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Budapest XIV. Ungvár u. 64-66.
1525 Budapest, P.O.Box 15.
Hungary, H-1525
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(361) 201-7471
Fax: (361) 251-9878
(361) 201-7471

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EDITORIAL

This special issue of the Journal on Communications is devoted to the living memory of Professor András Ambrózy. We decided to collect some of our new results demonstrating that the iniciative done by this outstanding Hungarian scientist resulted in new results in the field of electronic noise especially $1/f$ noise in different laboratories of the world. At the beginning of this issue Professor K. Tarnay gives on overview on Professor Ambrózy's activity.

In this issue the first paper deals with a very interesting problem: Low frequency noise measurements as a diagnostic tool in the semiconductor technology.

The next paper describes the application of $1/f$ noise measurements as a characterization and testing technique in solid-state devices.

The third paper considers a problem connected with the reliability prediction of digital integrated circuits.

The fourth contribution deals with the phenomena of conductance noise and percolation in high T_c superconductors.

The next paper gives a model for the influance on the noise of trimming cut-path in thick film resistors.

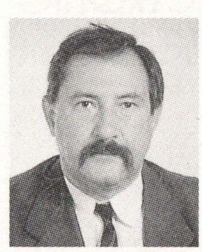
The sixth paper contains the results of the contact noise measurements of metal-GaAs multispot test patterns.

The seventh paper describes the $1/f$ phonon number fluctuations.

The last paper considers the problem of extraction of noise sources in BJTS and HBTS.

I do hope that this special issue will give an overview of contemporary results on $1/f$ noise in electronic systems and will stimulate new research activity and perpetuate Professor Ambrózy's name. It was a good feeling for acting as the guest editor of this special issue because a number of interesting papers were presented for publication. I would like to thank all the authors for the cooperation.

I. MOJZES



Imre Mojzes received MS degree in electrical engineering from the Moscow Power Institute in 1972. He received Doctor techn. degree (Ph.D.) from the Technical University of Budapest in 1979, Candidate of technical science degree (C.Sc) from the Hungarian Academy of Sciences in 1988. Presently he is working for the TU Budapest, as a professor and head of the Electronics Technology Department. His present research fields are metal contacts for GaAs devices, electronic noise and quality assurance problems.

IN MEMORIAM PROF. ANDRÁS AMBRÓZY

Prof. András Ambrózy an outstanding Hungarian scientist who initiated and carried out high level research activity in Hungary concerning noise theory of physical and electronic systems died five years ago on June 1, 1990. He liked music, and his hobby was sailing. The experienced and always careful sailor died while doing his favourite sport, as a result of the fatal combination of the unpredictable paths of nature and fate.

Electronics has aroused his interest very early, as a young teenager student of the secondary school. This interest — combined with his talent — remained continuous during his life.

His career was strongly associated with the Technical University of Budapest, he received his degrees at this institution: diploma in electrical engineering (1953), the postgraduate degree dr.techn. ("summa cum laude", 1962). The Hungarian Academy of Sciences awarded him in 1966 with the degree "Candidate of Technical Sciences", and in 1978 with the degree "Doctor of Technical Sciences". Since 1953 he was a lecturer at the Technical University of Budapest, first at the Department of Wireless Telecommunications, then at the Department of Electron Devices. He joined the Department of Electronic Technology as the head of the department in 1970. He became a full professor in 1979.

He led the educatioal and research work of the Department of Electronic Technology for 20 years. He also played a determining role in the Faculty of Electrical Engineering. He served and developed the Hungarian higher engineering education for many years as a deputy dean and as a member of the Faculty Senate and of the University Senate.

He published several books and more than 120 scientific papers. He was a world-wide known expert of the field of electronic noise. One of the most important of his works is the book "Electronic Noise", which was published several editions in Hungarian language [1], received many awards, and was also published in English by McGraw Hill [2] and in Japanese by Keigaku Publishing House [3]. He was a member of the international organizing committee of "Noise Phenomena in Physical Systems" conferences, and the editor of the Proceedings of the 1990 issue [4].

He participated and acted as a chairman in several committees of the Hungarian Academy of Sciences and the National Committee for Technical Development (OMFB). Since 1970 he was a member and since 1985 the chairman of the Electronics and Computer Science subcommittee of Committee of Scientific Qualifications (TMB).

He played an outstanding role in improving the niveau of the Hungarian hybrid circuit fabrication. For this he was awarded by the State Award of the Hungarian State.

He was one of the first Hungarian members of the international scientific society "Institute of Electrical and Electronics Engineers" (IEEE) and also became an "IEEE Fellow", which is the most prominent membership level. He took part in the activities and organization of "Hungarians for the Technical and Scientific Development of the World" conferences.

Since 1960 he was a member of the Scientific Society for Telecommunication and the chairman of its Award Committee and restlessly worked for the Hungarian electronics industry. He also had an important role in the League of Technical and Scientific Societies, mainly in the Educational and Award Committees.

The works initiated by the late Prof. Ambrózy have been continued during the past five years:

- the low frequency excess noise in epitaxial GaAs and InP layers has been studied;
- a new method for qualifying the contacts of high frequency devices by noise measurement has been developed;
- a method for realiability testing of CMOS integrated circuits by measurements of the supply current noise has been developed;
- the properties of polyimid layers are evaluated by noise measurements.

These research activities has been supported by Hungarian Research Fund (OTKA) projects, and the results are published in several papers, diploma and thesis works.

[1] Ambrózy A.: "Elektronikus zajok", Műszaki Könyvkiadó, Budapest, 1972. p. 283.

[2] A. Ambrózy: "Electronic Noise", McGraw-Hill, New York, 1982. p. 281.

[3] A. Ambrózy: "Denshi Noizu", Keigaku Publishing, Tokyo, 1988. p. 259.

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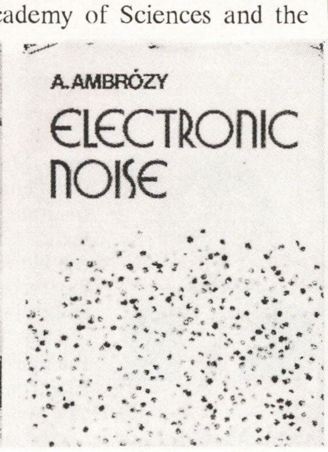
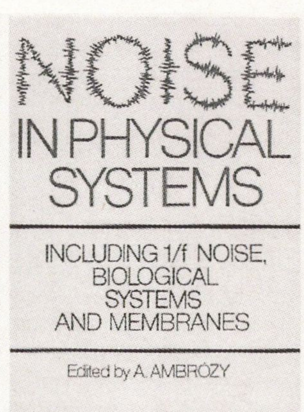
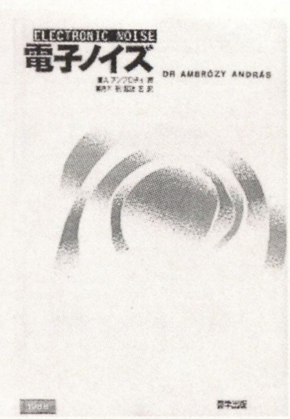
Prof. KÁLMÁN TARNAY







LOW-FREQUENCY NOISE MEASUREMENTS AS A DIAGNOSTIC TOOL IN THE SEMICONDUCTOR TECHNOLOGY

P. GOTTWALD

TECHNICAL UNIVERSITY OF BUDAPEST
FACULTY OF ELECTRICAL ENGINEERING
H-1521 BUDAPEST

We report on results of low-frequency noise measurements, which have been carried out on specially designed, strongly surface sensitive patterns prepared on GaAs and InP epitaxial layers. Generation-recombination (G-R) and $1/f$ noise components have been separated from the obtained noise spectra by curve fitting. Deep level analysis has also been carried out by observing the temperature dependence of the G-R noise components. In cases of unpassivated surface, which adsorbs contaminating particles from the atmosphere, the effects of commonly used cleaning solvents of the semiconductor technology were studied intensely. In these cases the shape and the time transient feature of the noise enhancement – occurring after the exposure of the contamination – have been analyzed. As it was observed the additional noise components do not correspond regularly to a distribution of discrete deep levels of a finite number. For the InP e.g. a quite pure $1/f$ character of the noise enhancement has generally been found. Next, different kinds of the passivation technologies and surface treatments have been applied to the samples, and the influence on the low-frequency noise spectra was analyzed in each case. We have stated, that the additional noise components to the $1/f$ noise, which are generated in the surface passivated samples, exhibit mostly G-R like characteristics. The nature of the deep level distribution was quite different for the GaAs than for the InP. An exact deep level structure was only detectable for InP. Finally, the lattice perfection of different GaAs epitaxial layers was investigated. In this case correlation between the $1/f$ noise component and the technology of the epitaxial crystal growth has clearly been observed.

1. INTRODUCTION

Because of their direct band structure and excellent electron transport properties, GaAs and InP have engaged an indispensable role in the high speed and high temperature electronics and in the optoelectronics as well. Combination of their related compounds by means of the advanced heteroepitaxial crystal-growth methods has opened the possibility of the bandgap engineering. On this basis a lot of electronic and optoelectronic devices, with extremely high operating speed have already been developed.

Nevertheless, there are more origins of crystal defects in the compound semiconductors, than e.g. in silicon. As a further contrast to silicon, they do not have a stable native oxide. So, it has to be replaced by other non-native dielectric, such as SiO_2 or Si_3N_4, for passivating layers and for mask applications. However there are many critical points of these technological steps, where considerable damage in the interfacial region may likely be caused.

Low-frequency noise measurements, carried out on suitable designed test patterns, have definitely been proved to be an effective inspection method for these problems. Therefore, the aim of this work is to give a review about this possibility.

2. THEORY

In the last decade, low-frequency fluctuation phenomenon in thin GaAs resistors has engaged much interest [1], [2], [3]. It is agreed upon, that the source is a resistance fluctuation, which can be converted to a voltage or current fluctuation by applying constant driving current or voltage to the resistor, respectively. Further, it is also evident, that both of the number- and mobility-fluctuation of the charge carriers in the resistors can cause resistance fluctuation.

2.1. Fundamental 1/f noise

Observations have shown, that the power spectrum of the low-frequency fluctuations always contains a $1/f$-type component, but more or less part of the power is attributed to any other type of spectra. However, the $1/f$-type component (thereafter only $1/f$ *noise*) seems to be a quite general phenomenon, which appears in the nature in almost all physical quantities. It was still be observed experimentally down to a frequency as low as 10^{-6} Hz [4].

There are particular models for the explanation of the origin of the $1/f$ noise [5], [6], [7], however a general theory giving insight undoubtedly into the fundamental physical reasons is missing till now.

Two of these models are based on the relaxation mechanisms which are always present in the metals and also in semiconductors. They can lead e.g. to number fluctuation of the charge carriers, which than consequently causes the resistance to fluctuate. In this case the necessary condition of $1/f$-type fluctuation is a $1/\tau$-type distribution of the relaxation times τ.

McWhorter has pointed out, that for semiconductors having an oxide (or dielectric) layer on the surface such a distribution of the relaxation times is realistic, if the trap distribution in the oxide is uniform [5].

The other model – the Dutta-Dimon-Horn (DDH) model [6] – is a quite popular one for metals. It is assumed in this model, that the physical background of the fluctuation is due to some thermally activated processes being present in the metal. It can be pointed out, that a constant distribution of the activation energies gives a $1/\tau$-type distribution of the relaxation times, thus $1/f$ noise arises.

Both of these theories fail for the bulk semiconductors. Namely, McWhorter's model is obviously surface oriented. Further the requirement of the DDH model is not exactly fulfilled too, since in semiconductors only the traps in the vicinity of a few kT to the Fermi energy level can give significant contribution to the noise.

The other group of models is based on the mobility fluctuation. They seem to be closer to the fundamental reasons. The most applicable theory for semiconductors is the quantum $1/f$ theory [8]. This points out, that the "*Bremsstrahlung*" induced by the scattering processes of any charged particle in a medium represents itself the basic physical phenomenon responsible for $1/f$ noise. This theory predicts $1/f$ spectrum down to $f = 0$ in limit, and supports the empirical relation stated by Hooge earlier [9]. However, modification was necessary later [10], because in certain cases the predicted noise levels were much smaller, than those of measurement results.

A quite different way to approach to $1/f$ noise – disregarding the physical reasons – is the empirical relation given by Hooge [9], as is follows:

$$S_R/R^2 = \alpha/fN. \qquad (1)$$

Here S_R stands for the spectral density of the resistance fluctuation at a frequency f, R is the resistance value and N is the total number of charge carriers in the resistor. The parameter α – called as *Hooge parameter* – was initially thought to be an universal constant with a value of about 2×10^{-3}. In contrast to that, values of α in a wide range of about 10^{-6} to 10^{-3} have been measured for various devices, depending on the device scale and material as well [11]. This fact is a subject of further investigations.

In spite of that, – in certain cases – important comparisons can be made using Eq. 1. For this, not too much differing materials should be compared on test samples of identical structural geometry, as we demonstrate it later. Such comparisons give information relating to the crystal lattice quality, since the fundamental $1/f$ noise has already been proved to be of mobility fluctuation origin.

2.2. Additional noise

The most important additional noise component is the generation – recombination noise (thereafter simply *G-R noise*). By its generation charge carriers will be trapped and released randomly by special energy states, causing thus typically number fluctuation. The most effective energy states are in the vicinity of the Fermi energy level, since the medium population allows the greatest fluctuation. Such allowed energy states are called as *deep levels*. Deep levels can be present in the bulk, but in the surface region as well. The physical origin of theirs is very different (defects, damages, complexes, strange atoms, adsorbed contamination, etc.). They usually represent themselves troubles in the structures, which possibly should be avoided.

G-R noise has a Lorenzian spectrum:

$$S(f) = S_0/[1 + (2\pi\tau f)^2]. \qquad (2)$$

Here $1/(2\pi\tau)$ is the cut off frequency of the spectrum, and the amplitude S_0 depends upon the density and other parameters (*activation energy, capture cross section*) of the deep levels. Both of τ and S_0 have a strong and characteristic temperature dependence. By plotting the temperature dependent values of $\lg(\tau T^2)$ versus $1000/T$ one can obtain the Arrhenius plot, which is normally linear. From the slope of this the activation energy and from its crossing value at $1000/T = 0$ the capture cross section can be determined.

Because of its known frequency dependence, G-R noise components can be separated by curve fitting from the measured total noise. Thus, deep levels can be analyzed and correlated with the parameters of the critical steps of the device technology, such as surface passivation, etching or cleaning.

3. EXPERIMENTAL

For the experimental investigations of the bulk and surface properties of the GaAs and InP, planar resistors have been prepared in both cases on the structure of a thin n-type epitaxial layer grown on semiinsulating substrate. The layout and the dimensions of the test structure are shown in Fig. 1. The different (*length/width*) relations serve for the appropriate resistance value at the different parameters of the epitaxial layer, such as thickness, doping and mobility. Each resistor has four contacts, making possible the four-point measurement, and the guard ring may be used to eliminate the surface leakage currents.

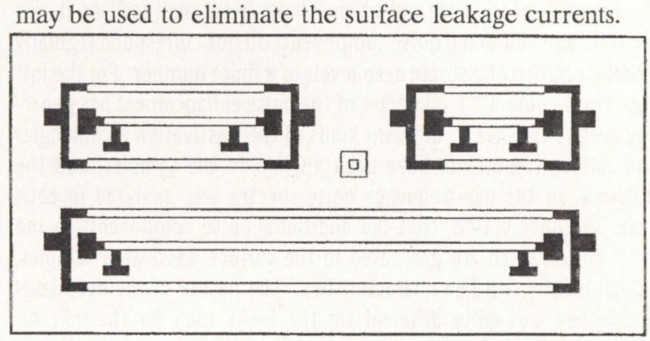


Fig. 1. Layout of the test structures used for the low-frequency noise analysis. (The line-width is 40 μm the length to width ratios are 10, 20 and 50.)

To realize this structure standard MeSFET technology has been applied, using mesa etching and lift-off technology for the AuGe-Ni ohmic contacts. In the case of the InP an n-type lattice matched InGaAs cap layer was additionally grown serving as a contact layer under the metallization. To get an uncovered, free InP surface, this cap layer was later removed from the resistor body by a selective wet chemical etching.

The samples have been subjected to different surface treatments or to different passivation technologies, and low-frequency noise measurements have been carefully performed in a frequency range of 1.6 Hz to 20 kHz. The DC driving current was matched to about 0.6 mA, if the sample had a resistance at about 2 kOhm. Four-point measurement technique was used, thus contact noise has been omitted. The sample temperature was regulated.

The obtained spectra were systematically analyzed in the following manners:

- While investigating the surface treatments, the time transients of the noise spectra were monitored and compared for different contaminating materials.

- For comparison of passivation technologies the noise spectra were split by curve-fitting to a $1/f$ spectrum, and to additional G-R noise components. In this case the temperature dependence of the G-R noise components has been analyzed.
- To compare the crystal perfection, good passivated samples should be measured. After extracting the isolated G-R noise components from the total noise, the remaining nearly $1/f$-type component can characterize the crystal perfection.

4. RESULTS AND DISCUSSIONS

4.1. Contamination on the free surface

First, surface effects on the low-frequency noise of thin GaAs and InP layers will be demonstrated [12], [13]. In these experiments the investigated resistors had an unpassivated, free surface. Our investigations have been focused on the contaminating effect of the wet cleaning materials normally used in the standard cleaning process in the semiconductor technology, such as acetone, methanol, chloroform and water. They were brought onto the semiconductor surface in form of saturated vapour in a N_2 atmosphere. The noise spectra and the DC resistance values were monitored during the course of the treatments.

To have comparable results for the different materials, the same resistor chip was used for each experiment. Therefore an effective method had to be found, to clean and regenerate the surface in any case before starting with a new material. Various attempts using heat treatment in inert gas or vacuum had failed. However, excellent results have been achieved by illuminating the warm surface by ultraviolet light in inert gas-flow. (The surface temperature and the light power density were 373 °K and at about 20 mW/cm^2 respectively.) Furthermore, this method is successfully applicable for both of the GaAs and InP.

In Fig. 2. three noise spectra demonstrate the effectiveness of the method for GaAs, if the surface regeneration is consequently performed after each contaminating cycle. It can be seen, that even a series of treatments does not alter significantly the noise spectrum (curve 1 and 2), although the noise spectrum shows a large enhancement e.g. after 1 hour methanol treatment (curve 3). Otherwise, curve 4 illustrates that the system noise is small.

According to our observations, there is a long time transient (roughly 20 hours) of the noise spectra accompanied by tendentious changes of the resistance values as well, for both of the GaAs and InP. The time transients were characteristic for the contaminating materials and were also different for the GaAs and InP. For InP the shape of the noise enhancement was $1/f$-type for all of the used contaminating materials, and had a scope of about 40, 25, 6 and 2 dB for water, methanol, acetone and chloroform, respectively.

For GaAs a nearly $1/f$-type noise enhancement was only observed for the water and acetone, while for the chloroform and even for methanol a much greater enhancement was observed in the lower frequency part of the spectrum. The range of the enhancement was at 20 Hz at about 19, 18, 8 and 4 dB for water, methanol, acetone and chloroform, respectively.

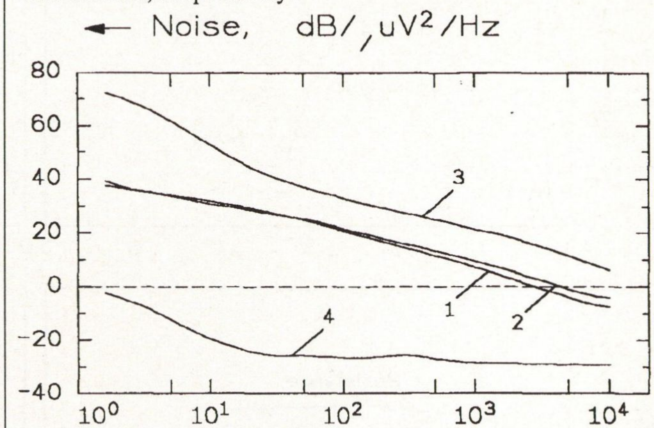


Fig. 2. Noise spectra measured on GaAs structure with unpassivated free surface. 1) before, 2) after a series of contaminating investigations followed by surface regeneration, 3) after 1 hour methanol treatment. Spectrum 4) corresponds to the measuring system noise.

The resistance values were essentially growing, with a tendency of saturation. The changes were greater for water and methanol and much smaller for acetone and chloroform.

Figs. 3. and 4. illustrate the above statements with the detailed measurement results.

To explain the phenomenon a modification of the surface state structure of the semiconductor can be supposed, while contaminating particles interact with the surface, and will lastly be adsorbed on it. The adsorbed particles create new surface states, and a new balance builds up between the charges on the surface and in the neighbouring depletion layer. The monotone increase of the sample resistance indicates, that the width of the depletion layer grows. This corresponds to an increase of the negative surface state charge, occurring through the electron-acceptance of the adsorbed particles. The fact that the removing of the contamination can be promoted by ultraviolet light, supports the assumption of the electron transfer.

The origin of the excess noise is the random fluctuating motion of the charged particles on the surface. As these fluctuations should result in equivalent fluctuations of the depletion layer width, which modulate the thin conducting channel, the effect can be more pronounced as the average increase of the resistance. Further, the noise intensity is strongly dependent on the surface coverage. At high coverage the fluctuation is restricted, diminishing the noise (see e.g. water contamination on InP).

Further details and explanation of the phenomenon can be found in the literature [12]. As a conclusion, it should be stated, that the wet cleaning materials of the semiconductor technology may strongly be adsorbed on the surface, and require special treatment to desorb them. Low-frequency noise measurements have been proved to be a sensitive tool to investigate these phenomena.

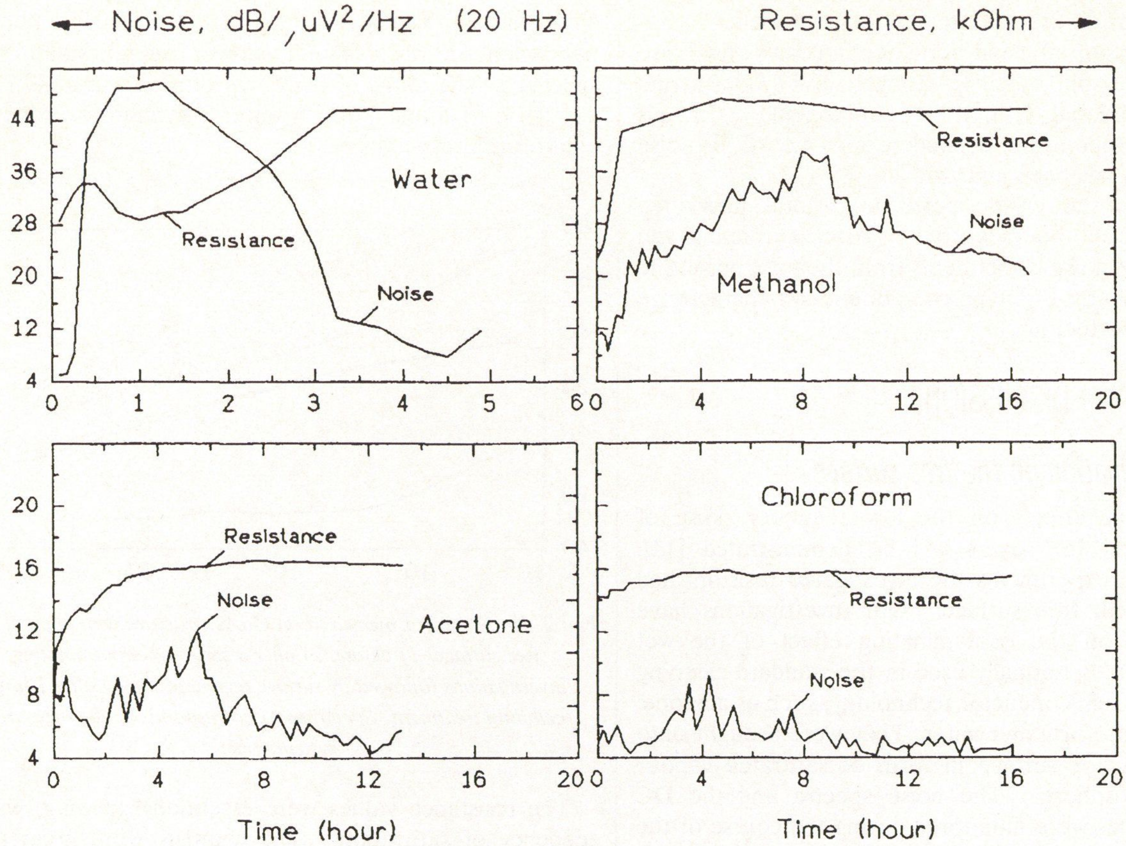


Fig. 3. Time transients of the DC sample resistance and of the noise spectrum level at 20 Hz, measured on an unpassivated InP structure in the course of different contaminating treatments.

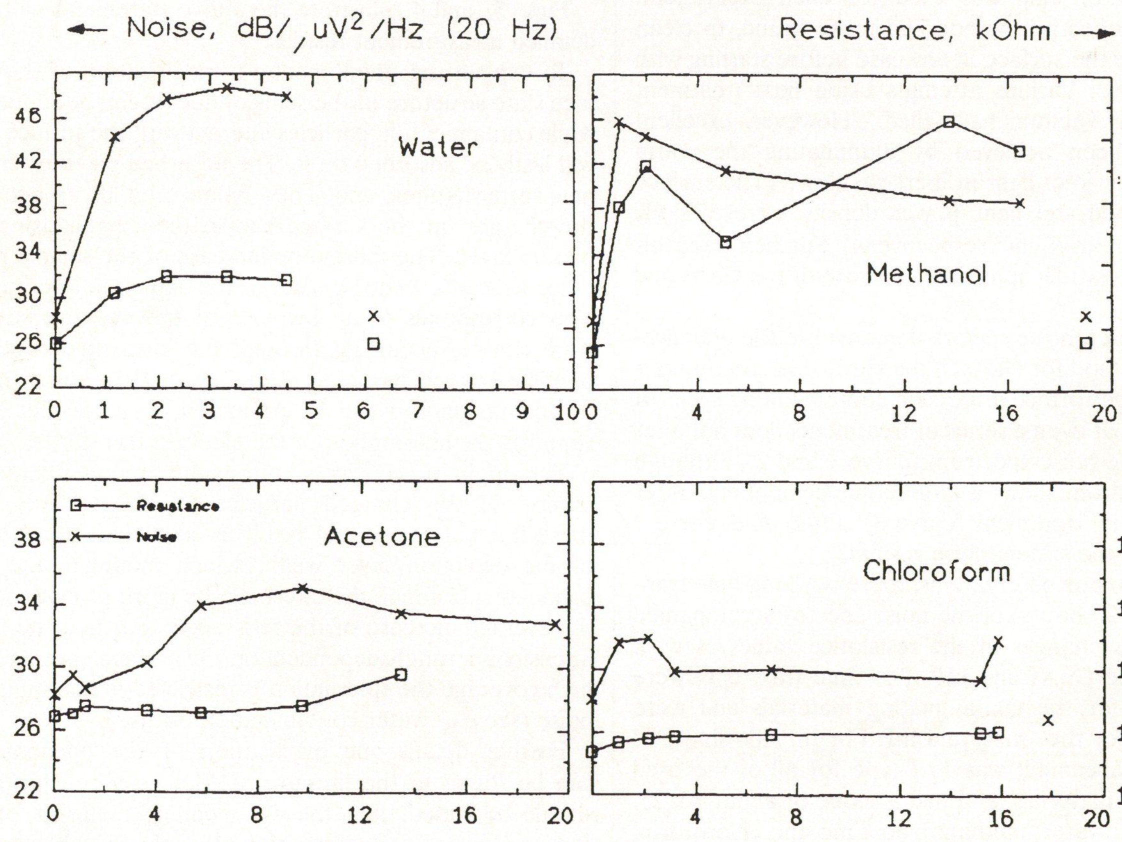


Fig. 4. Time transients of the DC sample resistance and of the noise spectrum level at 20 Hz, measured on an unpassivated GaAs structure in the course of different contaminating treatments.

4.2. Comparison of passivating processes

As the previous section has shown the unpassivated, native surface of the GaAs and InP is an unsatisfactory boundary for the high quality and stable semiconductor devices. Therefore, a good passivation is extremely important, even for very small, surface dominated devices. The best results are obtained by the deposition of SiO_2 or Si_3N_4 layers, but the deposition technology should carefully be optimized. To avoid any out-diffusion of As or P and subsequent degradation of the surface the deposition at low-temperature is essential.

Plasma-enhanced chemical vapour deposition (PECVD) has been widely applied since uniform film deposition at temperatures in the range of 200–400°C is possible. However, high energetic ions present in the plasma, impinge on the surface, and may enter up to an extent of 200 nm. [14]. Alternatively, much interest has emerged in photo CVD (in our case: Hg sensitized) as a low-damage deposition process at substrate temperatures of about 100–200°C. Still these films have a lower density and are chemically less stable than the plasma-grown films due to the incorporation of various intermediate reaction products. Further, photo CVD passivation gives basically different result for the InP than for GaAs. Recently, it has been shown, that a modified plasma process for SiO_2 deposition in a phosphorous ambient [15], which results in phosphosilicate glass (PSG) with a small amount of phosphorous ($<$ 2 wt%), considerably improves the film and interface properties. (In the modified process an electrostatic shield is used to diminish the energy of the bombarding ions.)

According to the theory, the interface and oxide properties (degree of damage and concentration of traps of different physical origin) strongly influence the contribution of the additional G-R noise components to the $1/f$ noise in the low-frequency spectrum. Therefore, low-frequency noise measurements have been carried out on the surface sensitive resistor chips described earlier. The measurements have been performed at different temperatures in the range of 0–80°C. $1/f$ noise and G-R noise components were separated and analyzed by computer-aided methods.

In this section characteristic results will be reviewed for different passivation technologies for GaAs and InP. All of the details of the deposition parameters have been published in [14], [15].

4.2.1. Results for GaAs passivated by SiO_2

As an important feature, we first investigated the reproducibility. By comparison of the noise spectra of several chips at different temperatures we have stated, that the PSG passivation of GaAs was very well reproducible. Considerably worth reproducibility has been found for PECVD passivation, and much worth was the reproducibility for the photo-CVD passivated chips. Further, in the case of photo-CVD, qualitatively differing spectra have been found, which show a strong temperature dependence in a specific temperature range. In most cases, a strong enhancement of the noise power density was detected around room temperature (300 K), and at lower frequencies (mostly below 20 Hz) as it is shown in Fig. 5. For the supposed explanation of these observations by photo-CVD we refer to the literature [16].

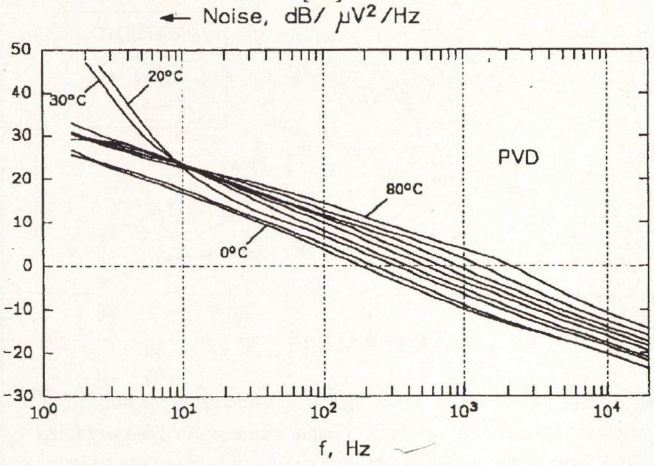


Fig. 5. Typical enhancement of low-frequency noise between 10 and 30° C of photo-CVD passivated samples.

Furthermore, we tried to analyze the deep level distribution for each of the technologies. The original spectra were quite well approximated by the sum of a $1/f$ noise and two different G-R noise components. However the Arrhenius plots, that have been constructed from the temperature dependence of the G-R noise components, have a strongly disperse nature, and they do not yield two discrete trap levels. This may be an indication of a more complicated and spread trap level distribution.

Nevertheless, a good characterization of the different passivation processes has been achieved by plotting the temperature-dependent Noise-Power-Ratio (NPR), that is defined as:

$$NPR = \frac{total\ additional\ noise\ power}{power\ of\ 1/\mathrm{f}\ noise}, \quad (3)$$

where the power is integrated in an arbitrary bandwidth of 2 Hz – 2 kHz.

In Fig. 6. curves of NPR are plotted for two photo-CVD passivated structures, showing considerable deviation of the peak heights. This underlines the strong temperature dependence of the noise power density at lower temperature, regularly at about 300 K, and illustrate the bad reproducibility of the passivation as well. Both facts are connected with the ill-defined nature of the interfacial region by photo-CVD, as discussed in the literature [16].

The temperature dependence of NPR is illustrated in Fig. 7. for PECVD of SiO_2 and for the deposition of PSG. It can be seen, that the noise enhancement is obtained in a wider temperature range, and the maximum is generally lower, than it was for the photo-CVD process. Further, the result for PSG deposition is better than for the PECVD of SiO_2. In comparison to the photo-CVD the basically different curves are the consequence of the essentially other type of the interface layer due to the effect of the high-energetic ions present in the plasma, and due to the higher substrate temperature. Though, we want only to focus now on the illustration of the differences between the technologies, for a detailed theoretical explanation we refer to the literature [16].

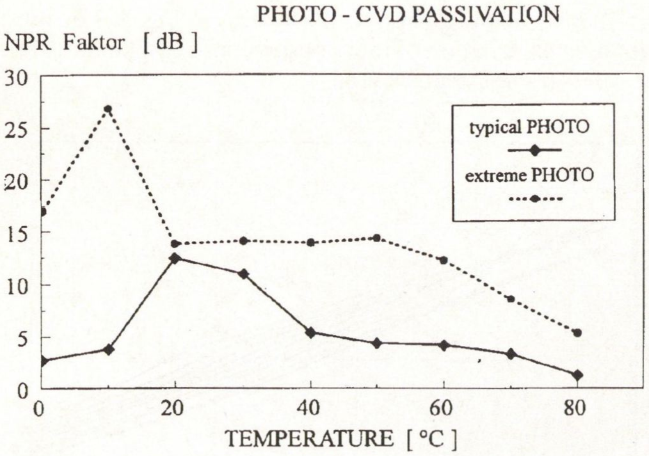


Fig. 6. Plots of the noise factor NPR vs. temperature for photo-CVD passivated GaAs resistor chips. Typical values have been obtained from most of the analyzed chips. Extreme values are not unique, they did occur on the same wafer.

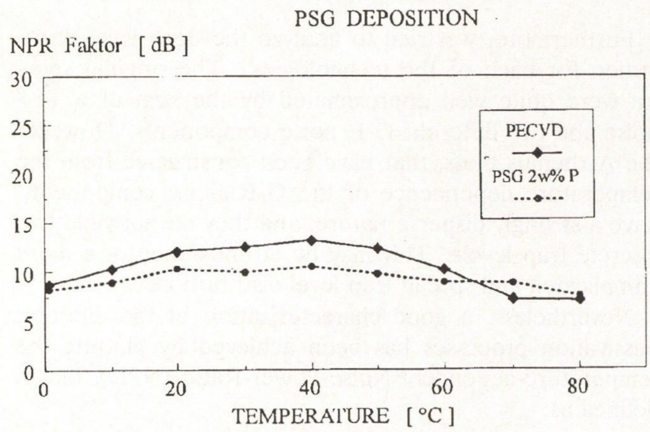


Fig. 7. Typical plots of the NPR vs. temperature for PECVD passivation. Conventional PECVD and modified PECVD in phosphorous ambient (PSG) are compared for GaAs.

As a further proof of our method the remaining $1/f$ components were also analyzed, and the Hooge parameters were calculated. At room temperature the values of the Hooge parameter were at about 10^{-3} with a temperature coefficient around 3 %/K. These results are in a good agreement with those of the literature [17].

4.2.2. Results for InP passivated by Hg sensitized photo-CVD and PECVD of SiO_2.

Practically, the photo-CVD do not cause surface damage. However, incorporation of HgO in the oxide, – especially at lower deposition temperature, – may cause traps close to the surface and an additional noise can be detected. At higher deposition temperature the HgO incorporation diminishes, but the out-diffusion of phosphorous increases.

In our case the deposition temperatures were 150 and 300 °C. At the lower temperature extremely low additional noise was detected, which did not allow reliable deep level analysis. (The temperature dependence of the noise spectra was well below 1 dB/80°C.) However, the additional noise was somewhat greater for the samples passivated at 300°C, and a deep level (at 0.35 eV), that is characteristic for the phosphorous loss, have nearly been detected (at 0.32 eV) by means of the Arrhenius plot.

The PECVD of SiO_2 was carried out at a substrate temperature of 300°C at an RF power of about 50 W, and we did not use electrostatic shield against the ion bombardment (i.e. the process was not "modified"). As a consequence a damaged interface was created under the surface, hence a considerably increased additional noise has been measured. In this case the temperature dependence of the noise spectra was already at about 10 dB/80°C, as it can be seen on a typical example of a temperature dependent noise spectra in Fig. 8. Further, in Fig. 9. results for five samples have been analyzed by Arrhenius plots. This shows, that two deep levels at nearly the same energies (at 0.53 and 0.46 eV) with a capture cross section in the range of 10^{-15} cm^2 can be detected. We note, that deep levels around 0.5 eV are commonly known values appearing in the literature.

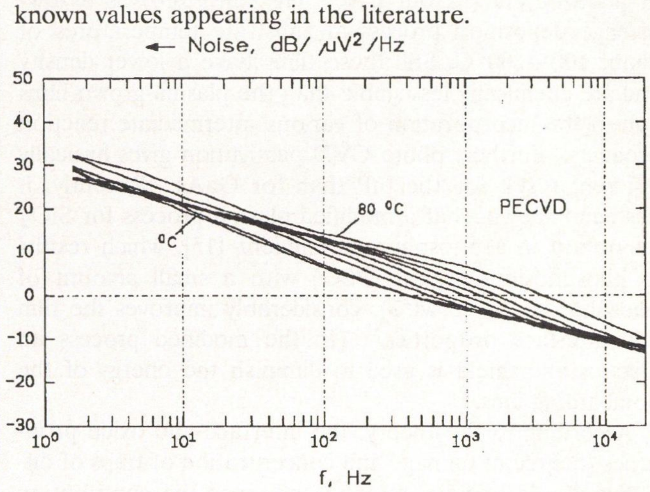


Fig. 8. Typical temperature-dependence of the low-frequency noise spectrum for InP in the temperature range of 0–80° C. The sample was passivated by conventional PECVD technology.

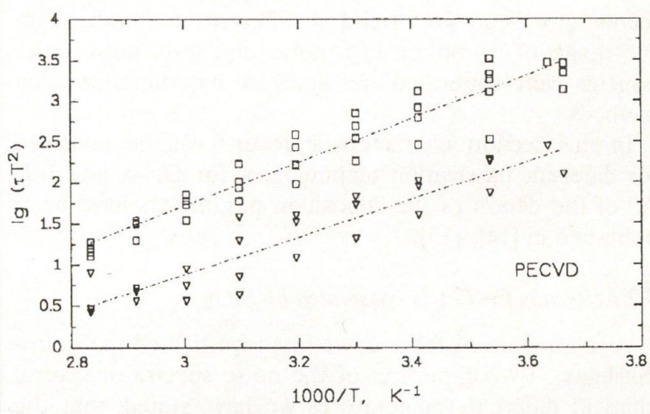


Fig. 9. Arrhenius plot constructed from the results of five InP chips. Conventional PECVD technology was used. Two dominant G-R noise components were investigated.

In conclusion it can be seen, that the passivation processes can also sensitively be analyzed both for the GaAs and InP by means of low-frequency measurement technique.

4.3. Low-frequency noise measurement as an indication tool for the crystal lattice quality

To demonstrate this ability five different GaAs layers will be investigated. The samples were passivated by CVD SiO_2 layer, deposited from silane at 400°C. The other parameters are summarized in Table 1.

Table 1. Parameters of the investigated GaAs layers

No.	Technology		Thickness (nm)	Doping (cm^{-3})	Mobility (cm^2/Vs)
1	MBE	*	1000	1.8E 14	7000
2	MOCVD	#	300	1E 17	3900
3	VPE	+	300	1E 17	3700
4	LPE	+	300	1E 17	3700
5	VPE	•	300	1.1E 17	3600

(* – Tampere University of Technology; # – Inst. of Semiconductors, Kiev, Acad. of Sciences; + – Research Inst. for Technical Physics of H.A.S, Budapest; • – TESLA, VUST, Prga)

As already mentioned, the fundamental $1/f$ noise should now be extracted and analyzed on the basis of Eq. 1. However, Eq. 1. should somewhat be modified for semiconductors. Namely, the impurity scattering does not contribute to the fundamental $1/f$ noise [18], but influences the resistance value. Therefore, values of α which are formally calculated from Eq. 1. should be distinguished as α_{exp}. To describe the fundamental $1/f$ noise the corrected value α_l should be used, that is:

$$\alpha_l = (\mu_l/\mu_{exp})^2 \alpha_{exp}, \tag{4}$$

where μ_{exp} is the experimental measurable mobility, and μ_l stands for the lattice mobility. At low temperature this correction is essential, but at higher temperature the mobility will be even more lattice dominated. Thus, at room temperature for approximate calculations $\alpha_l \approx \alpha_{exp}$ can already be accepted.

First, the noise spectra were measured, and the G-R noise components were separated and extracted. Thus, the $1/f$ noise components yield.

In the concrete case noise voltage spectra were measured at a known DC voltage across the samples. Thus $S_R/R^2 = S_U/U^2$. Further, the known geometry and scales, the measured mobility and the sample resistance determine N, that is the number of charge carriers. Using Eq. 1. and the approximation of $\alpha_l \approx \alpha_{exp}$ for room temperature the following values have been found for α_l:

MBE	MOCVD	VPE +	LPE	VPE *
1.3×10^{-4}	5.2×10^{-4}	1.1×10^{-3}	1.5×10^{-3}	3×10^{-3}

Though, the lower value indicates the higher degree of crystal lattice perfection, the above order may well agree with the experiences.

In conclusion, we have demonstrated the applicability of the low-frequency noise measurements for the characterization of the crystal quality.

5. CONCLUSIONS

The measurement and analysis of low-frequency noise generated in thin GaAs and InP layers have been proved to be a powerful technique for monitoring technological processes. Using this method a sensitive analysis of the contaminating effects on unpassivated semiconductor surfaces has been carried out. An effective surface cleaning method has been found, that is able to remove adsorbed particles of cleaning solvents widely used in the semiconductor technology. Different passivation technologies of the GaAs and InP were successfully investigated and qualified by low-frequency noise measurements giving insight into the deep level structure of the interface layers as well. Finally, the crystal perfection was qualified in case of GaAs for various crystal growth technologies.

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Péter Gottwald received the Dipl. Eng. degree in Electrical Engineering from the Technical University of Budapest in 1966. He joined the Department of Electronic Devices of Budapest Technical University in 1966, where his research and teaching activities were focused on the theory and operation of the active electronic devices. He was a Visiting Scientist at the Technical University of Aachen (RWTH-Halbleitertechnik) in 1978 and at the University of Duisburg in 1986 and 1990. Since 1991 he is co-operating with the Technical University of Darmstadt and the German TELEKOM. Dr. Gottwald has been a Consultant to the Research Institute for Technical Physics of the Hungarian Academy of Sciences for many years. His research fields include microwave semiconductor devices and technological problems of the inter-metallic semiconductor materials. Since 1990 most of his work has been focused on the noise-phenomena of GaAs and InP. Dr. Gottwald has been awarded by the Best Paper Award "Pollák-Virág" of the Scientific Society for Telecommunication in Hungary in 1990.

1/f NOISE MEASUREMENTS AS A CHARACTERIZATION AND TESTING TECHNIQUE IN SOLID-STATE DEVICES

Z. ÇELIK-BUTLER, A. JAHANZEB, W. HE and R. ZHANG

SOUTHERN METHODIST UNIVERSITY
ELECTRICAL ENGINEERING DEPARTMENT
P.O. BOX 750338, DALLAS, TX 75275-0338

An overview of $1/f$ noise measurement techniques is done where a brief description of experimental techniques and results is given about methodologies developed for characterization and testing in microelectronic systems. Three such evaluation techniques are discussed as examples. The first one involves the characterization of electromigration parameters and prediction of the life-times in Very-Large-Scale-Integrated (VLSI) circuit metallization layers and vias. By performing $1/f$ noise measurements in accelerated electromigration conditions such as elevated temperatures and at stressing current densities, one can evaluate the electromigration parameters like activation energy and current exponent. Then, these parameters are used to establish a correlation between the life-time of the metallization layer and the $1/f$ noise the film exhibits, which in turn can be used as a predictor of life-time. Specific examples are given in relation to Al-Cu-Si/TiW/Al-Cu-Si multilayered via electromigration. Secondly, $1/f$ measurements can be utilized to characterize thin films and to determine the quality of the film related to its composition, crystallinity, and electrical characteristics. Specific examples are given on $YBa_2Cu_3O_7$ thin films in relation to the superconductive behavior, transition temperature and crystal orientation. The third, and perhaps the most widely reported application of the $1/f$ noise characterization techniques is the measurement of semiconductor-insulator interface state densities. This can be done by varying the bias conditions of the structure, temperature or both. In this paper, HgCdTe-ZnS interface state characterization will be discussed.

1. INTRODUCTION

Inherent fluctuations in electronic devices, especially $1/f$ noise, have generally been seen as a nuisance by engineers and scientists; as a phenomenon that should be avoided until its presence makes it absolutely necessary to deal with. To a very few of us who actually like doing research on fluctuation phenomena, noise is an outcome of a complex series of physical mechanisms that change with changing bias conditions, temperature, device structure and material. Not only understanding and minimizing the fluctuations, but also making some use out of these is the goal. In that context, researchers have begun to use noise measurements as a characterization and testing tool in microelectronics.

One of the first examples for utilization of electronic noise as a characterization tool has been in electromigration studies in thin metal films [1]. Later similar methods have been developed for other systems and devices. Although no single "Noise Technique" exits most of the applications of noise measurements as a detection and testing method have been by utilizing Low-Frequency Noise (LFN), and mostly $1/f$ noise. All such methods involve noise measurements at the lower ranges of the frequency spectrum either as a function of temperature or bias conditions to extract parameters that are inherent to the mechanism(s) proven to be the origin of these fluctuations [2]. Therefore, first there should be profound evidence to the fact that the characterized mechanism indeed causes these low- frequency fluctuations. Second, the proposed LFN method should be easy to implement and should surpass other existing techniques in at least one aspect. The maturity of LFN technique varies from technology to technology. In silicon based VLSI circuits and components, the method is more developed and more widely used whereas for superconductivity it is still at its infancy.

2. DETECTION OF ELECTROMIGRATION

Electromigration is defined as mass transport under the action of an electric field. Electromigration causes shorts and openings in VLSI metal interconnections through the development of hillocks and voids. As the dimensions continue to scale down, the high electric fields introduced in the metal films, and elevated temperatures due to current crowding effects worsen the effects of electromigration causing it to be the most dominant mode of failure in metal interconnections. Therefore, it is of paramount importance that electromigration mechanisms are characterized for each aluminium alloy used in the metallization lines and for the specific geometry. Prediction of life-times of these metallization layers is essential for reliability purposes. More often prediction of life-times is done by performing so-called "Sweat" tests where test structures are stressed with high electric fields at accelerated conditions and sometimes at elevated ambient temperatures until a predetermined failure criterion is met (such as 10% resistance increase). These tests are time-consuming and destructive. Moreover, there is a question to whether the electromigration mechanisms characterized at the accelerated conditions correspond to those at normal operating conditions. It is also known that the measured life-time and electromigration activation energy is a function of the failure criterion. A simple alternative is currently being researched at least four laboratories [3] – [8] where a low-frequency noise technique is used to predict the metallization life-times. In this paper, the discussion will be limited to the results obtained in the authors' laboratory.

According to Black, the mean-time-to-failure (MTF) can be expressed in terms of the current density J, temperature T, activation energy E_a, and Boltzmann's constant k as [9]:

$$MTF = A\frac{J^{-n}}{kT}e^{(E_a/kT)}. \quad (1)$$

Here, A is a material and geometry dependent constant and n is the current exponent measured to be around 2 in most samples. There is growing evidence that at stressing current densities, the low-frequency noise measured on thin metal films is inversely proportional to MTF. At stressing current densities, the voltage noise power spectral density S_v for low-frequency noise has the following form [4], [7]:

$$S_v = B\frac{J^n}{Tf^\gamma}e^{(-E_a/kT)}. \quad (2)$$

Here, J, n, E_a, k, and T have the same meaning as in (1), f is the frequency, γ is the frequency exponent and B is another constant that depends on the material, geometry and microstructure of the film. One should note that expression (2) represents the noise under stressing conditions where vacancy migration dominates all other mechanisms as the origin of $1/f$ noise, such as local temperature fluctuations. Eq. (2) was derived for $n = 1$ and $\gamma = 1$ by Neri *et. al* [10] who based their model on the fluctuations in the number of migrating vacancies. How these fluctuations lead to electronic fluctuations were left unclear, however. A more general form was derived by Yang and Çelik-Butler [6]. They based their model on electron mobility fluctuations due to the fluctuations of the number of vacancies and gave an equation similar to (2) with $n = 2$ and $\gamma \geq 1$. This equation agrees well with the experimental results since both $1/f$ and $1/f^2$ noise are observed. Moreover predictions of the model on the temperature and current density dependence of the noise magnitude and the frequency exponent were experimentally verified for single layer,1% Si Aluminum thin films [6].

The application of noise measurement technique to reliability in metallization layers has not been without controversy. In single layer Al-based alloys, the electromigration activation energies measured through both noise and sweat techniques were found to agree closely with values being in the range of 0.6 eV which represents grain boundary vacancy motion. This is reasonable considering that the microstucture of these thin-films are dominated by grains. Such an agreement implies that the mechanism leading to the low-frequency fluctuations is the same as or closely related to grain boundary vacancy motion which is known to be the dominant electromigration mechanism in these interconnects. In metallization layers where there is an underlying refractory metal, such as TiW, a discrepancy in the measured activation energies has been found between MTF and noise techniques. The sweat techniques reveal $E_a \approx 0.6 - 0.7$ eV even in the presence of the underlying refractory metal. The E_a measured through noise technique, on the other hand, was found to be considerably lower, in the range of 0.1 to 0.2 eV. Table 1 summarizes the results.

It seems that the underlying TiW layer not only changes the grain size and structure of the film (which is well known) but it also somehow makes the contribution of the noise component arising from the oxide-free surface electromigration to dominate over the other components. The observed activation energy values of 0.1 to 0.2 eV are closer to oxide-free surface vacancy motion which is computed to be in the vicinity of 0.28 eV for Aluminium alloy thin films [16]. It is interesting to note, however, that even with this discrepancy, researchers have found the noise to be a good indicator of reliability in multilayered metal strips and via structures. The $1/f$ noise magnitude was shown to be inversely proportional to MTF for TiW/W/2% Cu-Al thin films [5]. Average voltage noise spectral density S_v at 10 Hz vs. MTF is depicted in Fig. 1 for 5 wafers. Each point represents an average of noise data taken from at least five strips on one wafer. The least squares fitted line shows an inverse relationship between parameters with $S_v \propto (MTF)^{-1.1}$.

Table 1. Comparison of activation energies

Method	via (Al-Si-Cu/TiW)	strip TiW/Al	strip Al single layer
LFN	0.11&0.23 eV [11]	0.14 eV [5]	0.6-0.7 eV [6], [10]
MTF	0.48 eV [12]	0.35-0.7 eV [12]-[15]	0.6-0.7 eV [7]

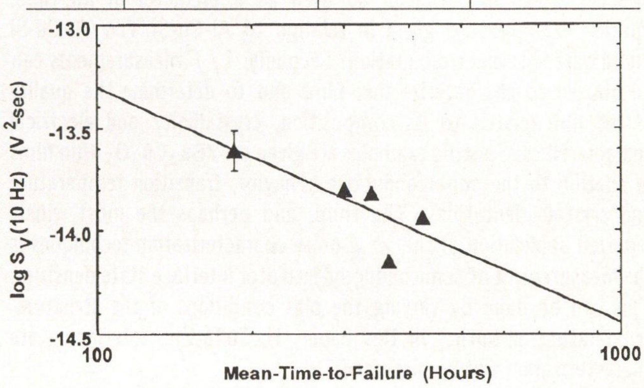

Fig. 1. TiW/W/2% Cu-Al thin films: Voltage noise power spectral density at 10 Hz vs. Mean-Time-to-Failure for five wafers. Noise measurements were performed at $T = 150$ °C with the samples biased with $J = 4 \times 10^6$ A/cm^2. Life-time tests were done at $T = 200$ °C ($T_{stripe} = 216$ °C) and $J = 2 \times 10^6$ A/cm^2. The least-squares fitted line shows an inverse relationship between these parameters.

The current technology is going towards smaller dimensions and therefore higher aspect ratios in vias, which underscores the issue of via electromigration reliability. The next challenge in utilization of low-frequency noise in metallization reliability lies in via structures.

3. SUPERCONDUCTING THIN FILMS

Since the discovery of high-temperature superconductivity, applications of superconductive materials to medical, defense, and consumer electronics have increased due to the lower cost of operation made possible by the higher operation temperature, and ease of fabrication. The inherent low-frequency noise in high-T_c superconductors affects the performance of the devices and therefore deserves careful analysis and characterization. It can also be used as an indicator of material quality, in terms of the degree of crystallinity, crystal orientation, superconducting transition temperature, and transition width. Some preliminary investigations have been reported relating the $1/f$ noise magnitude to the crystal structure and orientation. R. D. Black et. al [17] showed that $1/f$ noise magnitude measured on c-axis-normal epitaxial $YBa_2Cu_3O_{7-\delta}$ films on $SrTiO_3$ substrate was two to three orders of magnitude lower than that measured on granular $YBa_2Cu_3O_{7-\delta}$ films on ZrO_2 buffer layers with different substrates. It was also found that as the R-T characteristics deteriorated, the $1/f$ noise component increased in these films [18], [19]. The $1/f$ noise magnitude increased with time as the film degraded when it was left exposed to air for a prolonged period of time.

Currently, perhaps the biggest hurdle in utilization of noise measurements as a reliability tool in superconductivity is the lack of method for normalization of the observed voltage noise by the fluctuation quantity. Without such normalization, comparison between noise magnitudes observed on samples of different size and composition becomes impossible. Hooge's expression of $1/f$ noise voltage noise spectral density S_V is given by [20]:

$$\frac{S_V}{V^2} = \frac{\alpha_H}{N}\frac{1}{f^\gamma}. \tag{3}$$

Here, α_H is Hooge's parameter commonly used as a normalized noise parameter for comparison of different samples, $\gamma \approx 1$ and $N = nV_S$ where n is the charge carrier concentration and V_S is the sample volume. In order to perform a proper comparison between different samples, the quantity α_H has to be computed from the measured noise spectral density S_V. Although for metals and semiconductors this is trivial, it turns out to be a major task for superconductors in superconducting transition, since the number of quasi-particles, n, is a strong function of temperature. For rigorous calculations, this quantity has to be directly measured on the sample simultaneously with the noise measurements. Hall-Effect measurements have been used [21] to obtain Hall carrier density which in turn was used as charge carrier concentration to normalize the measured noise in Eq. (3). Fig. 2. shows the voltage-normalized noise special density (a), and the voltage- and Hall-carrier-normalized noise spectral density (b) measured on a c-axis normal $YBa_2Cu_3O_{7-\delta}$ film deposited on a $LaAl_2O_3$ substrate. The critical temperature of the film at zero applied magnetic field is 82.3 K. It can be clearly seen that the increased noise at transition in Fig. 2(a) with the increasing magnetic field is actually an anomaly caused by the lack of proper normalization that is corrected in Fig. 2(b). When the noise is properly normalized by the Hall carrier density, such an increase due to magnetic field is not observed. In fact, the normalized noise is not sensitive to applied magnetic field at the superconducting transition. This underlines the importance of normalization with respect to the fluctuating quantity instead of merely with respect to volume.

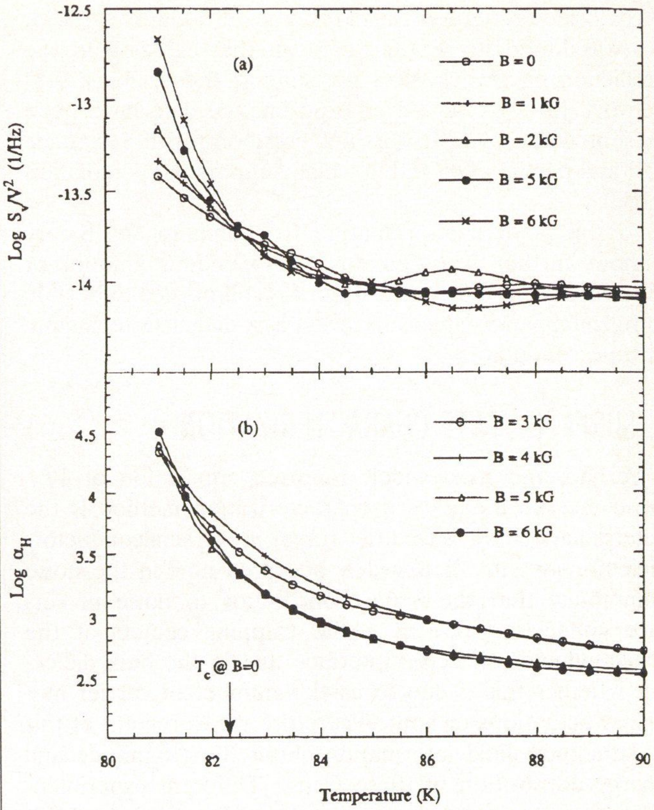

Fig. 2. $YBa_2Cu_3O_{7-\delta}$ thin film on $LaAl_2O_3$: Normalized voltage noise power spectral density at 5 Hz vs. temperature for different values of applied magnetic field, B. In (a) the normalization was done with respect to only applied voltage. In (b) the normalization was done with respect to voltage and Hall-carrier density.

Another problem that the researchers are facing in using noise measurements as a characterization tool for high-T_c superconductors is the lack of a universally accepted model that accounts for the fluctuations in the superconducting transition. Although several mechanisms have been proposed, there are only three that survived the experimental tests: flux hopping, equilibrium temperature fluctuations, and percolation models. The first is due to the motion of a flux trapped between the non-superconducting regions of the film causing fluctuations in the observed voltage in the superconducting transition. Telegraph-like noise observed in some samples is usually shown as an indication of this mechanism [22], [23]. The second, temperature fluctuations theory, has had mixed support from the experimental community. Although some investigations [24], [25] indicated a β^2 dependence for the normalized voltage noise spectral density, taken to be a proof for the validity of the temperature fluctuations theory, others failed to establish such a relationship [21], [26]. Here β is the temperature coefficient of resistance: $\beta = (1/R)(dR/dT)$. Percolation theory, on the other hand, seems to be very promising. If the voltage noise spectral density measured at superconducting transition is properly normalized, i.e., with respect to voltage and carrier density, then the classic percolation model is confirmed, according to which the normalized noise is proportional to $R^{-k'/s}$, where R is the macroscopic sample resistance and k' and s are the critical exponents. For a c-axis oriented film on $LaAl_2O_3$, the zero-Gauss value of k'/s was found to be 1.06, a result that agrees with the prediction of the classical percolation theory for a 2-D network [21]. A wealth of experimental data have been presented also by L. B. Kiss and collaborators in reference [27] and papers cited within which support the percolation model.

At this point it is premature to submit to any theory without further investigations. Tremendous amount of experimental and theoretical work is still needed to be able to fully apply noise measurements as a diagnosis technique in superconductors.

4. SURFACE-STATE CHARACTERIZATION

Perhaps the most widely reported application of $1/f$ noise measurements as a characterization method is the determination of dielectric traps in a semiconductor-dielectric system. It is widely accepted now in the noise community that the major contributor to noise in surface conduction devices is the trapping centers at the semiconductor-dielectric interface and in the bulk dielectric whether this is due to carrier number or carrier mobility fluctuations or both. Therefore measurement of this noise should yield information about the magnitude and energy distribution of these traps. Different experimental techniques have already been developed where MISFETs [28], [29], [30], MIS capacitors [31], ungated oxide-semiconductor structures [32] have been probed to extract interface trap densities. There have been also variations in the type of models used. There are several advantages of using noise techniques to measure interface trap densities: first, capability to probe into band gap energies close to band-edges and to energies even beyond the semiconductor band-gap; second, ability to probe into slow interface states which reside within the dielectric.

Among the semiconductor-insulator interfaces, HgCdTe-ZnS provides the greatest challenge, since (1): the amount of work done on this material is considerably less than that on Si and GaAs due to the limited applications of HgCdTe, and (2): HgCdTe is a material with more imperfections than that observed on Si and GaAs. HgCdTe and ZnS form the semiconductoı and insulator components of MIS structures, most commonly used for infrared (IR) radiation detection. There are many ways of operating an MIS detector which might produce varying amounts of $1/f$ noise in the detected signal. This, however, would not change the underlying mechanism(s) inherent to the MIS capacitor causing these $1/f$ fluctuations, as long as all other sources of noise such as photon noise, system noise and any other noise source caused by the signal processing circuitry are subtracted.

In our laboratory, we performed $1/f$ noise measurements on $HgCd_{0.71}Te_{0.29}$ MIS infrared detectors in the dark and subjected to IR radiation of 300 K [31]. These devices are used for midwavelength infrared detection. The devices were operated at 90 K in the correlated-double-sampling mode where the voltage across the MIS capacitor was sampled when the charge well is empty and right after the accumulation of minority carriers due to IR radiation generation. The noise magnitude was investigated as a function of integrated charge, integration time, and dark current. The normalized charge noise magnitude when compared for two different integration times was found to disprove bulk originated Hooge's theory. The charge noise spectral density was also independent of dark current at this temperature. Charge fluctuations due to interface state trapping and detrapping were explored. The k-p theory was utilized to compute the position of the Fermi level during the integration of charge in the dark. Then the effective trap state density was extracted from the noise data using the McWhorter Theory [33]. Within 40 meV around the valence band-edge, the effective trap density was found to vary between 3.3×10^{19} and 3.3×10^{18} cm^{-3} eV^{-1}, monotonically increasing towards E_V, which agrees closely with independent studies done on similar structures using capacitance and conductance techniques. Fig. 3 shows the effective trap density extracted from dark noise measurements for charge integration time of 30 μsec and various reset voltages which determine the depth of the charge integrating well under the ZnS [31]. Although at 90 K surface originated noise theory seems to agree with the experimental data, at lower temperatures ($T \leq 65$ K) another source is identified as the origin of $1/f$ fluctuations [34]. At these temperatures, the charge noise power spectral density was found to depend quadratically on the dark current. At higher temperatures, this quadratic dependence did not exist. We attribute the dark current to a mixture of tunneling and depletion-region-originated minority carrier generation which seem to be responsible for $1/f$ fluctuations in these structures for temperatures below 65 K.

It might be surprising at first sight that the trapping states are not responsible for $1/f$ fluctuations at lower temperatures in HgCdTe MIS photodetectors. As the temperature is varied, the position of the Quasi-Fermi level for electrons changes, in effect activating states with different band-gap energies. Since the energy distribution of the interface and dielectric states is not necessarily uniform, this leads to varying contribution of trapping-state-originated charge fluctuations to the total noise as the temperature is changed. It seems that at lower temperatures, this contribution is low enough that the $1/f$ fluctuations are dominated by the dark current due to tunneling and depletion-region-originated minority carrier generation.

The above discussion underlines the importance of locating the exact origin of $1/f$ fluctuations before attempting to utilize them for characterization purposes. A widely accepted model that has been experimentally proven at one temperature might not necessarily be used as a diagnostic tool at another temperature or even for another material.

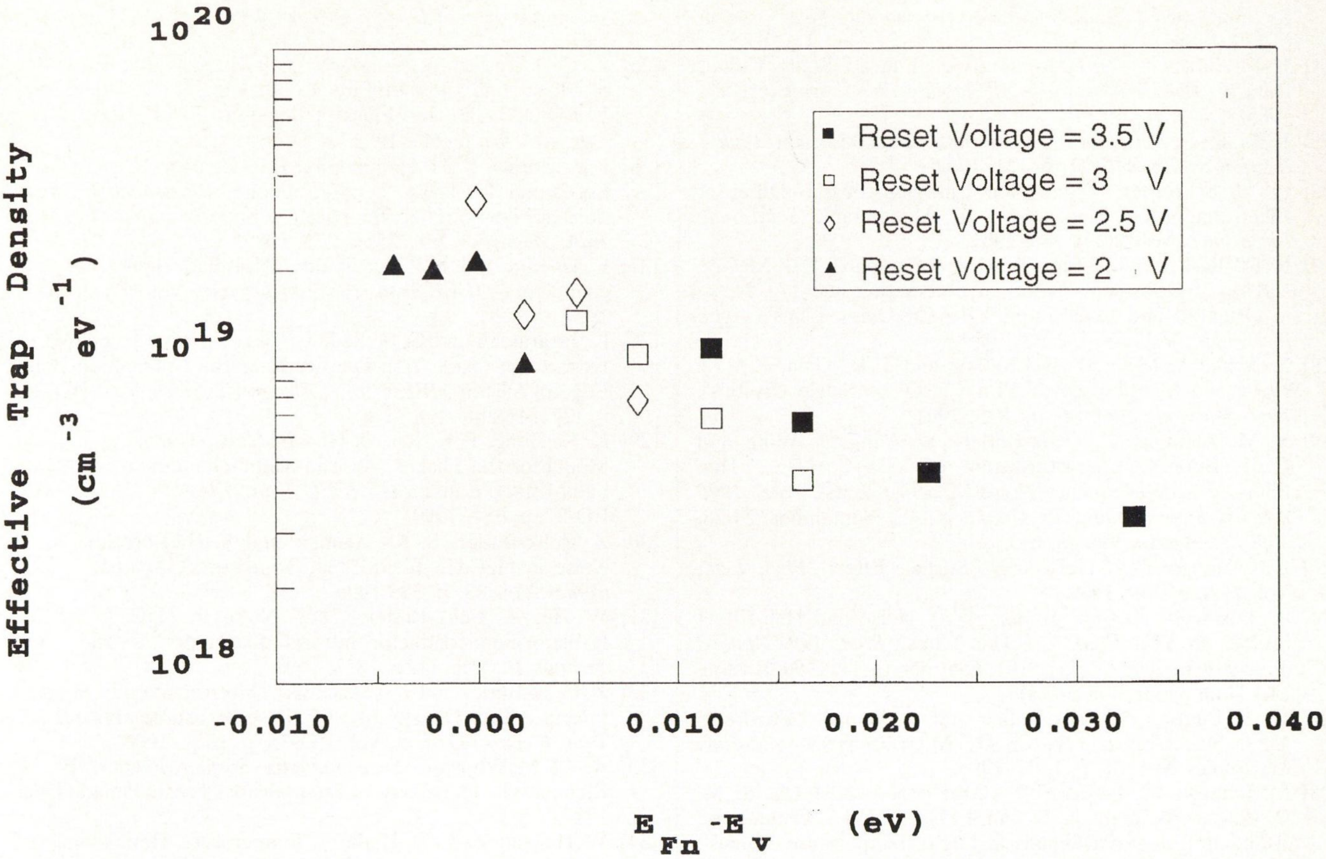

Fig. 3. $HgCd_{0.71}Te_{0.29}$-ZnS Interface States: Effective trap density extracted from dark noise measurements for 30 μsec. charge integration time. The reset voltage is the voltage applied to MIS gate before the charge integration starts to reset the device after the previous cycle.

5. CONCLUSION

Low-frequency noise measurements are finding more and more applications as a diagnostic tool in microelectronics. The method is simple, non-destructive, and inexpensive. Moreover, when developed fully, it provides as good or even better information about the probed parameter than conventional techniques.

6. ACKNOWLEDGMENTS

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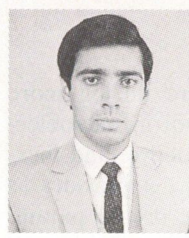
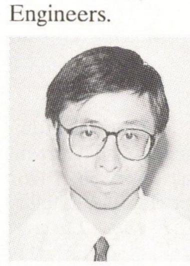
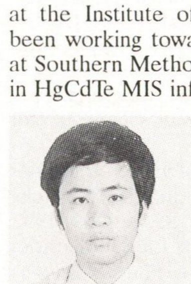
Zeynep Çelik-Butler received the B.S. degrees in electrical engineering and physics from Bogaziçi University, Istanbul, Turkey, in 1982. She received the M.S. and Ph.D. degrees in electrical engineering in 1984 and 1987, respectively, from the University of Rochester. She was an I.B.M. Predoctoral Fellow from 1983 to 1984, and an Eastman Kodak Predoctoral Fellow from 1985 to 1987. She joined SMU, Electrical Engineering Department in 1987, and held the J. Lindsay Embrey Trustee Chair for junior faculty from 1990 to 1993. She is currently an associate professor. She served in several technical review committees including the 1988 and 1989 IEEE International Electron Devices Meetings and the Annual Symposiums on ECS Electronic Materials, Processing and Characterization in 1988-1991. Dr. Çelik-Butler served in the Dallas Chapter of the IEEE Electron Devices Society first as a treasurer (from 1989 to 1992), then as a vice-chairman (from 1993 to 1994) and finally as a chairman (from 1994 to present). From 1988 to 1991 she was the IEEE student branch advisor at SMU. Dr. Çelik-Butler's research interests are characterization, and modeling of solid-state devices with a focus on noise properties of infrared devices, VLSI interconnects, sub-micron Field-Effect Transistors and high T_c superconductors. She has authored or co-authored more than 40 journal and conference papers in these areas.



Agha Jahanzeb received bachelor of science degree in electrical engineering from University of Wisconsin-Madison in 1992. He is currently a graduate student at Southern Methodist University in Dallas, Texas. His research interests include noise characterization and transport properties of high-T_c superconductors. He is a member of American Physical Society and Institute of Electrical and Electronics Engineers.



Wenmu He received the B.S. degree in electrical engineering from Peking University, Beijing, China, in 1985, and the M.S. degree in applied physics from the Institute of High Energy Physics, National Science Academy, China, in 1988. From 1986 to 1988, he was with Microelectronics R&D Center, National Science Academy working on X-ray lithography research. In 1988, he joined the Synchrotron Radiation Lab at the Institute of High Energy Physics. Since 1991, he has been working toward the Ph.D. degree in electrical engineering at Southern Methodist University. His research area is $1/f$ noise in HgCdTe MIS infrared detectors.



Ruojian Zhang received his M. S. in electrical engineering from Nanjing University (P. R. China) in 1990. He worked for Shenzhen Modern Advancement Co. from 1990 to 1991 on equipment maintenance of auto-testing system. He studied Physics at University of South Florida in 1992. Since 1992 he has been working on VLSI circuit reliability research at Southern Methodist University, Electrical Engineering Department, Ph.D. program. He is a member of IEEE.

RELIABILITY PREDICTION OF DIGITAL INTEGRATED CIRCUITS

ZS. KINCSES and P. BÁNLAKI

DEPARTMENT OF ELECTRONIC TECHNOLOGY
TECHNICAL UNIVERSITY OF BUDAPEST
H-1521 BUDAPEST

Nowadays it is getting higher and higher request for more reliable electronic instruments, which must work within extreme environmental conditions but they must be absolutely error-free in whole of the intended lifetime. This work describes a scrutiny method for digital silicon integrated circuits to detach ones that are damaged or having hidden failures. The introduced method is quite efficient and has no danger for perfect chips.

1. INTRODUCTION

The development of the integrated circuit's technology reached the reduction of number of the outgoing parts with defect and the increase of reliability. Typical industrial levels are less than 100 DPM (Defect Per Million) and 100 FIT (Failure In Time) [1], [2]. Moreover much better values can be found in the industry [3].

In present years a lot of expensive high-tech instruments and devices are frequently used. They have to work within excessive conditions in many cases. Repairing is almost never possible, or its cost is very high. Several times these instruments shield personal life or establishment of great worth. The sudden failure can cause great loss or injury. From this aspect the manufacturer of the high-reliability instrument is not interested in the statistics of the lot above all, but the individual properties of the given one that would be built in. There is a need to throw off the parts of shorter lifetime from the lot, without critical risk for faultless ones.

The conventional tests of digital circuits are difficult and need very much time. The complete examination is impossible in most cases of large integrated circuits, or its efficiency is little.

The noise analysis is widely used to examine materials, devices, circuits, instruments and systems. It is a common observation that inadequate material structures generate excess noise [4]. The noise spectrums [5] of the devices to fail earlier is markedly different from the better ones. By using suitable scrutinies the hidden signs can be shown. Moreover the reasons and the location of the fault can be found occasionally.

The new method accomplished in the frame of our investigations gives a powerful possibility to inspect comprehensively the whole internal structure of the digital integrated circuit. The technique is advantageous from several aspects. It also can give information to IC designers. This work is based on experimental proofs.

2. DIGITAL IC'S INSIDE

There are several types of digital integrated circuits from the viewpoint of internal construction and operation. It can be classified as voltage level logic or current logic. Another way of assortments is saturation. From the technology there are bipolar, MOS, and up to now CMOS IC's. The general construction is the next: there is a protection on the input comparators. It is followed by the logical function. It is followed by the level regenerator and output driver with its shielder. These parts are not separated physically on all occasions.

Because the CMOS technology is at the front of development nowadays the present work concentrates on it. The CMOS structure is relative simple and easy to consider. In this technology the fundamental gate, the inverter, is as shown in Fig. 1.

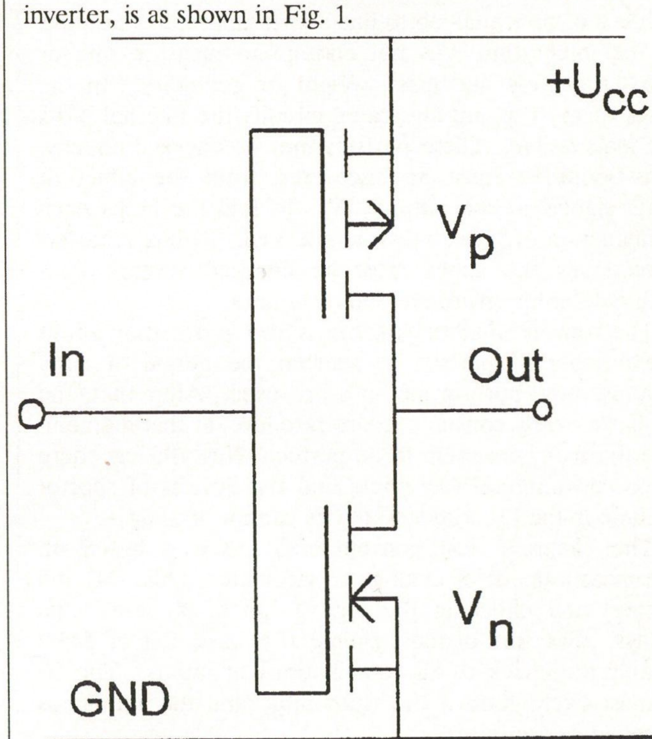


Fig. 1. A CMOS inverter

It is a serial connection of a p-channel enhancement MOS transistor and an n-channel one. When the input voltage is high the n-MOS transistor is open and the other one is closed. Fig. 2 shows the operating ranges from the aspect of threshold voltages related to the supply voltage. In an ideal inverter the threshold voltage is equal to the half of supply voltage. The changes on the output are abrupt-like. In the case of higher threshold voltages there is a voltage range, where both transistors are closed, so the output is a floating potential. On the contrary when the thresholds are low, there is a voltage range where both transistors conduct. That time some very high current is flowing through the inverter and dissipation can destroy the IC. To avoid it the CMOS inputs mustn't be on floating potential.

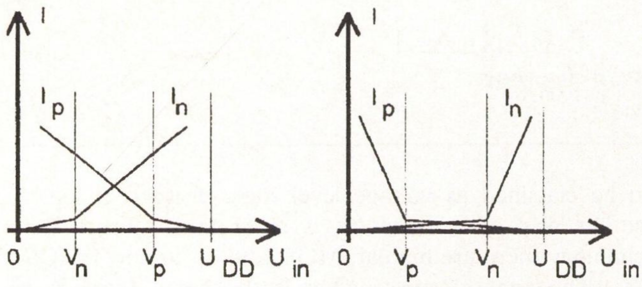


Fig. 2. Operating ranges of CMOS vs. threshold voltage

In the static situation there is no current flowing through. The dynamic dissipation is rising with the frequency.

3. CONVENTIONAL TESTS

It is a complicated job to find a part failureless. Because of the integration it is not enough to measure one or two parameters like mass, weight or geometry. In the cases of most digital integrated circuits the internal parts are inaccessible. These parts cannot be checked directly. The operation must be discovered from the effect of input signals to the outputs [6]. To find the faults each combination of states must be checked. To find parasites successfully the states must be checked several times within different environmental conditions.

The conventional reliability tests give information about the number of parts. To shorten the period of early failures some burn-in methods are used. After that, the lot has a nearly constant failure rate level at that moment, and the members seem to be perfect. Nevertheless there is no conventional way could find the devices of shorter lifetime in the lot. Hidden failures cannot be shown.

The electrical test conventionally used is based on the measurement of analogous parameters (AC, DC and timing) and checking the logical functions, levels and delays. The test of most simple IC's (like SSI or MSI) is only the check of all combinations of inputs. The IC is most likely good if the truth table and the analogous parameters are in order.

In the cases of LSI, VLSI and ULSI circuits the analogous parameters of inputs and outputs cannot say too much about the inside region of the IC. Because of the digital operation they are nearly independent from the inner part. The test of splited parts is generally unsuccessful, because the number of leads is fewer than the gates. In most cases the control signals for input routes are different and conflicting with the controls for output routes. The largest part of nodes inside is not verifiable.

Designers can give a little support. Nowadays there are libraries of implementations of digital functions. The greatest part of the work of the designers is to make the circuit suitable for testing [7]. A relatively new method in the design is making auxiliary shift-sequence to see internal nodes, or construct built-in selftests. But its efficiency is limited.

4. SUPPLY CURRENT TESTS AS A NEW METHOD

There is a way to examine the whole integrated circuit without physical destruction or accelerating failure mechanisms. There are two power supply connections that are very close to each element. At normal conditions the supply voltage is constant.

The power supply current carries detailed information about the whole system.

There are several ways to study the supply current as the carrier of information about degradation. The static or the quasi static current test can show higher dissipation. The measurement of static current can be a usual part of conventional quality investigations. The current-noise test can show other information of physical structure, because the noise is increased in cases of damaged or less homogeneous structures.

The most interesting part of this area is the transient current test. When a state is changed in an IC it produces a current pulse according to its physical construction. TTL ICs change the current level. TTL and CMOS ICs produce a current peak at the moment of the state change. Because of the time delay the changes at following gates appear at different times. In the course of the state changes a signal flow is running through the device. It is a well-observable current transient.

Analysing a current transient it can show higher or lower current level between the peaks, larger or smaller current peaks, abnormal peaks with unusual shape or timing. It can be considered what physical or technological problems are causing them. Theoretically, the exact place of the fault can be found.

5. EXPERIMENTAL RESULTS

In the frame of this investigations metal gated CMOS ICs were examined. It was 1000 pieces of HCF 4027 dual JK flip-flop. First the circuit was studied to generate an adequate input test pattern that can change the output of each internal gate caused by each internal or external input. The signal was programmed into a crystal-controlled pattern generator that gives not only the test vectors, but the controls for the digital storage oscilloscope too. It was HITACHI Type VC 6065 in our case. The current was measured on a 50 Ω resistor. The digital results were accepted and evaluated by PC through IEEE 488 bus.

Before the transient current measurement of any IC they were checked by the conventional way. According to the level of the quality no wrong part was found. The lot seemed to be perfect by this viewpoint.

The measured current transients were averaged (Fig. 3) and the results were evaluated and considered. Fourteen devices seemed to be questionable. There were some peaks with higher or lower amplitude (Fig. 4), irregular peak at wrong place (Fig. 5).

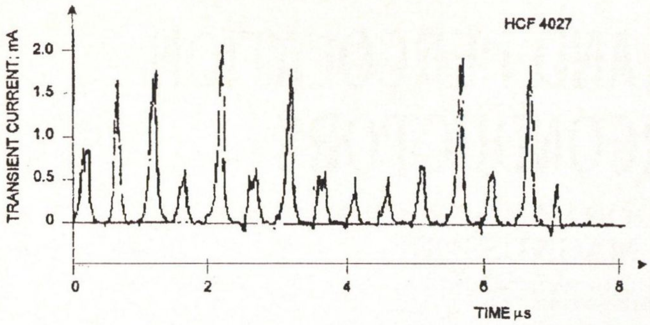


Fig. 3. Average current transient

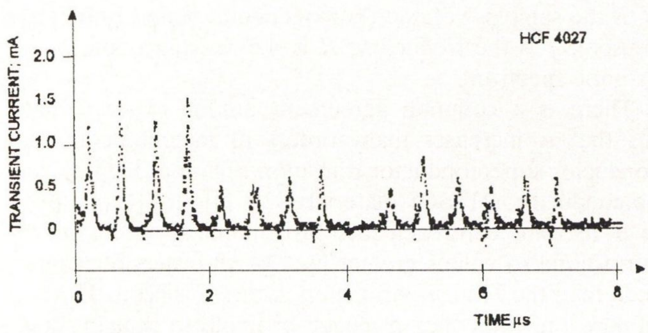


Fig. 4. Smaller peaks

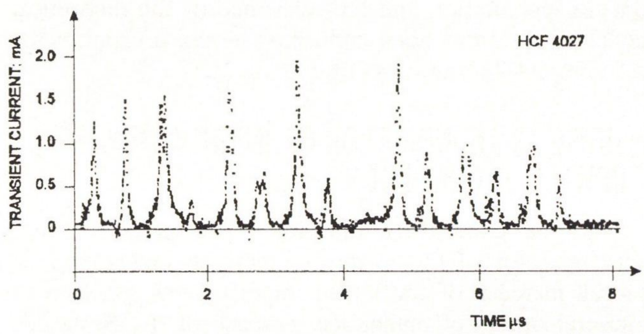


Fig. 5. Irregular peaks

After the new current test all ICs were taken into a furniture for 2000 hours at 150°C maximum storage temperature. It is equivalent to $0,6 - 200 \times 10^6$ hours at 20°C depending on the effective failure mechanism.

Proceeding the high temperature storage the functional and parametric test was done again. Two devices were defected. Both were members of the questionable 14. It is possible that longer term of high temperature storage or another investigation (e.g. storage under supply power, work under higher supply voltage, etc.) would give closer correlation between the obtained and the estimated results.

6. SUMMARY

The request for high reliability devices makes necessary new guidelines in reliability prediction. In contrast to the conventional methods the new technique allows to investigate the given device to be built in, without any danger for the perfect ones. The new way is the transient current test.

Theoretical and experimental studies indicate a relation between irregular current characteristics and hidden failures. The considerations were presented. Experimental justification was given with 2000 amount of JK flip-flops. The first results suggest a great promise, but there is a need for more investigations and a serious demand for further results in practice.

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Zsolt Kincses is a postgraduate student for Ph.D. degree at the Department of Electronic Technology. He studied mechanics in secondary school, and graduated in electrical engineering at the Technical University of Budapest in 1992. The National Scientific Research Found (OTKA) granted him support for his researches. His interest is low-frequency electronic noise, integrated circuits, high-frequency semiconducting devices and reliability aspects.

Pál Bánlaki graduated in electrical engineering at the Technical University of Budapest in 1980 and received dr. univ. degree in 1983. He took part in reliability investigations and space research project VEGA and Phobos. He has engineering practice in Canada and Germany. Now he is a teacher in a secondary school and a part time research worker at the Department of Electronic Technology. His interest is electronic noise measurement and physics, and reliability investigations.

CONDUCTANCE NOISE AND PERCOLATION IN HIGH Tc SUPERCONDUCTORS

L. B. KISS*, P. SVEDLINDH, M. BJÖRNANDER, P. NORDBLAD
F. MASSZI and J. MAGNUSSON

UPPSALA UNIVERSITY, INSTITUTE OF TECHNOLOGY
BOX 534, UPPSALA, S-75121, SWEDEN

There is a common agreement among experimentalists that the normalized resistance noise increases many orders of magnitude in the conductor-superconductor transition region of high T_c superconductor (HTSC) materials. The normalized noise can reach such "astronomical" values which correspond to values greater by 13–18 orders of magnitude than the Hooge-parameters found in silicon. Another important property is the scaling behaviour of the normalized noise versus the resistance. Experimentally verified theoretical attempts to explain this particular behaviour of the noise in the transition are reviewed. Critical remarks on interpretation in terms of the Voss-Clarke theory are presented. Open problems about the role of non-Ohmic behaviour are considered and some relevant experimental results are shown.

1. INTRODUCTION

Random conductor-insulator mixtures and random conductor-superconductor mixtures show excessively strong $1/f$ noise in their resistance. Under proper conditions, the strength of the noise is a power function of resistance. These effects can be understood in terms of percolation models of these materials. This modern topic concerns not only fundamental science but also modern electronical technology due to the applications of high T_c superconductor (HTSC) materials and thick film resistors. Professor Ambrózy had been involved in the research of noise in thick film resistors [1]. In this paper we are concerned in the complementary topic of percolation: the noise in HTSC materials.

In high T_c superconductors (HTSC), the carrier concentration is (usually) unknown and it is a steep, (usually) unknown function of the temperature T. For that reason, at the characterization of the $1/f$ noise of the HTSC materials, the noise is usually normalized to the atomic number of the sample ($\sim 10^{22}$ cm^{-3}), or to a somewhat smaller (but also fixed) number of atoms ($\sim 10^{21}$ cm^{-3}) which provide free charge carriers [2], [3], [4]. Another reason to use the atomic numbers is that the percolation models of noise, which have been proven to be working here, are based on this kind of (volume) normalization [2], [4]. As a consequence, the $1/f$ noise parameter α of HTSC materials is given in the same way as the Hooge-parameter [5], [6], [7] of metals:

$$\alpha(T) = n_a V f \frac{S_r(f,T)}{R^2(T)}, \tag{1}$$

* On leave from: JATE University, Dept. of Experimental Physics, Dóm tér 9., H-6720 Szeged, Hungary

where, n_a is one of the above mentioned atomic densities, V is the samples volume (homogeneous sample geometry is used), f is the frequency, R is the resistance and S_r is its noise spectrum.

There is a common agreement among experimentalists that α increases many orders of magnitude in the conductor-superconductor transition region of high T_c superconductor (HTSC) materials [2], [4], [8]-[19], moreover, its value can reach such "astronomical" values which correspond to values greater by $13-18$ orders of magnitude than the Hooge-parameters found in silicon. Up until now, only percolation models are able to explain these peculiar characteristics of the conductance noise of HTSC materials [2], [4], [10]-[15]. Note that until now, the existence of percolation and the relevance of the theoretical model [4], [11] has been confirmed by noise experiments in 5 different laboratories [10]-[15].

2. NOISE AS AN INDICATOR OF IMPROVEMENTS IN HTSC TECHNOLOGY

In normal conductors, the Hooge-parameter is a sensitive indicator of the quality of material technology, as a small increase of the lattice imperfections can lead to a several orders of magnitude increase of α. Naturally, the same effect controls the α values found in the normal conductor state of HTSC materials.

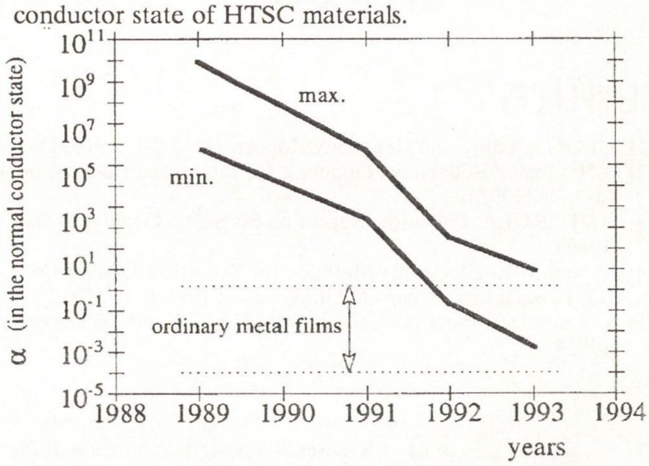

Fig. 1. Evolution of α during the last years, as an indication of the improvements of HTSC technology

On Fig. 1, the evolution of α values can be seen during the last years [2], [4], [9], [16], [17], [18]. As other technical parameters in film samples are concerned, the critical current density and the temperature-width of the transition are: $\sim 10^4$ A/cm^2 and $\sim$ 10 K at the left-hand side; and 10^6 A/cm^2 and $\sim$ 10 K at the right-hand side of the figure.

3. NOISE IN THE TRANSITION TEMPERATURE REGION

First of all: note that the most useful information is contained *not* in the α versus T plot (Fig. 2) or not in the noise versus T plot (Fig. 3), but in the $\alpha(T)$ versus $R(T)$ plot (Figs. 4, 5). The most informative characteristics has turned out to be the behaviour of the $\alpha(T)$ versus $R(T)$ in the transition temperature region (Figs. 4, 5).

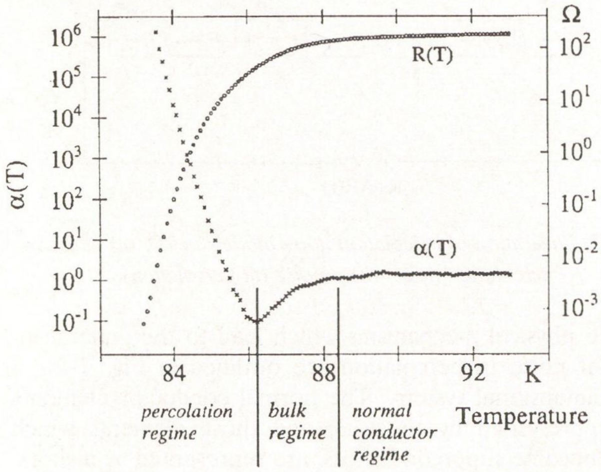

Fig. 2. General behaviour of R and a versus T in an in-situ annealed YBCO sample made by co-evaporation (at 10^5 A/cm^2 current density)

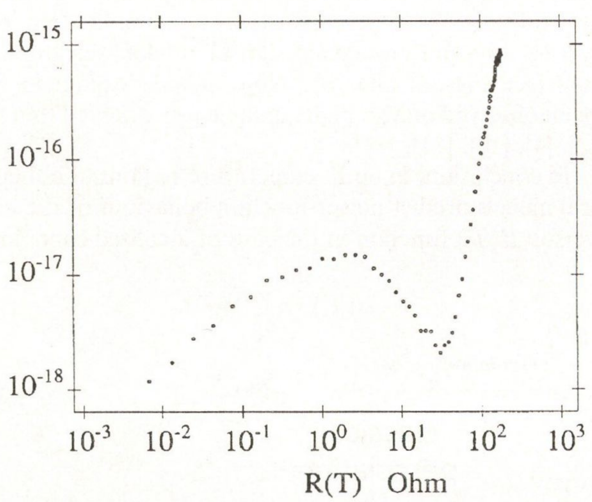

Fig. 3. The same results as plotted in Fig. 2, plotted here in a different fashion. The noise versus R (or versus T) often show a peak in the transition region. Still, these fashions of plots have been usually the less informative ones.

First, in the upper part of the transition temperature region, a is often decreasing with decreasing temperature (with decreasing $R(T)$) (see Figs. 2, 4) or staying at a constant value (see Fig. 5). We call this temperature regime *bulk* region [2], [4], because the behaviour of α is interpreted as the result of temperature-independent current density distribution, just like within the bulk of a grain. The constant $\alpha(T)$ is interpreted as the result of *number fluctuations* of normal charge carriers which induce the same relative number fluctuation of Cooper-pairs in a correlated way. On the other hand, when $\alpha(T)$ is decreasing with decreasing temperature and the decrease is significant, $\alpha(T)$ turns out to be proportional to $R^2(T)$, see Fig. 4. That observation is interpreted as the result of *mobility fluctuations* of normal charge carriers, which are obviously not causing fluctuations in the conductance of Cooper-pairs and this fluctuation decays as $R^2(T)$ when superconductivity takes over. Those cases, where $\alpha(T)$ is decreasing with decreasing temperature, but the decrease is small and less steep than $R^2(T)$ are interpreted as the result of simultaneous number fluctuations and mobility fluctuations.

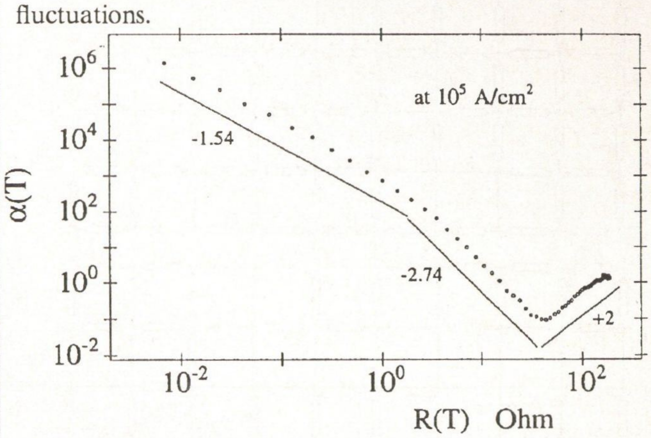

Fig. 4. The most informative fashion of plots: α versus R in a log-log scale. The same results are plotted here as on Figs. 2 and 3. From right hand side to left, the different scaling exponents represent the following noise phenomena: mobility fluctuations (+2); p-noise in 3D ($\cong$ 2.74) and p-noise is 2D ($\cong$ 1.54).

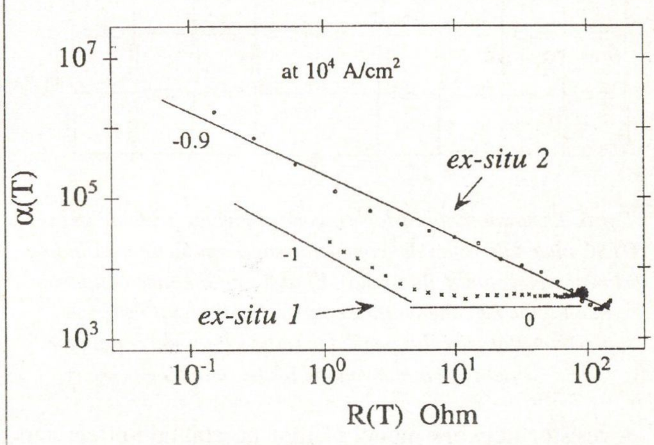

Fig. 5. Scaling of the noise at 10^4 A/cm^2 in ex-situ annealed samples made by co-evaporation. Number fluctuations (0) and classical percolation noise ($\cong -1$). The last phenomenon often occurs in weaker quality samples, like the ex-situ annealed ones or sintered ones are.

In the lower temperature part of the transition region (see Figs. 2, 4, 5), or sometimes in the whole region (see Fig. 5), α is radically increasing with decreasing temperature (with decreasing $R(T)$). We call this temperature regime percolation region [2], [4], because the behaviour of α is interpreted as the result of percolation of current through a conductor-superconductor random mixture with a temperature-dependent random composition and current density distribution.

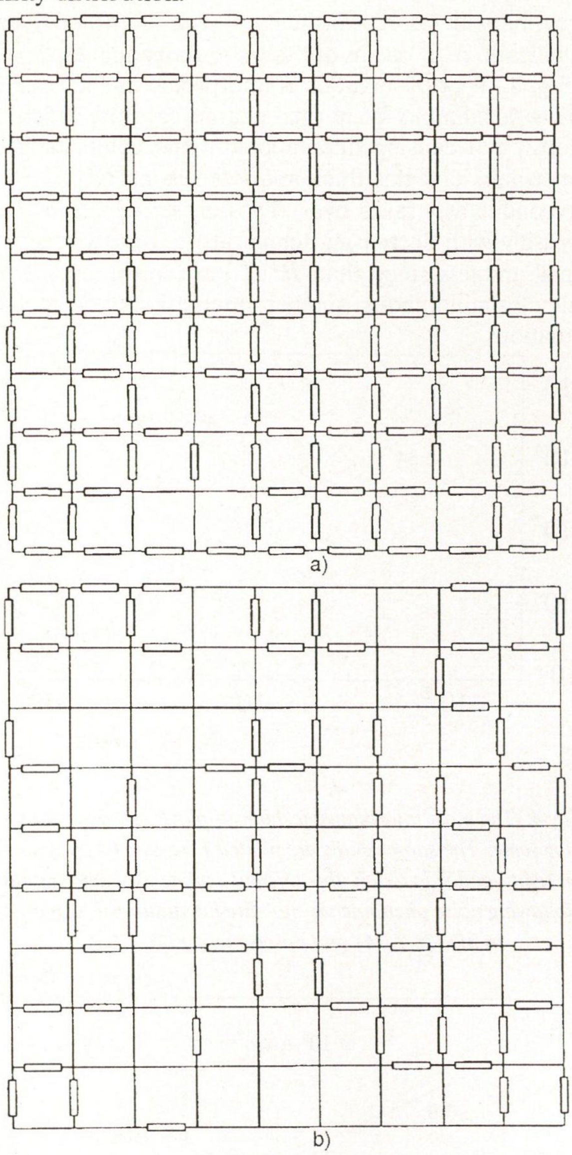

Fig. 6. Illustration of resistor-network model for percolation in HTSC materials: a) in the conductor-superconductor transition (below the percolation threshold); b) at T_c, in the superconductor state (at the percolation threshold). The shortcuts represent superconductor elements, while the resistors normal conductor elements in the volume of the sample.

A resistor network model of that percolation effect can be seen in Figs. 6 a), b): the lower the temperature the larger the number of superconductor elements (shortcuts) in the network. At T_c, a superconductor random island formed by the elements will make a shortcut between the two measuring electrodes. The sharply increasing $\alpha(T)$ with decreasing temperature is a natural consequence of approaching the percolation threshold, which in the language of $1/f$ noise research, is equivalent with a sharp decrease of the *effective noise volume* of the sample [6]. The existence of percolaticon is proven by not only the sharply increasing $\alpha(T)$, but rather by the fact that $\alpha(T)$ is most often scaling with $R(T)$ with power exponents very close to the universal values predicted by different percolation theories.

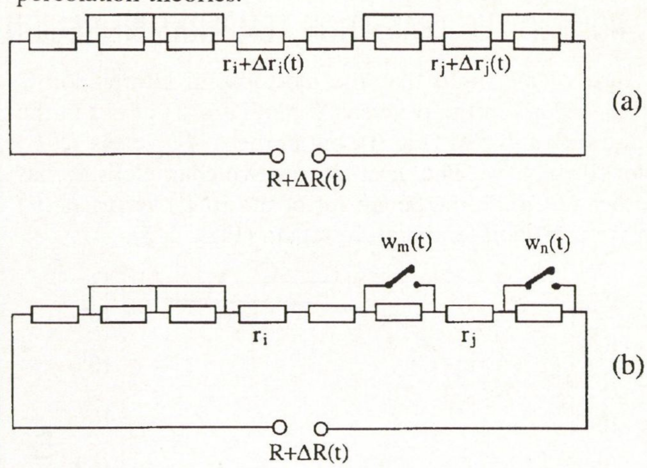

Fig. 7. Illustration of percolation noise models in 1D. (a) classical percolation noise; (b) p-noise (at percolation).

The physical mechanisms which lead to the generalization of noise in percolation are outlined in Fig. 7 for a one-dimensional system. The normal conductor elements are represented by resistors, and those elements which have become superconductors, are represented by a short-circuited resistor. In the case of *classical percolation noise* [2], [4], [11], [20], the resistance of normal conductors is fluctuating (see Figs. 5 and 7a). In the case of *p-noise* [2], [4], [11] (a newly discovered percolation noise phenomenon), some elements are randomly switching between normal state and superconductive state and the total noise will be a properly weighted sum of random-telegraph signals (see Figs. 4 and 7b). Note, *p-noise* exponents have been observed only in high quality c-axis oriented thin films [2], [4], [10], [11], [13].

In conclusion, in both temperature regions, the theoretical models predict power-function behaviour of the $\alpha(T)$ versus $R(T)$ function in the case of idealized conditions:

$$\alpha(T) = R^X(T). \tag{2}$$

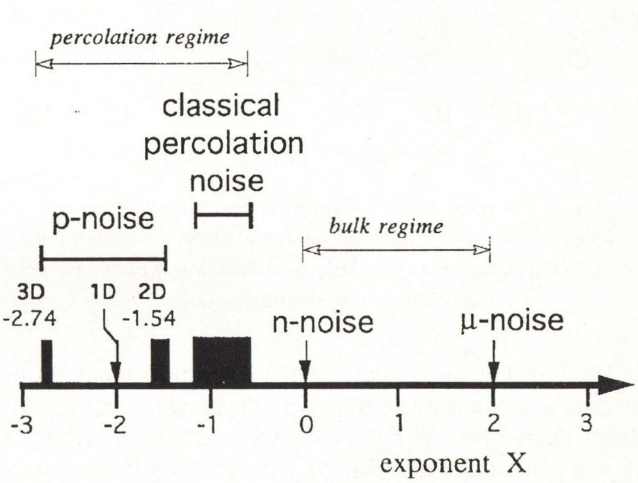

Fig. 8. Summary of those theoretical exponents which one's existence has been experimentally verified in HTSC materials

In Fig. 8, we summarize those exponents X which are given by theoretical predictions and verified by noise-measurements on HTSC materials (mostly thin films).

4. NONLINEAR EFFECTS

The HTSC materials do not show Ohmic behaviour in the transition temperature region. This fact rises serious questions about the effects of non-linearity and about the possible limitations of linear noise models. Furthermore, in the framework of percolation, one can expect that not only the noise and the resistance but other transport quantities become scaling quantities, when the system is sufficiently close to the percolation threshold.

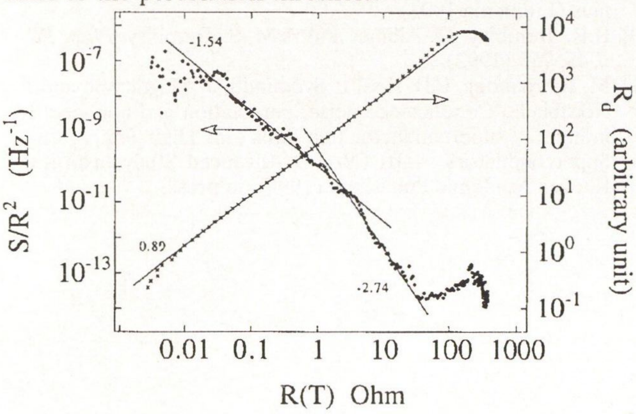

Fig. 9. Scaling of α and the differential resistance R_d in laser ablated YBCO films. The results indicate the necessity to develop nonlinear percolation models.

The last assumption has experimentally been checked for the differential resistance $R_d = dU/dI$. We found [21] that the differential resistance is scaling with the resistance $R(T)$ and the empirical scaling exponent is somewhere between 0.7 and 0.9, which might mean that there is a new universal exponent here. In Fig. 9, one of the best plots can be seen. As a comparison the plot of $\alpha(T)$ is also presented there. As it can be seen, the scaling of the differential resistance and the scaling of the noise ($\alpha(T)$ by the theoretical p-noise exponents) are setting on remarkably in the same range. This observation is another convincing evidence of the existence of percolation in HTSC thin films.

5. CRITICAL REMARKS

We would like to note that the experimental study of high quality HTSC materials in the transition region needs serious technical efforts due to the large temperature-derivative of the resistance. The noise of the temperature regulator often has to be smaller than $10^{-8}K/\sqrt{Hz}$ to avoid measurement of noise due to the noise of the temperature control. There are several erroneous papers in the literature, where the authors try to interpret their artefacts due to noisy temperature control in terms of the Voss-Clarke equilibrium temperature fluctuation model, which is nonsense in this case. For details, further considerations and advice, see [2].

6. ACKNOWLEDGEMENTS

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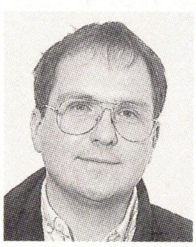
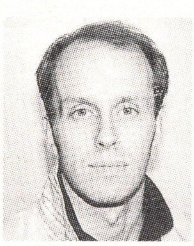
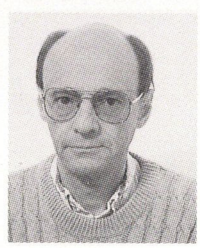


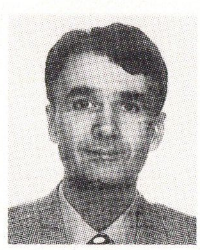
László Béla Kiss received the M.Sc. degree in semiconductor physics and Dr. degree in experimental statistical physics at JATE University (Hungary), in 1980 and 1984, respectively. He received the Docent degree (habihtation) in solid state physics at Uppsala University (Sweden), in 1994. In 1988, he established a noise research laboratory at JATE University, Department of Experimental Physics (Hungary), where he holds a permanent faculty position. He had been a Guest Scientist at Eindhoven University (The Netherlands), The University of Kln (Germany), the National Research Laboratory for Metrology (Japan) and The University of Birmingham (England). Presently, he is a Visitmg Scientist at Uppsala University (Sweden) where his mean scientific responsibility has been noise in high T_c superconductors. Other fields of his activity: stochastic resonance, 1/f noise, universal conductance fluctuations, diffusion noise, and application of noise measurements as "least disturbing way" of obtaining information on various systems.



Peter Svedlindh received the M.Sc. degree in material physics and the Ph.D. degree in solid state physics from the Institute of Technology, Uppsala University, in 1980 and 1985, respectively. His current research interests are in disordered magnetic materials, ultrafine magnetic particles, materials showing giant magnetic resistance effects and high T_c superconductors. Since 1986, part of his professional time has been devoted to applied magnetic research at the National Defence Research Establishment in Stockholm.



Per Nordblad received his Ph.D. in Solid State Physics at Uppsala University, in 1980. Presently, he is Assoc. Prof. at Uppsala University. His current research activities concern experimental studies of fundamental physical properties of magnetic and superconducting materials. Author of about 110 articles in international physics journals.



Ferenc Masszi received the M.Sc. degree in electrical engineering and the Dr.Techn. degree in semiconductor electronics from the Budapest Technical University in 1976 and 1979, respectively. Since 1984, he has been with the Electronics Department at the Institute of Technology, Uppsala University, Sweden, now as Assoc. Prof., being responsible for the computer simulation group in the Scanner Lab. His interests include all aspects of physical modelling of semiconductor devices. He is a member of the IEEE.

Malin Björnander and **Johan Magnusson** are Ph.D. students at Uppsala University. At the completion of the paper there was no bio and photo available on them.

THE INFLUENCE ON THE NOISE OF TRIMMING CUT-PATH IN THICK FILM RESISTORS

G. PAPP, ZS. KINCSES and I. HAJDÚ

TECHNICAL UNIVERSITY OF BUDAPEST
DEPARTMENT OF ELECTRONIC TECHNOLOGY
H-1521 BUDAPEST, HUNGARY

Manufacturers of thick film circuits often use trimming algorithms cutting the resistors by laser-beam to adjust the resistance to the desired values. The electronic noise in thick film circuits is related closely to material, technological and topological parameters. This work describes the influence on the noise of the cut-path, confirmed with experimental data base; and introduces a strategy helping to find optimal trimming route, yielding resistors with higher reliability and reduced noise.

1. INTRODUCTION

Parameters of laser trimmed thick film resistors are different from the untrimmed ones'. Cutting off the resistor the resistance rises continuously. In trimmed resistors not the whole area of the film takes part in the current transfer, but the current density becomes higher in the active part nearby the cut-off. The current density distribution of the whole resistor depends on the shape of the cut-path seriously [1].

Another important effect originates from the heat of the laser-beam. In the edges of the laser kerf the material of the resistor melts. Therefore its metallurgical and electrical parameters are changed. This area is named as Heat-Affected-Zone (HAZ). In most cases the resistivity is increased after the heat treatment.

Using the circuit at maximum dissipation the resistor can be affected by the heat of the increased temperature. If the affect increases the resistivity, it rise the level of the current density. So the affected point being on higher temperature is moving accelerated till it finally 'cuts' off the resistor. The period of this effect can be in a wide range of time.

There is a close correlation between the $1/f$ noise and the lifetime of thick film resistors. The reason of the shorter lifetime and higher noise is the inhomogenity of the electronic potential as it is described below.

2. NOISE AND CURRENT DENSITY DISTRIBUTION

Resistor pastes generally used contains several components: conductive phase, glass frit, organic solvents, vehicles. The conducting mechanism in fired thick film resistors is very complicated. Till now there is no consistent model that can be explain each features.

The most published, but even they most mysterious area of the noise research is the flicker-noise. Its spectral density (1) is nearly $1/f$. In the low frequency range the flicker-noise is the dominant component of the noise in thick film circuit.

$$S(f) = const\frac{I^{\beta}}{f^{\alpha}}\Delta f, \qquad (1)$$

where I is the current, f is the frequency, Δf is the bandwidth; $\beta \approx 2$ and $\alpha \approx 1$. The sources of this noise is not known yet [2], [3].

Under the trimming process a part of material of resistor film is removed by the laser-beam. The streamlines of current are curved, and the current transfer cross-section is decreased. The electric field and the current density comes higher, depend on geometrical parameters, and finally it determinates the noise power of the resistor [4] (2).

$$\overline{(u)}^2 = \frac{\alpha \cdot \Delta f}{f} \cdot \frac{R^2 \cdot t}{\rho \cdot V^2} \int\int |\mathbf{E}|^4 dxdy, \qquad (2)$$

where R is the resistance, t is the thickness, ρ is the resistivity of the resistor and V is the voltage. The values of variables depend on material properties, the value of integral is depend on the geometry.

As it can be seen in (2) the inhomogenity of the electric field (E) and strongly curved trajectories of potentials has considerable effect on the noise. And as (1) shows the higher current increases the $1/f$ noise.

3. THE BASIC SELECTION OF CUTTING PATHS

The Fig. 1 illustrates some typical types of the trimming cuts, potential trajectories and the circle shows the points of the maximum dissipation.

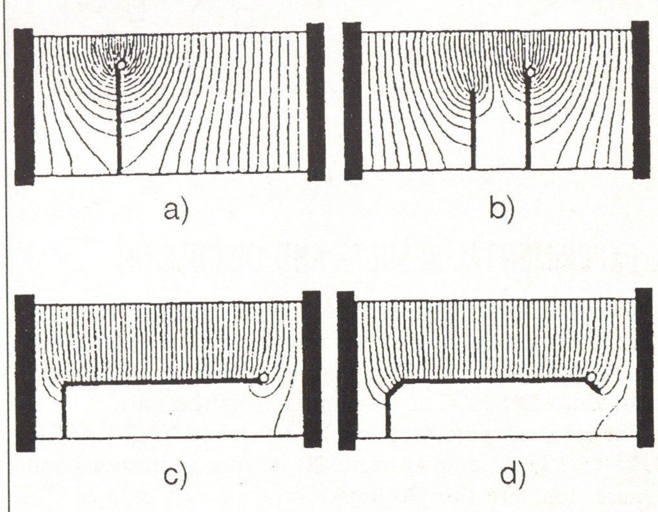


Fig. 1. Trimming cuts

On Fig. 1 the **A** illustrates the single plunge, **B** the twin plunge, **C** the conventional L-cut and **D** the Modified L-cut (ML). Because of the inhomogenity of trajectories the calculated maximum dissipation at the same resistance in this setup in single plunge cut was 11 times higher than in the modified L-cut [1]. It means there is a close relation between current inhomogenity and lifetime.

4. NOISE MEASUREMENT METHOD

According to IEC recommendation [5] Fig. 2 shows the applied measurement setup. In this work a modified and rebuilt preamplifier of that one fabricated by Prof. Ambrózy was used [6].

The current was flowing through R_m was provided by four 4.5 V standard batteries. Since the current is:

$$I = \frac{U_{DC}}{(R_s + R_m)}, \tag{3}$$

where I is the current, U_{DC} is the biasing DC voltage, R_s is the serial separator resistor, R_m is the sample resistor under test. The total squared noise voltage is:

$$u^2 = 4kT(R_s \times R_m)\Delta f + k^2 U_{DC}^2 \frac{R_m^2 R_s^2}{(R_m + R_s)^4} \frac{\Delta f}{f}, \tag{4}$$

where the first term is the thermal noise and the second is the excess one [6]. The amplifier and the batteries were enclosed in a grounded steel box. The noise spectra was measured in the frequency range 1.6 Hz...20 kHz with real-time digital frequency analyser type Brüel&Kjær 2131 controlled by PC 486. During the measurement the excess noise was separated, and noise index was calculated.

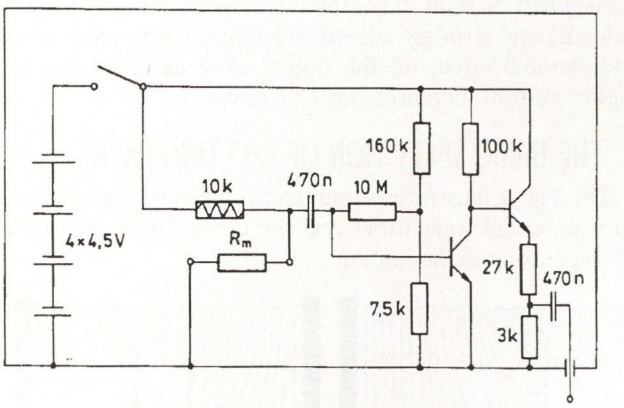


Fig. 2. Applied measurement setup

5. EXPERIMENTAL RESULTS AND DISCUSSION

Several resistor samples, size 1 and 6 mm, were examined. The material of resistors was DuPont HS8021 paste on alumina substrate. The peak temperature of the heat treatment was 850°C at 10 minutes from the start.

The resistors were prepared by Nd doped YAG LASER (Q-rate: 2 kHz, lamp current: 20 A, speed: 50 step/s and 5 pulse/step, bite size: 80 μm.)

A group of resistors were cut by the conventional L-cut. Another group of resistors were cut with the same depth and length as the L-cut, but the corner was chopped by 700 μm length and 45 degree lines. Its name is Modified L-cut.

The remark HAZ mean that resistor was cut by the LASER from one contact to other with parallel lines, which did not disturb the streamlines. The small changes of noise index caused by damaged resistor material in HAZ.

Table 1 shows that, the main changes of noise index caused by the geometrical effect were supported by simulation results [7].

We found that the Modified L-cut results in more homogeneous distribution of electric field and current density, but its effect on the resistance changes caused directly by the cut is negligible.

Table 1. Noise index of samples [dB]

Resistor	Trim routes shape			
size mm	untrim-med	HAZ	ML-cut	L-cut
6×6	-18.95	-18.79	-18.26	-17.87
6×3	-18.79	-18.61	-17.47	-17.37
6×2	-19.43	-18.69	-17.23	-16.68
3×6	-18.82	-18.21	**-18.34***	-17.60
3×3	-19.03	-18.67	-18.48	-18.04
3×2	-19.27	-18.64	-18.13	-18.06
2×6	-19.11	-18.04	**-18.30***	-17.78
2×3	-18.70	-18.27	-18.09	-17.69
2×2	-19.46	-18.93	-18.23	-17.08
1×6	-18.48		-17.41	-17.29
1×3	-19.11		-18.32	-17.37
1×2	-19.70			-17.52

* means devious results

6. SUMMARY

In thick film circuits the electronic noise and the reliability of the circuit have close relation with material, technology and topological parameters. In this work the influence on the noise of the cut-path was considered. Geometrical dependence of $1/f$ noise was presented and illustrated. Material and geometrical actions of laser trimming were summarized. Applying this knowledge a model was derived from which the impact of the inhomogenity of electric field and current density distribution could be calculated to analyze trimming routes.

We found that the Modified L-cut, which is chopped by 45 degree line at its end, has more homogeneous electric field, lower noise level and lower temperature at the point of maximum dissipation although the resistance is not changed seriously.

7. ACKNOWLEDGEMENT

Experimental works and the noise measurement were partly done in the laboratory supported by the Hungarian National Scientific Research Found through the Grant No. 7557.

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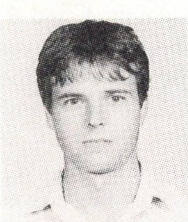
Zsolt Kincses Photograph and biography on page 19.



Gábor Papp received his degree in electrical engineering from the Technical University of Budapest (TUB), Hungary in 1991. From 1991 to 1994 he works as a postgraduate student in the TUB, Department of Electronic Technology, with particular interest in computer simulation of laser trimming process. He is member of ISHM Hungary. Recently, he works as a technology adviser of several private companies.

István Hajdú graduated in electrical engineering at the Technical University of Budapest, in 1969. He is working at the Department of Electronic Technology TUB, and gives lectures both in Hungarian and English on electronics technology. His special interest area is the thick film integrated circuits. He is a vice president of the International Society for Hybrid Microelectronics Hungary. He has more than 40 publications.

CONTACT NOISE MEASUREMENT ON METAL-GaAs MULTISPOT TEST PATTERNS

I. MOJZES, B. KOVÁCS, and ZS. KINCSES

TECHNICAL UNIVERSITY OF BUDAPEST
DEPARTMENT OF ELECTRONIC TECHNOLOGY
H-1521 BUDAPEST, HUNGARY

The metal-semiconductor contacts are essential building blocks of any semiconductor device. This work describes a method that is applicable to evaluate the noise spectra of ohmic contacts applying the widely used TLM (Transmission Line Model) test pattern. The results of the noise measurement are compared with results of the contact parameter measurement. The comparison shows that the contact noise has an 1/f component exceeding the thermal noise attributed to the contact resistance (R_C).

I. INTRODUCTION

The metal-semiconductor contacts are important parts of semiconductor devices. Although a lot of effort was devoted to study the physics and the technology of metal-semiconductor junction, there are certain open questions in this field. One of these questions is the relation between the obtainable noise spectra and the state of degradation of the ohmic contacts. To study this question an accurate method is required to separate the noise spectra of the ohmic contact from the noise generated in other parts of the device or the test pattern.

Several device models are widely used. These models generally do not care about the location of noise sources. Localization of the noise source can be carried out by measuring the bias or temperature dependence of the noise. In contrast to that in the present we propose a new method based on comparison of noise spectra measured on different points of a TLM ([1]), which is in this respect a multi-contact test pattern. According to our new method proposed, the noise data measured on different parts of the TLM pattern will be evaluated by a subsequent numerical calculation, that is capable to separate the noise of the ohmic contact from the complete noise of the test pattern.

2. THE NEW CONTACT NOISE SEPARATION METHOD

During the measurement of the contact noise a low noise voltage source (alkaline dry battery) and a low noise preamplifier (fabricated by Ambrózy as described in [2]) were applied. The noise spectrum was measured in frequency range of 1.6 Hz...20 kHz with real-time digital frequency analyzer type Brüel & Kjær 2131 controlled by PC 486.

The theoretical background of our new method is based on the next reasoning. In a device several noise voltages are combined. Measuring the noise of the device we can measure the combination of these noise voltages. For the measuring of this voltage a simple test lay-out can be used, which is a plain homogeneous resistive layer on the semiconductor wafer with two ohmic contacts. Connecting this device into the measuring circuit, the contacts are flowed through with current. Comparing to the IEC recommendation [3], in the frame of this work a new modified measurement philosophy was used, where in the device under test (DUT) the driving and measuring points are separated.

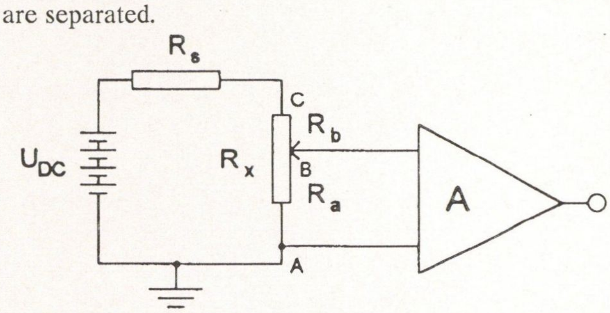


Fig. 1. Proposed measurement setup

Considering a currentless probe applied in the circuit, we shared the resistance of DUT described above (Fig. 1.). The biasing DC current is flowing through the test pattern from the extreme contact to the other extreme contact. The noise signal is measured between the first (extreme) contact and the currentless probe. According to the placement of this probe the obtained noise signals are different. Assuming the noise of the auxiliary parts of the measuring circuit to be zero, the measured noise spectrum originating from the two contacts and the resistive layer is as follows:

$$S = P_c \left[\frac{(R_s + R_b)^2 + (R_a)^2}{(R_s + R_x)^2} \right] + \\ + P_{r0} \left[\frac{R_a(R_s + R_b)^2 + R_b(R_a)^2}{(R_s + R_x)^2} \right], \quad (1)$$

where:
R_s is the separator serial resistance in the circuit, R_x is the resistance of the resistive layer, R_a is the resistance of the part of the resistive layer between points **A** and **B**, R_b is the resistance of the other part of the resistive layer (between **B** and **C** points), P_c is the noise power of the contact flowed through by current, P_{r0} is the noise power of the unit resistance of the resistive layer.

Curves calculated as the function of the ratio R_a/R_x are shown in Figs. 2 and 3. The parameter of the curves is the noise level of P_c or P_{r0}, when the other parameter is assumed to be zero.

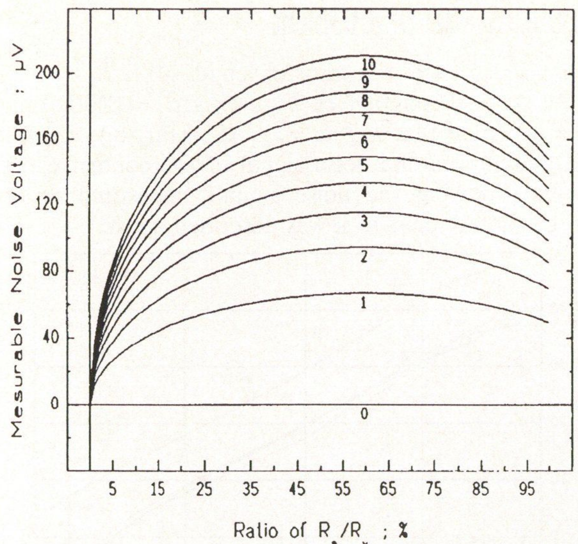


Fig. 2. Resistor noise
(parameter is the resistor noise level P_{r0}, index step is 1 μV^2, contact noise $P_c = 0$)

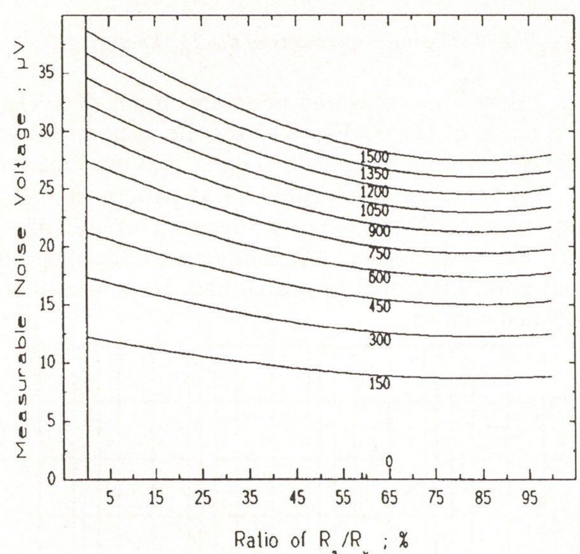


Fig. 3. Contact noise
(parameter is the contact noise level P_c index step is 150 $\mu V^2/k\Omega$ resistor noise $P_{r0} = 0$)

Using this equation the contact noise (P_c) and the noise of a unit resistance (P_{r0}) can be calculated by curve-fitting. This formula is very simple, it does not take any consideration for non-ideal properties of the test pattern.

3. OHMIC CONTACTS AND TLM

In practice a potentiometer-like probe cannot be used because of its high noise level. This contact must be deposited onto the surface as contacts are usually made. Transmission Line Model is a widely used multicontact test pattern (Fig. 4). There is a need of a new formula taking the impact of metal overlayer into consideration.

In this case the biasing DC current is flowing through the TLM pattern from the extreme contact to the other extreme contact. The noise signal is measured between the first (extreme) and an other contact. Fig. 5 shows the equivalent circuit of the noise measurement on the TLM pattern.

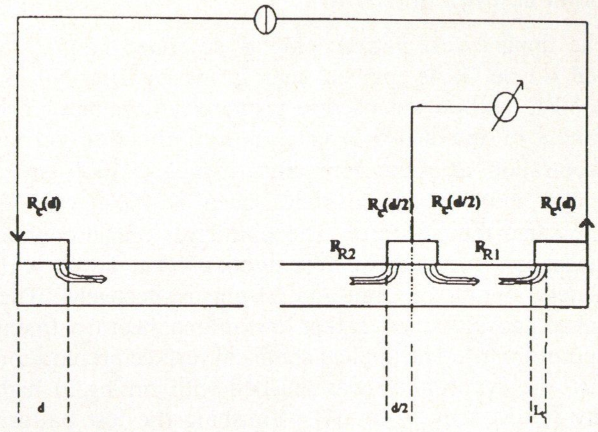


Fig. 4. Cross-section of TLM probe

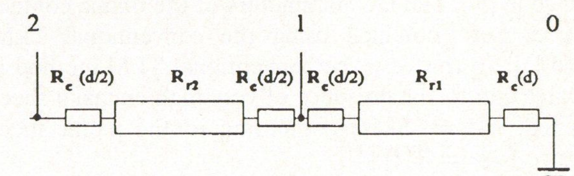


Fig. 5. The equivalent circuit of the TLM pattern

As Figs. 4 and 5 show, there are two different types of contacts: extreme and internal. In practice the internal contact cannot be currentless. When a metal overlayer is deposited onto the semiconductor surface the current flowing through the semiconductor resistive layer it goes in and out the metal. The internal contact can be understood as two serial contacts with half of contact length. Contact resistances can be evaluated [4].

In this point of view the equation for the given noise to the outlet is modified. Assume that R_a is the resistance between contact No. 0 and i, R_x is the resistance from the beginning to the end of TLM structure, u_{c1} is the noise voltage of each extreme contact, u_{c2} is the noise voltage of each internal contact, P_{r0} is the noise voltage of a unit resistance of the resistive layer. The n is the number of internal contacts.

$R_b = R_x - R_a$, and R_s is the serial separator resistance in the measuring circuit. In the case of this model the following noise power can be measured:

Noise power originated from contacts is:

$$S_{c_i} = \left[\left(\frac{R_s + R_b}{R_s + R_x} \right)^2 [P_{c1} + (2i-1)P_{c2}] + \right.$$
$$\left. + \left(\frac{R_a}{R_s + R_x} \right)^2 [P_{c1} + (1 + 2n - 2i)P_{c2}] \right] . \quad (2)$$

Noise power originated resistive layer is (since the noise of R_a resistance is $R_a P_{r0}$):

$$S_r = \left[\left(\frac{R_s + R_b}{R_s + R_x} \right)^2 R_a P_{r0} + \left(\frac{R_a}{R_s + R_x} \right)^2 R_b \right] P_{r0}. \quad (3)$$

Noise powers denoted by S_c and S_r are added to the output.

4. SAMPLE PREPARATION

The applied test pattern chip is described in [5]. S doped n-type GaAs epitaxial layer grown by Effer-Nozaki type VPE method was used to prepare the samples. The thickness of the active layer is 3 μm, the free carrier concentration at room temperature is 1.5×10^{15} cm^{-3} and drift mobility of the active layer is 10000 cm^2/Vs at the same temperature. The multilayer ohmic contact metallization contains AuGe(eutectic)/Ni/Au layers with thickness 75 nm, 12.4 nm and 20 nm, respectively. This metal composition has a rather low optimal heat treatment temperature [6]. The applied chemical surface preparation before the evaporation was finished with rinsing in high purity (18 MΩcm) water [7]. Finishing the test pattern with the preparation of the Schottky gate of FATFET structures, the layer parameters were measured as it was described in [5]. The DC parameters of the ohmic contact structures were obtained using the conventional TLM method [1]. In this case the conventional TLM method is applicable due to the presence of very high epitaxial sheet resistance ($\rho_\square = 5k\Omega$) comparing to the metal sheet resistance ($\rho_M = 1\Omega$) [4].

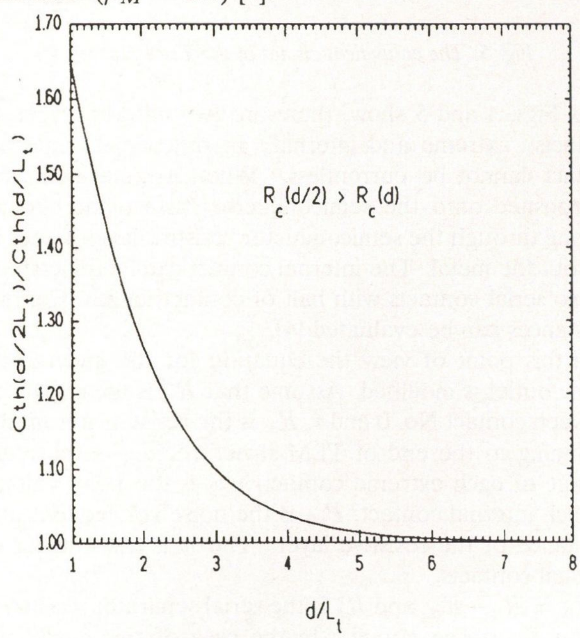


Fig. 6. Contact resistance ratio of extreme and internal vs. contact length comparing to transfer length

The evaluated specific contact resistance was $\rho_C = 1.8 \times 10^{-3}\Omega\text{cm}^2$. Although the real specific contact resistance is a little bit lower as it was proved [4] the obtained value is very high comparing our previous results obtained during MESFET-like device preparation [7]. However, in this case, the free carrier concentration of the applied epitaxial layer is lower with two ranges of magnitude than in the case of MESFET devices. This is why, the optimized contact preparation process results so poor contact. (See Braslau-theory in [8].)

5. RESULTS AND DISCUSSION

Since in our cases the transfer length [4] is $L_t = 6$ μm and the contact length (see Fig. 4) is $d = 100$ μm, it can be accepted that $u_{c1} \approx u_{c2}$. (See Fig. 6.) Applying these assumptions the noise signal of the contact can be evaluated knowing the noise signals, measured on the different contact pairs as it was described above.

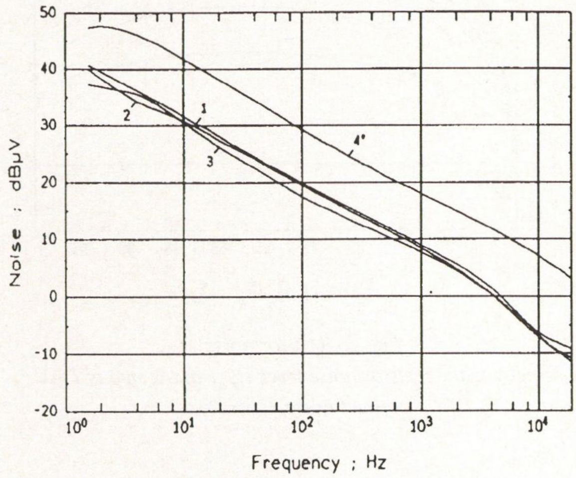


Fig. 7. The noise spectrums of the TLM pattern

Fig. 7 shows the measured noise spectrums of a TLM pattern based of GaAs. Fig. 8 shows the evaluated noise spectra of the ohmic contact. Taking into account that R_C is 300 Ω in the investigated TLM pattern, it should be emphasized, that the noise spectrum of the ohmic contact has a significant 1/f component exceeding the thermal noise attributed to the contact resistance of the investigated contact.

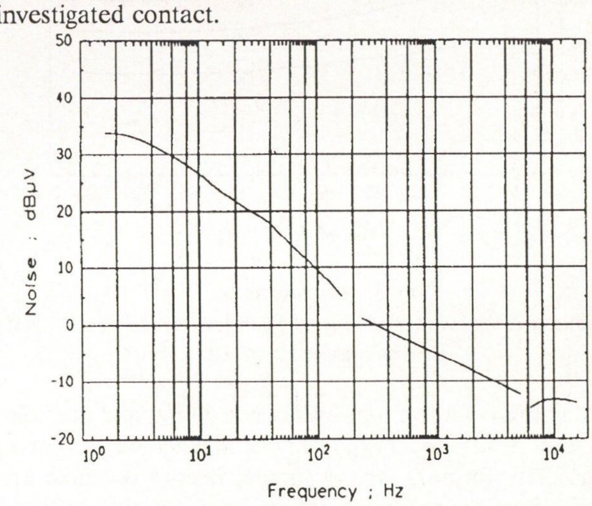


Fig. 8. The separated noise spectra of the ohmic contact

6. SUMMARY

In the frame of this work a new method was introduced for measuring and evaluation of the noise of metal-semiconductor contacts used in semiconductor devices. The measurement is based on a multi-contact test pattern allows to measure the distribution of noise in the function of the localization. A formula is presented which is capable to determine the noise parameters by curve fitting. Our results show that this method is useful in many cases we investigated. The spectrums of frequency distribution were found are nearly $1/f$.

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7. ACKNOWLEDGEMENTS

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The sample, being measured in the course of this investigation, were prepared in the Research Institute for Technical Physics of the Hungarian Academy of Sciences.

Imre Mojzes Photograph and biography on page 1.

Zsolt Kincses Photograph and biography on page 19.

Balázs Kovács received MS degree in electronic engineering at TU Budapest in 1978. He reached Dr. Tech. (Ph.D.) degree, in Electronic and Electrical Engineering from TU Budapest in 1982, Candidate of technical science (C.Sc.) degree, from the Hungarian Academy of Sciences in 1992. He is working for the TU Budapest, as associate professor of the Department of Electronics Technology. His scientific interests are the compound semiconductor technology, the computer aided measurement methods and the reliability assesment of electronic devices.

1/f PHONON NUMBER FLUCTUATION IN PURE LIQUIDS

G. BORBÉLY

DEPARTMENT OF TELECOMMUNICATION
COLLEGE OF ISTVÁN SZÉCHENYI
GYŐR, HUNGARY

These experiments of high purity liquids with Brillouin scattered laser light at the wavelength of 514.5 nm proved that the fractional phonon number fluctuations have $1/f$ type power spectral density. The level of the estimated $1/f$ spectrums at a given frequency in case of nine liquids was found inversely proportional to the number of phonon modes involved in the scattering. An empirical formula for fractional phonon number fluctuations per mode can be given as $S_{\Delta n}(f)/<n>^2 = const \cdot 1/f$, where the constant depends only on the sample.

These results suggest that the origin of the electronical $1/f$ noise and other $1/f$ fluctuations in the field of physics can be related to phonon number fluctuation of $1/f$ type PSD.

The acoustic wave changes locally the distance between the atoms, ions or molecules of material, and consequently the dielectric constant and the refractive index. If the material is transparent, then a small part of the light passed through the material is scattered by fluctuation of refractive index caused by lattice vibrations. The scattering of light by acoustic phonons is the so called Brillouin scattering, because Brillouin gave the first consideration about this effect in 1914 and 1922 [3]. (The light scattering by optical phonons is the well-known Raman scattering.)

The first experimental observation of Brillouin scattering was made by Gross in 1930 [4]. The discovery of laser and its supplement with ultrahigh-resolution Fabry-Perot interferometers accelerated the research of this area. Many investigations were carried out around 1965−70. The experiments were extended for liquids, solids and gases, too.

Usually the frequency spectrum of Brillouin scattered light has three peaks:

- the central, unshifted Rayleigh component,
- two Brillouin peaks with broadened line shape, shifted symmetrically from the incident frequency (f_i) downward and upward with the frequency of the acoustical phonons (f_p).

This frequency is equal to

$$f_p = 2 \cdot f_i \frac{u}{c} \cdot n \cdot \sin(\frac{\gamma}{2}), \qquad (1)$$

because the photon-phonon interaction gives satisfaction to conservation of energy and momentum. Here n is the refractive index of transparent specimen, c is the velocity of light in free space, u is the acoustic velocity in specimen and γ is the angle enclosed by the direction of the incident and scattered light. The frequency value of the acoustical phonons involved in the scattering depends on the scattering angle (the maximum is occurred at back-scattered case). It equals to about 10−50 GHz in solids, 3−6 GHz in liquids and more smaller in gases at the right angle of γ. Since this frequency shift is very small compared to the frequency of the incident light, single mode laser must be used for research.

The first theoretical treatment of the intensity ratio of the Rayleigh and the two Brillouin components was given by Landau and Placzek [5], which was refined by Fabelinskii [6], then Cummins and Gammon [7]. Since the intensity of the central, unshifted Rayleigh component depends also on the purity of the sample, this is the reason why the theory and the experiments give sometimes different results.

A lot of information was collected up to now, but these experiments were the first ones to determine the power spectral density (PSD) of the Brillouin scattered light. The experiments were carried out at the Department of Applied Electronics of Tokyo Institute of Technology, Japan.

A LEXEL Model 95-4 Ar^+-ion laser was used in the experiment as a light source. Its wavelength was 514.5 nm. The laser had 150 MHz longitudinal mode spacing according to 1 meter cavity length. The width of a single longitudinal mode was approximately 3 MHz and the coherence length was about 100 meters due to the temperature controlled intracavity etalon. The maximum laser power at single mode could reach 1200 mW and the fractional power fluctuation was smaller than 0.2%/hour.

The laser light with vertical polarization went through the sample.

For the samples, 14 liquids contained in quartz cell and two synthetic crystalline quartz crystals of different quality were used. (The results with quartz and water can be read in references [1] and [2], respectively.) The liquids were of electronical grade purity and each was carefully forced through Millipore filters of pore size of 0.01 μm in a clean room to remove the very small "dust".

To estimate the frequency spectrum of the scattered light I used a Fabry-Perot interferometer with piezoelectric mirror movement and its overall finesse was about 15. The Fig. 1 shows the frequency spectrum of water, acetic acid, xylene and acetone as typical examples.

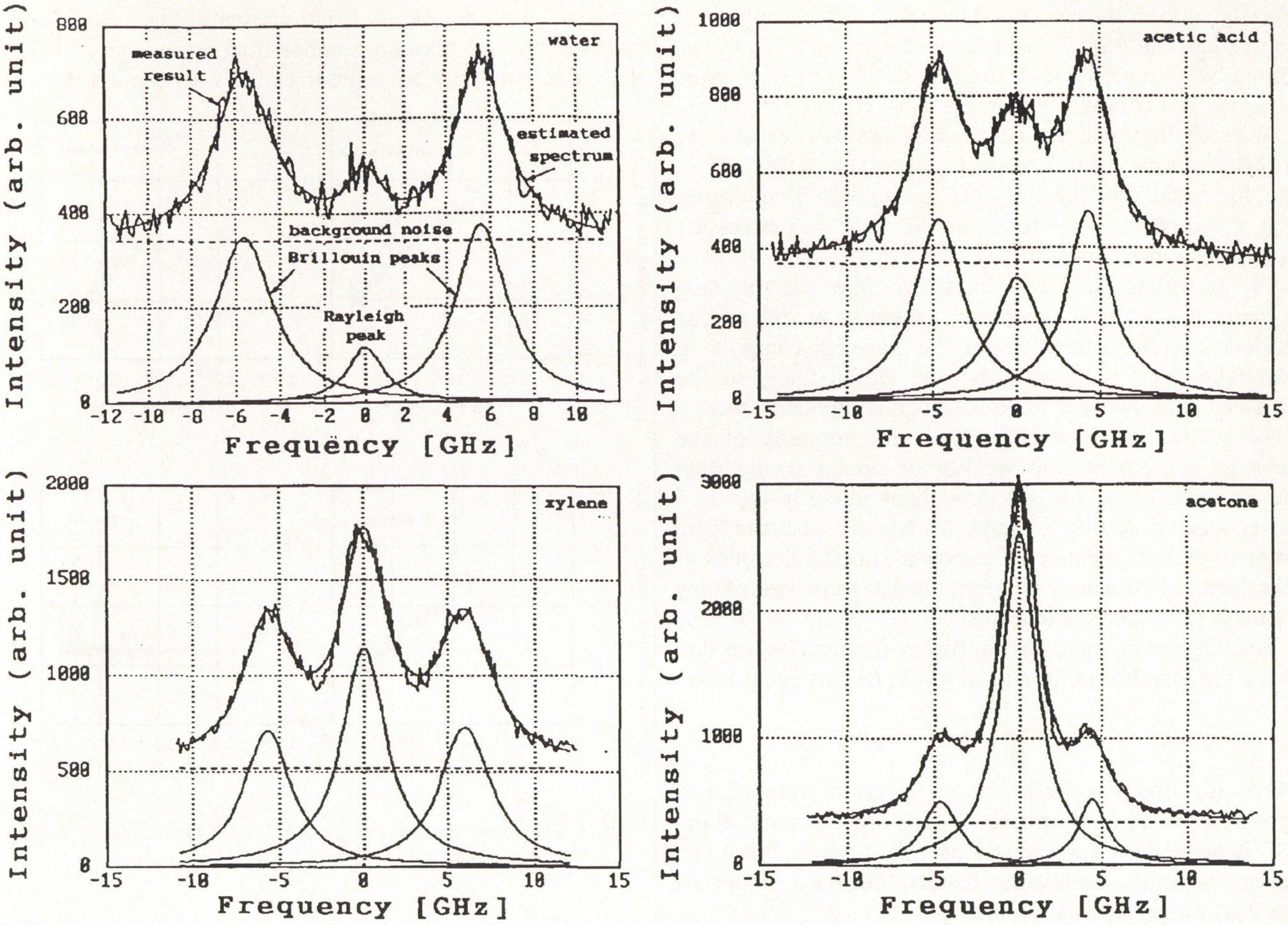

Fig. 1. The Brillouin spectum of water, acetic acid, xylene and acetone

The observed spectrums were approached with three Lorentzian curves by means of the least square method. In this way I could estimate the intensity of the three components. The Fig. 2 shows the intensity of Rayleigh peak in percent of total intensity. It can be said that the intensity ratio of the Rayleigh and the two Brillouin peaks is the smallest for water, the largest for alcohols.

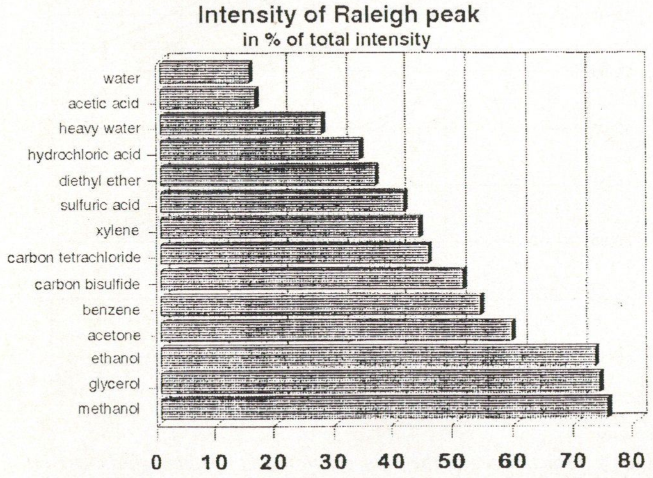

Fig. 2. Intensity of Rayleigh peak in % total scattered light intensity in case of 14 liquids

Since the Brillouin scattered light arises from the interaction of the photons and acoustic phonons, and the scattering cross-section and consequently, the intensity of the scattered light is also proportional to the phonon number, by help of investigation of the scattered light much more can be known about the feature of phonons.

To estimate the PSD of the scattered light I used the so called photon counting technique. The light source was the same as before. The sample was placed to the laser window as near as possible to decrease the angular fluctuation of the outcoming laser light beam. The light scattered in the sample passed through a vertical slit of 0.1 mm in width and 2 mm in height. It was situated 1.5 mm apart from the center of the laser beam.

The light scattered at $90 \pm 8°$ was collected by help of two uncooled photomultipliers Hamamatsu Photonics C716-02. The dark count rate of the photomultipliers was below 100 counts/sec and this value was more than 1000 times weaker than the count rate of the received scattered light.

It must be noted that the PSD-s of the closed photomultipliers were white and the PSD of the background noise of the experimental setup was more than one order of magnitude smaller than the PSD of the scattered light by any samples.

Since the intensity of the incident laser light had a slight fluctuation which appeared in the scattered light, too, I used the ratio of the count number of two photomultipliers to calculate the PSD of the scattered light. In this way the common fluctuation due to the unstable laser power could be eliminated completely from the final results.

The sampling time was adjustable from 0.01 sec to 1 sec in this experiment but usually I had to select the longest one for calculation of PSD at low enough frequencies. This is the reason why minimum 20 hours was necessary for a single observation.

On the other hand this length of time was too long to keep an etalon at the same frequency as one of the incident laser light to filter out the Rayleigh component. Consequently, I could not use the etalon filter, so the scattered light involved some Rayleigh component, too.

As my results show (Fig. 3), if the intensity of the Rayleigh component was weaker or similar strong than the Brillouin ones, the scattered light intensity has $1/f$ power spectral density. I found this type of spectrum from water to carbon bisulfide. The power spectral densities of other liquids (from acetone to methanol) were without any common characteristic features.

The flat part of some PSD at higher frequencies was due to the photon shot noise and its level (L) was equal to

$$L = \frac{1}{\overline{n}_A} + \frac{1}{\overline{n}_B}, \qquad (2)$$

where $\overline{n}_A$ and $\overline{n}_B$ were the mean count rate of two photomultipliers in count/sec. Therefore this part of the PSD depended on the received laser light power, but it can be claimed on the basis of my results, that the $1/f$ part of the PSD was independent on it.

To convince that the $1/f$ parts of the scattered light intensity PSD-s were originated from the phonon number fluctuations of samples, I measured the dependence of the $1/f$ spectral level on the number of the phonon modes involved in the scattering.

The number of phonon modes can be evaluated as

$$N = \frac{2V}{(2\pi)^3}\Omega q^2 dq, \qquad (3)$$

where V is the scattering volume in m^3, Ω is the solid angle determined by the scattered light cone in sterad and q is the phonon wave vector in m^{-1}.

$$dq = 2\pi\frac{B}{u}, \qquad (4)$$

where B equals to the halfwidth of the Brillouin peak and u is the acoustic velocity in the specimen.

Since the overall finesse of my Fabry-Perot interferometer was as low as 15, this experimental setup was not sensitive enough for estimation of dq, so I had to use the value of B from literature [8].

The width of the slit and the diameter of the aperture before the lens were altered to change the value of V and Ω, and together with them to vary the value of N.

The Fig. 4 shows the $1/f$ level at 1 Hz plotted in function of the number of phonon modes. Only these nine liquids had $1/f$ part of their PSD-s. It was found that the levels of the $1/f$ parts of the PSD-s at a given frequency were inversely proportional to the number of phonon modes (N) involved in the laser light scattering. These results verify the phonon origin of the $1/f$ type intensity fluctuations in Brillouin scattering. So it was proved that the phonon number fluctuations have a $1/f$ type PSD and they are not correlated with one another.

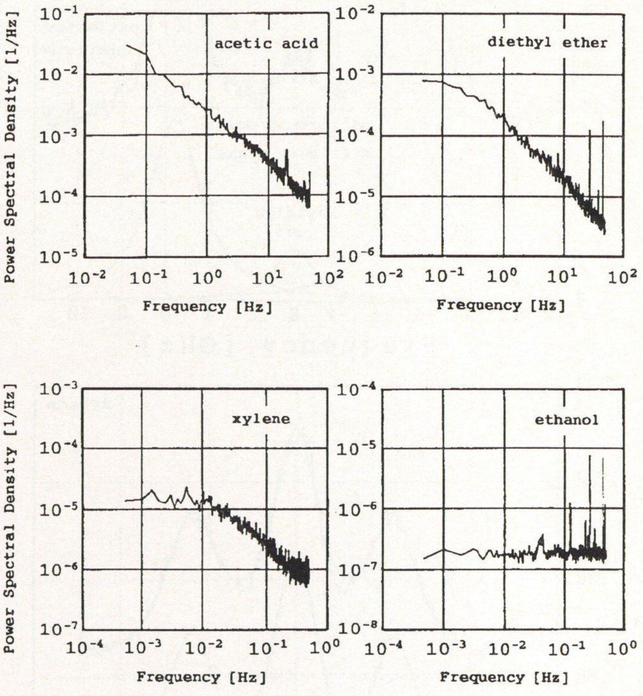

Fig. 3. The power spectral density (PSD) of Brillouin scattered light for acetic acid, diethyl ether, xylene and ethanol

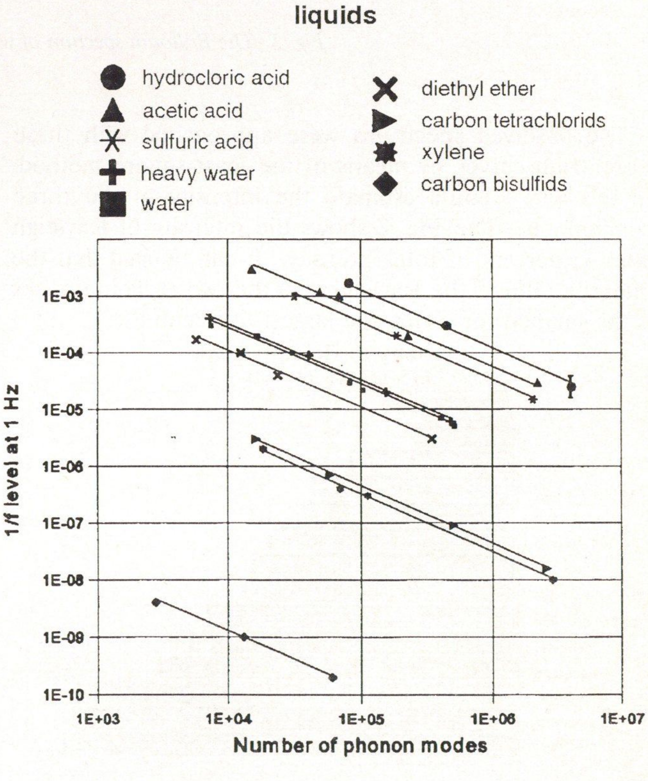

Fig. 4. Dependence of the 1/f spectrum level of nine liquids at 1 Hz on the number of phonon modes involved in the scattering

Relying upon the findings of the Fig. 4 an empirical formula of fractional phonon number fluctuations per mode can be given

$$\frac{S_{\Delta n}(f)}{< n >^2} = \text{const} \cdot \frac{1}{f}, \tag{5}$$

where the constant depends only on the substance of the sample. This value of some liquids can be seen in the following table:

hydrochloric acid	$1.2 \cdot 10^2$
acetic acid	$5.8 \cdot 10^1$
sulfuric acid	$3.4 \cdot 10^1$
heavy water	3.1
water	3.0
diethyl ether	1.1
carbon tetrachloride	$4.4 \cdot 10^{-2}$
xylene	$3.0 \cdot 10^{-2}$
carbon bisulfide	$1.2 \cdot 10^{-5}$

I tried to find a relation between the value of these constants and some invariants of the samples, like density, acoustic speed, viscosity, refractive index etc. Simple connections have not been found. Nevertheless, it can be noted that the liquids of high dielectric constant ($\epsilon_i > 75$), e.g. hydrochloric acid, sulfuric acid, heavy water and water, have high level $1/f$ spectrums and high values of this constant, respectively.

The recognition, that the fractional phonon number fluctuations have $1/f$ type PSD, is a significant advance to explain the origin of $1/f$ fluctuations in the field of electronics and physics. The electronical $1/f$ noises, which are in connection with scattering of electrons by phonons, have phonon origin. The phonon have $1/f$ fluctuations, while the electrons only intermediate these fluctuations in current or voltage noises.

Likewise the $1/f$ noise of Hall-voltage and the thermo e.m.f. can be interpreted. Probably the phonons have a great part in fluctuations of magnetic flux in superconductors, dielectric constant of ferroelectric crystals and decay rate of radioactive materials.

A very slow relaxation process due to the anharmonical connection between the phonons gives rise to longterm fluctuations around thermal equilibrium and to a $1/f$ type fluctuation.

The $1/f$ fluctuations measured in thermal equilibrium of samples lead to contradiction. To solve this problem our current models describing the phonons and the conception about phonon-interaction must be revized.

I would like to thank Professor Toshimitsu Musha for his valuable help and making the experiments possible.

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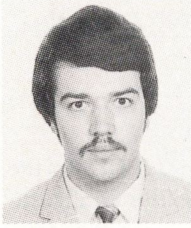

Gábor Borbély studied on the technical-physical branch of the Telecommunication Section within the Department of Electrical Engineering at the Technical University of Budapest. He granted the B.S. Degree with first-class diploma in 1985. He investigated the dry etching methods applied in IC manufacturing technology for two years and granted the M.S. degree with first-class diploma in microelectronics technology at the Department of Electronic Technology at the Technical University of Budapest. In 1988 he defended his thesis in microelectronics as a doctoral dissertation and he was titled with dr. univ. From 1989 he was a visiting researcher at the Graduate School of the Tokyo Institute of Technology, Japan for one and half years. His interest was the 1/f fluctuations in physical systems and the 1/f noise. From 1991 he had got a one-year scholarship from DAAD and researched the electronic noise at the Department of Semiconductor Electronics at the Technical University of Aachen, Germany. In 1995 he took the Ph.D. in microelectronics from the Hungarian Academy of Science. Currently he is an associate professor at the Department of Telecommunication at the College of Technology, Győr, Hungary.

EXTRACTION OF NOISE SOURCES IN BJTS AND HBTS

A. D. VAN RHEENEN
DEPARTMENT OF ELECTRICAL ENGINEERING
11000 UNIVERSITY PARKWAY
PENSACOLA PL 32514

L. R. CHAAR
EE. DEPT. UNIV. OF MINNESOTA
MINNEAPOLIS MN

A. YOUNG
TEXAS INSTUMNETS,
HOUSTON TX

Careful analysis of the noise in the base current and collector current of BJTs and HBTs as a function of the operating current and as a function of the applied feedback resulted in the identification of the noise sources and the magnitude of the base spreading resistance of these devices. In addition, the model we used to analyze the noise data is fully compatible with the noise analysis of the circuit simulator SPICE. The noise performance of silicon BJTs is compared with that of GaAs HBTs.

1. INTRODUCTION

In many analog circuit applications the low frequency noise performance of the devices is an important if not limiting parameter. Ideally, one measures the noise of a single or a few components, constructs a noise model, and uses it in circuit simulators such as SPICE to design circuits and verify their performance without having to build the circuits. To provide the necessary confidence that the model will give reasonable results one would like to base the model on elements that describe the physical operation of the device.

A second motivation for trying to physically locate the sources of noise, is to gain insight into what physical process produces the noise and how device layout, processing material composition, etc. affect the noise.

Third in recent work [1] we showed, as others have suggested earlier [2], that there is a strong correlation between the low-frequency noise of individual transistors and the phase noise of an oscillator based on these transistors. Characterization, modeling, and understanding of the low-frequency noise processes is important to circuits operating in the tens of Ghz range.

In 1986 Van der Ziel et. al [3] proposed a model for the location of the noise sources which is largely based on the current components flowing through the device. They presented a method to estimate the magnitude of the noise sources in a consistent fashion. The important feature of their extraction technique is to make use of feedback, i.e. a resistor in the emitter lead of the transistor. The disadvantage of their model is that it is not fully compatible with the SPICE circuit simulator. We intend to revisit this model and to propose a simple transformation to make the model fully compatible without losing the physical insight. We shall re-evaluate of the data which Pawlikiewicz and Van der Ziel [4] presented as an illustration of the use of their model. We clarify some of the inaccuracies in their interpretation. Finally we present an interpretation of noise data taken on Tektronix AlGaAs/GaAs based HBTs and show that a very simple SPICE model is capable of describing the noise performance, over a range of operating currents, with very few adjustable parameters.

2. THE VAN DER ZIEL MODEL

In this section we shall briefly review the essential features of the model proposed by Van der Ziel et. al. [3]. Fig. 1 defines the model, which is a simple hybrid-π model augmented with three current noise sources: i_{be}, i_{bc}, and i_{ce}. The location of the noise sources is based on current components flowing through the transistor [3]. The base-spreading resistance is split up into two parts r_1 and r_2, and R_E is the resistor between the emitter lead and ground. We use a slightly different notation than did Van der Ziel. Relation between the notations is given in Table 1.

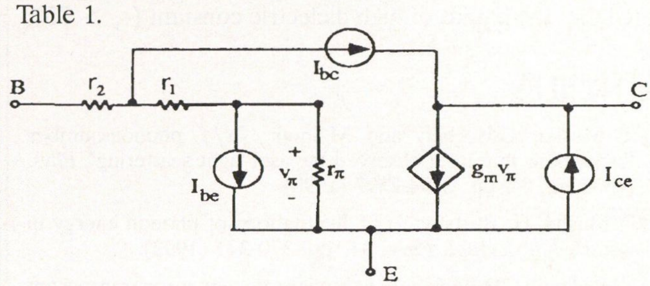

Fig. 1. Model for the bipolar transistor augmented with three noise sources, as proposed by van der Ziel et. al., see [3]

Table 1. Relation between notations in [3] and in this work

This work	Ref. 3	Definition
r_π	$1/g_{be}$	kT/qI_B
g_m	g_{mc}	qI_c/kT
r_e (not used)	$1/g_{me}$	kT/qI_E

Straightforward analysis shows that the shortcircuit base (i_b) and collector (i_c) noise currents are found to be:

$$i_b = \frac{(r_\pi + \beta R_E)i_{be} + [r_\pi + r_1 + (1+\beta)R_E]i_{bc} + R_E i_{ce}}{r_\pi + r_1 + r_2 + (1+\beta)R_E} \tag{1}$$

$$i_c = \frac{\beta(r_1 + r_2 + R_E)i_{be} + [r_\pi + r_i + (1+\beta)(r_2 + R_E)]i_{bc}}{r_\pi + r_1 + r_2 + (1+\beta)R_E} + \\ + \frac{(r_\pi + r_1 + r_2 + R_E)i_{ce}}{r_\pi + r_1 + r_2 + (1+\beta)R_E}. \tag{2}$$

In these equation β is the current gain I_c/I_B. In the limit when R_E tends to zero (common-emitter configuration) we find that

$$i_b = \frac{r_\pi i_{be} + (r_\pi + r_1)i_{be}}{r_\pi + r_1 + r_2} \tag{3}$$

$$i_c = \frac{\beta(r_1+r_2)i_{be} + [r_\pi + r_1 + (1+\beta)r_2]i_{bc}}{r_\pi + r_1 + r_2} + \\ + \frac{(r_\pi + r_1 + r_2)i_{ce}}{r_\pi + r_1 + r_2}. \quad (4)$$

It is possible to reconstruct Eqs. (5a) and (6a) in [3] by letting r_1, $r_2 << r_\pi$ and by assuming that the noise sources i_{be}, i_{bc} and i_{ce} are independent. Under the same conditions one can reconstruct Eqs. (12) and (13) in [3] from Eqs. (1) and (2) above. The noise currents for finite R_E are:

$$i'_b = \frac{(r_\pi + \beta R_E)i_{be} + [r_\pi + (1+\beta)R_E]i_{bc} + R_E i_{ce}}{r_\pi + (1+\beta)R_E} \quad (5)$$

$$i'_c = \frac{\beta R_E i_{be} + (r_\pi + (1+\beta)R_E)i_{bc} + (r_\pi + R_E)i_{ce}}{r_\pi + (1+\beta)R_E}. \quad (6)$$

The purpose of the analysis is to extract the noise source i_{be}, i_{bc} and i_{be} as well as r_1 and r_2 from measurements of the spectra S_{Ib} and S_{Ic}. These two measurements do not provide sufficient information to extract these quantities. Additional measurements S'_{Ib} and S'_{Ic}, with the resistance in the emitter lead are necessary.

3. SPICE MODEL

Fig. 2 shows a slightly modified model, where two base spreading resistance components are lumped into one, $r_b = r_1 + r_2$. The current noise generator (i_{bc}) between the base and the collector is replaced with one (i_z) at the input, parallel to the base spreading resistance. The idea behind this transformation is that one can think of the base spreading resistance as producing thermal noise. Since the circuit model is linear, it is possible to find a linear transformation that relates the new set of sources (i_x, i_y and i_z) to the old ones (i_{be}, i_{bz} and i_{ce}).

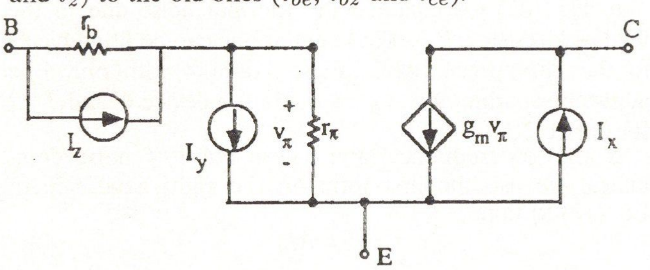

Fig. 2. Slightly modified noise model for the bipolar transistor, compatible with the circuit simulator SPICE

From straightforward circuit analysis we find the short-circuit base and collector currents:

$$i_b = \frac{(r_\pi + \beta R_E)i_y + r_b i_z + R_E i_x}{r_\pi + r_b + (1+\beta)R_E} \quad (7)$$

$$i_c = \frac{\beta(r_b + R_E)i_y - \beta r_b i_z + (r_\pi + r_b + R_E)i_x}{r_\pi + r_b + (1+\beta)R_E}. \quad (8)$$

By comparing Eq. (1) with Eq. (7) and Eq. (2) with Eq. (8), and realizing that they are equal for every value of R_E and every operating point (r_π) we find:

$$i_y = i_{be} \quad (9)$$

$$i_x = i_{ce} \quad (10)$$

$$i_z = (r_2/r_b)i_{bc}. \quad (11)$$

In the model proposed by Van der Ziel et. al. [3] all three sources are allowed to exhibit $1/f$ noise in addition to either shot noise (i_{be}, i_{ce}) or thermal noise (i_{bc}). However, by inspection of Eqs. (1), (2), (7) and (8), i_{ce} (or i_x) will only make a significant contribution when it is much larger than the other sources because the other noise contributions are amplified. In most cases because the contribution of i_{ce} (or i_x) may be neglected. Second, since now i_x and i_y are in series, we may lump all the $1/f$ noise into one source: i_y. We only keep i_z around to account for the thermal noise of the base spreading resistance. This model is essentially the one employed in the circuit simulator SPICE.

We shall think of the noise sources as follows:

$$S_{Iy} = A(I_B)/f + 2qI_B \quad (12)$$

$$S_{Ix} = 2qI_C \quad (13)$$

$$S_{Iz} = 4kT/r_b. \quad (14)$$

In Eqs. (12)–(14) I_B and I_C are the operating base and collector currents, A is a constant which may depend on the operating current. In this description there are only two adjustable parameters: r_b and A. Measurement of the noise as a function of the operating current will determine A. The only adjustable parameter is then r_b.

In the Van der Ziel model one needs to establish r_1 and r_2 as well as the individual strengths of three $1/f$ noise sources. In fact we shall see that it is not possible to uniquely extract the values of r_1 and r_2 when we use Van der Ziel's model.

4. DISCUSSION OF ANALYSIS IN [4]

If one takes Eq. (2) and calculates the spectral intensity in the limit $R_E \to 0$ one finds:

$$S_{Ic} = \frac{\beta^2(r_1+r_2)^2 S_{Ibe} + [r_\pi + r_1 + (1+\beta)r_2)]^2 S_{Ibc}}{(r_\pi + r_1 + r_2)^2} + \\ + \frac{(r_\pi + r_1 + r_2)^2 S_{Ice}}{(r_\pi + r_1 + r_2)^2}. \quad (15)$$

Neglecting r_1 and r_2 with respect to r_π we have:

$$S_{Ic} = g_m^2(r_1+r_2)^2 S_{Ibe} + [1 + ((1+\beta)/\beta)g_m r_2]^2 S_{Ibc} + S_{Ice} \quad (16)$$

which is Eq. (6a) of [3]. This equation should have been applied to interpret the data presented in Fig. 4 of [4]. We have redrawn this figure here and labeled it Fig. 3. When analyzing the high-frequency part of the spectra in that figure, the following equation was used:

$$S_{Ic} = 2qI_c + 4kTr_b g_m^2. \quad (17)$$

This is Eq. (3) in [4]. Equation (17) can be derived from Eq. (16) if the first term in (16) may be neglected ($S_{Ibe} = 2qI_B$) with respect to $S_{Ice} = 2qI_c$, which is reasonable as long as the prefactor $g_m^2(r_1+r_2)^2 << \beta$. Under the operating condition considered here, $I_c = 2$ mA, $g_m = 0,077$ S, $\beta = 170$ this limits the value of $r_b = r_1 + r_2$ to $<< 170\Omega$. Second, it is assumed that the second term in the square brackets of (16) is much larger than 1. This implies that $r_2 >> 1/g_m = 13\Omega$. This last condition is not satisfied in view of the numbers produced in Table 1 of [4]. Third, it is assumed that $S_{Ibc} = 4kT/r_b$, a choice that may not be entirely obvious in view of the splitting of r_b into two parts: r_1.and r_2.

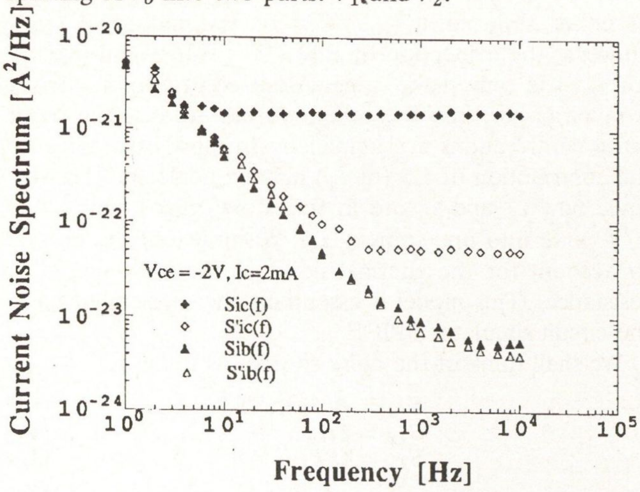

Fig. 3. Current noise spectra (base and collector, with and without feedback) taken from GE-82 pnp bipolar transistor biased at $I_C = 2$ mA and $V_{CE} = -2$ V. See [4] Fig. 4, device 1.

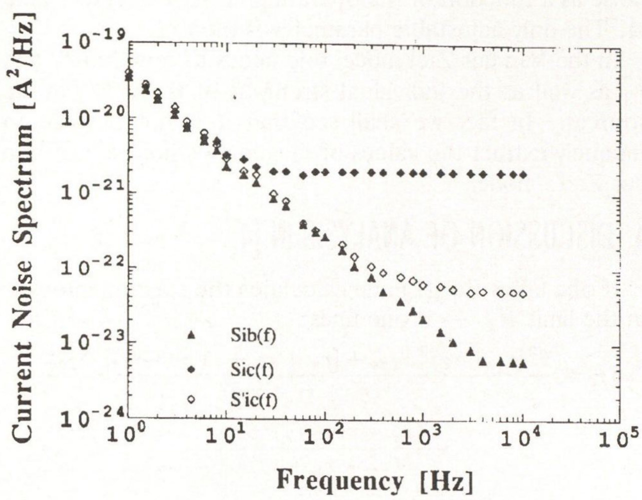

Fig. 4. Current noise spectra (base and collector, with and without feedback) taken from GE-82 pnp bipolar transistor biased at $I_C = 2$ mA and $V_{CE} = -2$ V. See [4] Fig. 6, device 2.

When one assumes $S_{Ice} = 2qI_C$ and $S_{Ibe} = 2qI_B$ and leaves S_{Ibc} as an adjustable parameter, losing all physical significance, one can fit Eq. (16) to the collector current spectrum in Fig. 3. However, S_{Ibc} will now be so large that one can never fit the base current spectra. Exactly similar arguments apply to the interpretation of the data in Fig. 6 in [4], which we have reproduced here and labeled Fig. 4. To overcome these difficulties we propose to apply the model described in Section 2. to analyze the data.

5. REINTERPRETATION OF PAWLIKIEWICZ'S DATA

Without feedback applied ($R_E = 0$) the spectral intensities of the base and collector current noise are (using Eqs. (7) and (8)):

$$S_{Ib} = \frac{r_\pi^2 S_{Iy} + r_b^2 S_{Iz}}{(r_\pi + r_b)^2} \tag{18}$$

and

$$S_{Ic} = \frac{(\beta r_b)^2 S_{Iy} + (\beta r_b)^2 S_{Iz} + (r_\pi + r_b)^2 S_{Ix}}{(r_\pi + r_b)^2}. \tag{19}$$

Under the assumption that r_b may be neglected with respect to r_π we can simplify these equations to:

$$S_{Ib} = S_{Iy} + (r_b/r_\pi)^2 S_{Iz} \tag{20}$$

and

$$S_{Ic} = (g_m r_b)^2 S_{Iy} + (g_m r_b)^2 S_{Iz} + S_{Ix}. \tag{21}$$

With feedback applied ($R_E > 0$):

$$S'_{Ib} = \frac{(r_\pi + \beta R_E)^2 S_{Iy} + r_b^2 S_{Iz} + R_E^2 S_{Ix}}{[r_\pi + r_b + (1+\beta)R_E]^2} \tag{22}$$

$$S'_{Ic} = \frac{\beta^2 (r_b + R_E)^2 S_{Iy} + (\beta r_b)^2 S_{Iz} + (r_\pi + r_b + R_E)^2 S_{Ix}}{[r_\pi + r_b + (1+\beta)R_E]^2}. \tag{23}$$

In the high-frequency limit, where the white noise dominates:

$$S_{Ib} = 2qI_B + (r_b/r_\pi)^2 4kT/r_b, \tag{24}$$

$$S_{Ic} = (g_m r_b)^2 2qI_B + (g_m r_b)^2 4kT/r_b + 2qI_C, \tag{25}$$

$$S'_{Ib} = \frac{(r_\pi + \beta R_E)^2 2qI_B + r_b^2 4kT/r_b + R_E^2 2qI_C}{[r_\pi + r_b + (1+\beta)R_E]^2}, \tag{26}$$

and

$$S'_{Ic} = \frac{\beta^2 (r_b + R_E)^2 2qI_B + (\beta r_b)^2}{[r_\pi + r_b(1+\beta)R_E]^2 4kT/r_b} + \frac{(r_\pi + r_b + R_E)^2 2qI_C}{[r_\pi + r_b(1+\beta) + R_E]^2} + 4kT/R_E. \tag{27}$$

In Eq. (27) we included the thermal noise due to the feedback resistor R_E. These equations can be fitted nicely to the data presented in Figs. 3 and 4 with only one adjustable parameter: $r_b = 6.7\Omega$ for device 1 and 11Ω for device 2.

In the low-frequency limit where the $1/f$ noise dominates, we use the first term on the right hand side of Eq. (12) to find:

$$S_{Ib} = A/f \tag{28}$$

$$S_{Ic} = (g_m r_b)^2 A/f \tag{29}$$

$$S'_{Ib} = \frac{(r_\pi + \beta R_E)^2 A/f}{[r_\pi + r_b + (1+\beta)R_E]^2}, \tag{30}$$

and

$$S'_{Ic} = \frac{\beta^2 (r_b + R_E)^2 A/f}{[r_\pi + r_b + (1+\beta)R_E]^2}. \tag{31}$$

The right-hand sides of Eqs.(28)$-$(31) are sufficiently close to A/f to claim this to be a good fit to the spectra presented in Fig. 3. We want to stress again that only two fitting parameters, r_b and A, have been used to describe the spectra in [4]. One may argue that the right hand side of Eq. 29 (about $0.26A/f$ for device 1 and $0.72A/f$ for device 2) is not close enough to A/f to fit the measured data. This objection may be overcome by including a second $1/f$ noise source, namely in the noise generator S_{Ix}. Let us call that contribution B/f. (28)$-$(31) are now modified to include this source and read:

$$S_{Ib} = A/f, \tag{32}$$

$$S_{Ic} = (g_m r_b)^2 A/f + B/f, \tag{33}$$

$$S'_{Ib} = \frac{(r_\pi + \beta R_E)^2 A/f + R_E^2 B/f}{[r_\pi + r_b + (1+\beta)R_E]^2} \tag{34}$$

and

$$S'_{Ic} = \frac{\beta^2 (r_b + R_E)^2 A/f + (r_\pi + r_b + R_E)^2 B/f}{[r_\pi + r_b + (1+\beta)R_E]^2}. \tag{35}$$

By inspection of Eqs. (34) and (35) we observe that B/f only contributes if $B >> A$, by a factor of about β^2 or 3×10^4 in our case. Fitting Eqs. (32)–(35) to the spectra in Figs. 3 and 4 we find:

Table 2. Strength of the 1/f noise sources extracted from measurements on GE 82 pnp transistors

	A ($10^{-21} A^2$/Hz)	B ($10^{-21} A^2$/Hz)
Device 1:	6	4
Device 2:	40	10

Obviously, the second device is a bit noisier than the first and we find that the largest difference is in the $1/f$ noise produced by the base-emitter junction.

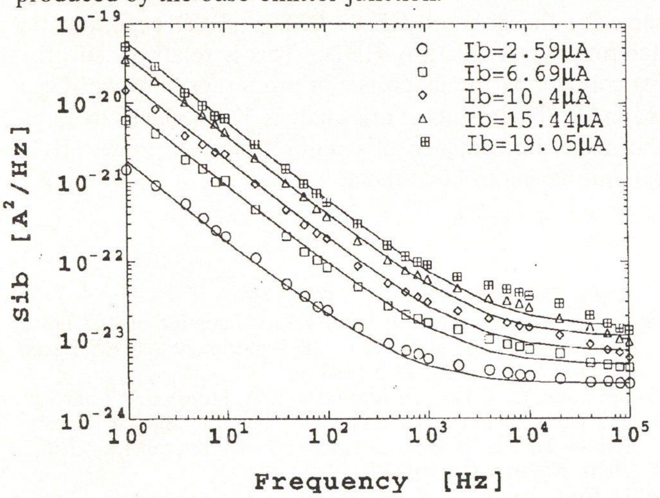

Fig. 5. Base current spectra of Tektronix GaAs/AlGaAs HBT. The symbols are experimental data and the solid lines are SPICE simulations. The collector-emitter voltage is 2 V.

In addition Fig. 7 in [4] (device 2, not reproduced here) indicates that S_{Ic} (1 Hz) scales with I^2. This is possible only when the first term on the r.h.s. of Eq. (33) is dominant and A has a weak dependence on the magnitude of the collector current. Again, the numbers we extracted support this assertion.

Although this fit is better, we now used an additional fitting parameter. A more important consideration is that now our model is not fully compatible with SPICE any more. However, this can be remedied by including an external noise source connected between the collector and emitter in the circuit simulation file.

We have showed how to extract the noise parameters which describe the noise performance of BJTs and applied this to a GE 82 pnp transistor. We found the noise parameters at one operating point only, namely $I_{Ic} = 2$ mA, $V_{Ce} = -2$ V.

In general, one would be interested in the dependence of the noise on the operating point. This knowledge would allow one to choose the operating points of the devices for a minimal circuit noise figure. We shall address this issue in the next section and apply the analysis to GaAs/AlGaAs HBTs.

6. CURRENT DEPENDECE OF THE NOISE AND APPLICATION TO HBTS

In the previous sections we compared noise spectra with and without feedback. Feedback controled the relative weight of the individual noise contributions in the terminal current noise spectra. With the simplifications proposed we show that it is sufficient to measure the base and the collector current noise spectra as a function of the operating current, without feedback. We show that we can extract the current dependence of the magnitude of the noise sources. To demonstrate the procedure we present base and collector current noise spectra taken on Tektronix GaAs/AlGaAs HBTs operating in the forward active mode: $\beta = 60$ with collector current I_C from 0.1 to 1.1 mA. The noise data, taken at room temperature, are presented in Figs. 5 and 6. The symbols represent the experimental data.

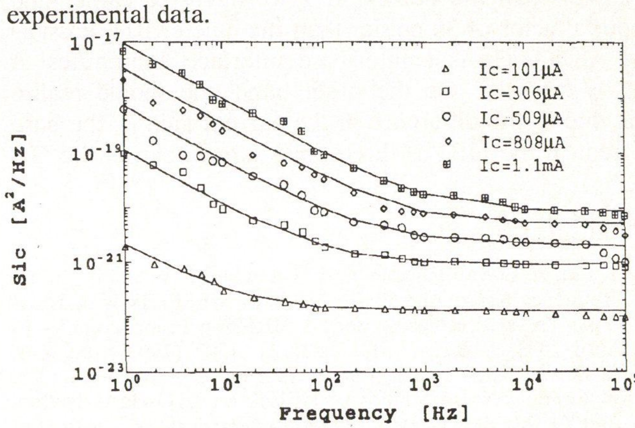

Fig. 6. Collector current spectra of Tektronix GaAs/AlGaAs HBT. The symbols are experimental data and the solid linea are SPICE simulations. The collector-emitter voltage is 2 V.

Let us first focus on the high-frequency part of the collector current spectra (Fig. 6). Comparing these data with Eq. (25) we can solve for r_b. We are not fitting just one spectrum but five spectra, with operating currents ranging over almost a decade, with only one value of r_b, namely 500 Ω. In principle this value could also have been extracted from the base current spectra. However, the base current spectra are less sensitive to r_b. In addition, measurements of the base current spectra suffer from the thermal noise of the base biasing resistor (10 kΩ here) which is comparable to the noise sources intrinsic to the transistor.

Having established the value of r_b we can now proceed to find the strength of the $1/f$ noise source as a function of the operating current (I_B or I_C), either by using the base current spectra and Eq. (28) or by using the collector noise spectra and Eq. (29). Let us concentrate on the base current spectra.

At 1 Hz the intensity of the base current noise is at least two orders of magnitude larger than the white noise, therefore we can safely assume that at 1 Hz the spectra are determined by the $1/f$ noise. A plot of the noise intensity at 1 Hz versus the base current tells us that $A = 7.6 \times 10^{-12} I_B^{1.7} A^2/\mathrm{Hz}$. The error in the prefactor is about 30 % whereas the error in the exponent is about 10 %.

The values for r_b and A have been used to simulate the noise performance of the transistor as a function of the operating current. The results of these simulations are presented in Figs. 5 and 6 by the solid lines. We conclude that this model simulates the experimental data quite well. Only three parameters needed to be established and spectra are fitted over almost a decade of operating currents.

The current exponent of the $1/f$ noise has a value between 1, which a typical of diffusion noise in the base, and 2, which is typical for generation-combination ($g - r$) noise in the space charge region. It is probably not too bold to suggest that indeed the $1/f$ noise is due to $g - r$ processes in the base-emitter space charge region. When we calculate the strength of the $1/f$ noise source at $I_C =$ 2 mA we find $1.9 \times 10^{-19} A^2/\mathrm{Hz}$ or about a factor of 5 noisier than the noisiest BJT we discussed earlier and about a factor of 30 noisier than the quieter BJT. Possibly the AlGaAs/GaAs emitter-base interface contributes to the $1/f$ noise. On the other hand one should realize that due to the difference in the current gain β, the base current in the HBT is three times larger than in the GE 82 BJT. This would imply that at a given base current the noisier BJT and the HBT produce about the same amount of $1/f$ noise in the emitter-base space charge region. We suggest that a suppression of the $g - r$ processes in the base region will reduce the $1/f$ noise of the HBT and of course reduce the base current thus improving the current gain. At that point the low-frequency noise performance of HBTs will challenge that of Si BJTs.

7. CONCLUSIONS

We have demonstrated that we can simply transform an earlier proposed noise model [3] for bipolar transistors into one that is fully compatible with a common circuit simulator: SPICE. The original model was firmly based on the physics of operation of the device. We argued that none of these aspects were lost in the transformation. Second, fewer parameters need to be fitted to describe the noise observed in experiments. We have reevaluated the data presented in [4], basically coming to the same conclusions but in a more rigorous way. We have also interpreted base current noise spectra and collector current noise spectra taken from GaAs/AlGaAs HBTs. The noise spectra could be fitted with three parameters over almost one order of magnitude of operating currents. We found that the strength of the intrinsic $1/f$ noise source, associated with the emitter-base junction, is roughly the same for the Si-based GE-82 BJTs and the experimental Tektronix GaAs/AlGaAs HBTs. This is relatively surprising considering the difference in properties between these materials. The results of our analysis give a means to study the noise performance of circuits based on these HBTs without having to build them.

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A D. van Rheenenreceived the Ph.D. in Electrical Engineering in 1987 from the University of Florida in Gainesville. At that time he joined the faculty at the University of Minnesota. Currently he is an associate professor in Electrical Engineering at the University of West Florida. van Rheenen has published extensively on electrical noise issues in electronic devices: generation-recombination noise, velocity fluctuation noise, 1/f noise, correlation between phase noise in oscillators and low-frequency noise in the components, correlation between the noise in the light intensity of semiconductor diode lasers and noise in the pump current. Currently, he is interested in circuit level modelling of noise in active devices.

L. R. Chaar's and **A. Young**'s photographs and biographies were not available at the time of printing.

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