ACCURATE LOW-FREQUENCY TRANSISTOR NOISE MEASURING INSTRUMENT WITH DIGITAL DISPLAY OF LOGARITHMIC NOISE FIGURE*

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(Received May 13, 1971)

The accurate measurement of low frequency, relatively wideband noise involves several difficulties [1, 2, 3]. To achieve the required accuracy, a quite long measuring time is needed. The most important parameters of the measuring system, like gain, bandwidth, detector characteristics etc. must not change — at least during one measurement. This latter requirement involves no difficulties; however, similar long-term stability, independence on the device parameters to be measured and so on would increase the complexity and costs.

For the above reasons the wideband audio frequency transistor noise measuring system to be discussed is in principle somewhat similar to double integrating digital voltmeters. The input of the transistor under test is connected to the specified source resistance. The amplified output noise voltage is integrated by an analogue integrator during a precisely preset time interval. After this the input noise power is increased by a known amount and the output voltage now discharges the integrator. The discharge time is related unambiguously to the noise figure (Fig. 1).

The increment of input noise power should be stable and precisely known both in amplitude and frequency spectrum. Gaussian white noise is preferred and flicker components should be excluded. Among the possible standard noise sources only two types meet these requirements: heated-wire resistor [4] or digital random/pseudorandom synthetizer with low-pass filtering [5—8]. The latter was abandoned because of its inherent complexity and of cost considerations. On the other hand, the known problems arising at the construction of a high frequency hot noise source are mostly dropped since the stray reactances are negligible in the audio-frequency band. Therefore an improved version of a hot noise source developed formerly partly by the author [9] was chosen. The requirements for the temperature control are discussed in the Appendix.

* On the basis of lectures given by the author at TH Aachen and LETI Leningrad.

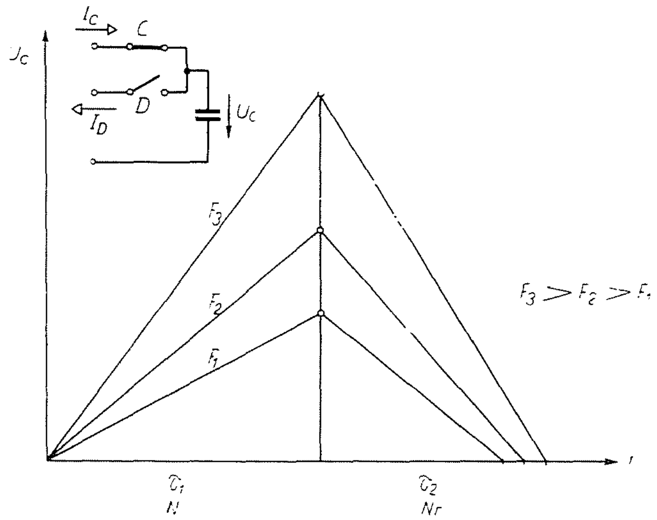

Fig. 1. Charging and discharging curves of the integrator

Principle of noise figure measurement

Fig. 2 shows the cold and hot noise sources and the transistor to be measured. Suppose that the squared noise voltage accross the cold source resistor is u_r^2 at room temperature. Then the equivalent squared noise voltage referred to the input of the transistor of the noise figure F is

$$u_1^2 = F u_r^2 \tag{1}$$

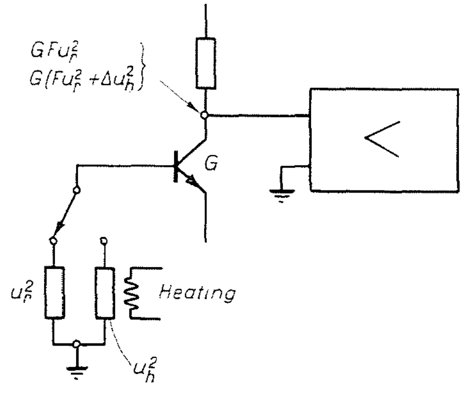

Fig. 2. Room-temperature and hot noise source in the input circuit of the tested transistor. The equivalent input square noise voltage can be obtained by dividing the output by the power gain G

Replacing the room temperature resistor for a hot one, the increased input is

$$u_2^2 = Fu_r^2 + \Delta u_h^2 \tag{2}$$

where

$$\Delta u_h^2 = u_h^2 - u_r^2 = u_r^2\left(\frac{T_h}{T_r} - 1\right) = Au_r^2 . \tag{3}$$

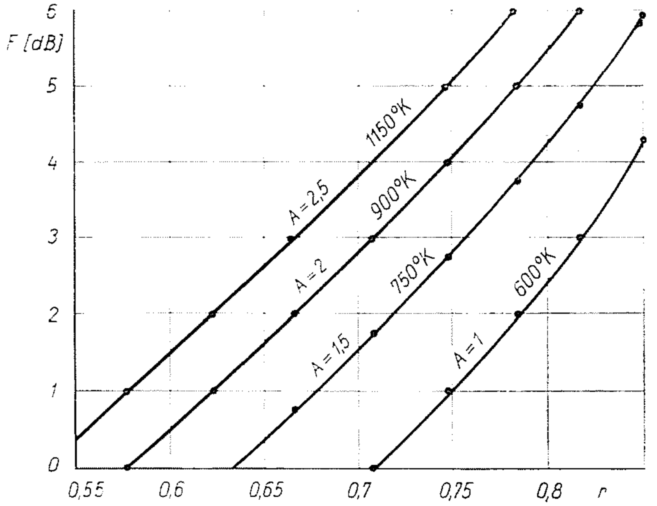

Fig. 3. $F(r)$ curves

Solving (1)—(3) for F and introducing the ratio

$$r = \frac{u_1}{u_2} \tag{4}$$

the equation

$$F^{dB} = 10 \log \frac{Ar^2}{1 - r^2} \tag{5}$$

results. Fig. 3 shows the $F^{dB}(r)$ relationships with four different A corresponding to different hot source temperatures. The lower the hot source temperature, the more nonlinear the function becomes. $A = 2$ or $T_h = 900$ °K is a good compromise. In that case the slope changes as 4 to 3 in the range of 0...6 dB.

Returning to Fig. 1 it appears clearly that for r neither u_1 nor u_2 should be measured, since

$$r = \frac{\tau_2}{\tau_1} \tag{6}$$

is also valid. Keeping τ_1 constant, τ_2 is almost proportional to the logarithmic noise figure. So the noise figure measurement may be carried out by counting elementary time intervals. Their most suitable length and number is calculated as shown in the next paragraph.

Design considerations

Fig. 1 shows the simplified functional diagram of the integrator and the voltage-time relationship. In fact, the output voltage of the integrator approximates the straight lines shown only stochastically. If 1% uncertainty is permitted with a confidence of 99.7%, the minimum charging time is [10]:

$$\tau_{\min} = \frac{f(p)}{\varepsilon^2} \frac{1}{B} = \frac{9}{10^{-4}} \frac{1}{15.10^3} = 6\,s \tag{7}$$

considering 15 kHz bandwidth, which covers the whole audio-frequency band.

The most economical way for measuring such time intervals is the use of counting decades driven with 100 Hz clock frequency, derived from the power line. The short term stability of co-operating power systems is better than 10^{-3}. Three counting decades provide maximum 10 s charging time with 10 ms resolution.

Referring to Fig. 1, the charging time is constant, controlled by the counter. The same counter measures the discharging time too, which depends on the noise figure of the transistor to be measured. Considering in addition the nonlinear relationship between F and r, what is the best choice for N (the number of elementary time intervals during the charging)?

The noise figure has to be displayed digitally in steps of 0.1 dB, therefore each step should correspond to an integer increment in rN, that is

$$N \Delta r = k = \text{integer}. \tag{8}$$

By definition

$$\Delta F = \frac{\mathrm{d}F}{\mathrm{d}r} \Delta r \tag{9}$$

and the initial slope derived from (5) and corrected for minimum error is

$$\left.\frac{\mathrm{d}F}{\mathrm{d}r}\right|_{F=0} = 22.95\,\mathrm{dB} \tag{10}$$

referred to the unity increment of r.

Combining the above equations and choosing $k = 4$:

$$N = \frac{k}{\Delta r} = \frac{k}{\Delta F} \frac{\mathrm{d}F}{\mathrm{d}r} = \frac{4}{0.1}\, 22.95 = 918. \tag{11}$$

It means that in the first measuring cycle the three counting decades are well utilized; the charging time (9.18 s) is sufficiently longer than the minimum calculated by Eq. (7).

The choice of $k = 4$ has another advantage: the end slope of the $F(r)$ curve is 4/3 times the initial one, so at the upper end the step of 0.1 dB corresponds to $k = 3$. In the middle range $k = 3$ or 4 provides everywhere the proper approximation. In fact, the $F(r)$ curve is approximated by a best fit polygon.

Table 1 shows numerically the nonlinearity to be linearized digitally.

Table 1

F	r	Nr	$N\Delta r$
0	0.5774	530	—
1	0.6216	570	40
2	0.6649	610	40
3	0.7067	649	39
4	0.7461	685	36
5	0.7827	719	34
6	0.8158	749	30

Sequential and block diagram

Fig. 4 shows the block diagram of the instrument and Fig. 5 the operation sequence. Fig. 4 does not include all units of the measuring equipment; parts

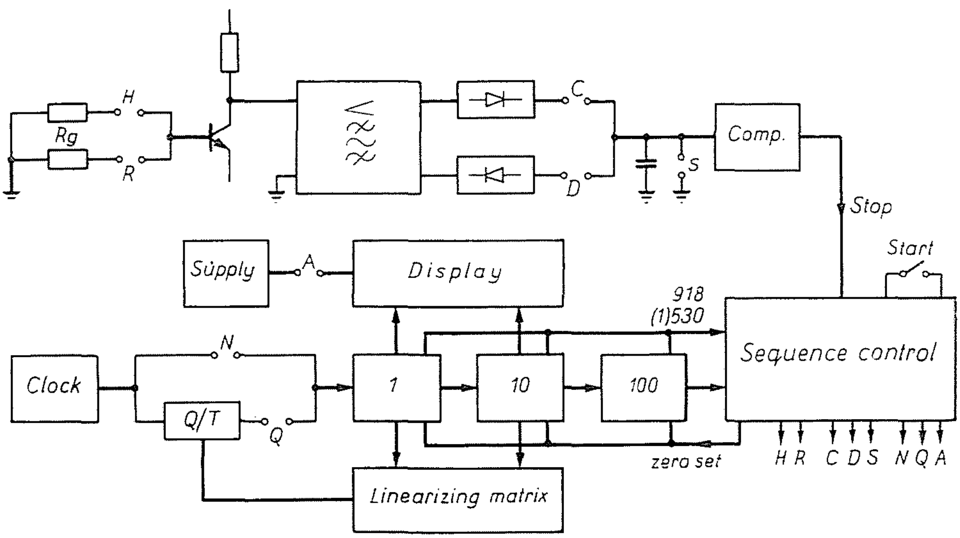

Fig. 4. Block diagram of the measuring instrument

urimportant from the point of view of operation sequence are omitted. The transistor to be measured is connected to the room temperature source resistor or to the hot one, by the switches R and H, respectively. C and D connect the charging or discharging circuit to the integrator while S serves as short circuit between the stop and the next start in order to avoid the integration of spurious signals. The comparator senses the zero-crossing of the discharging waveform, see Fig. 1.

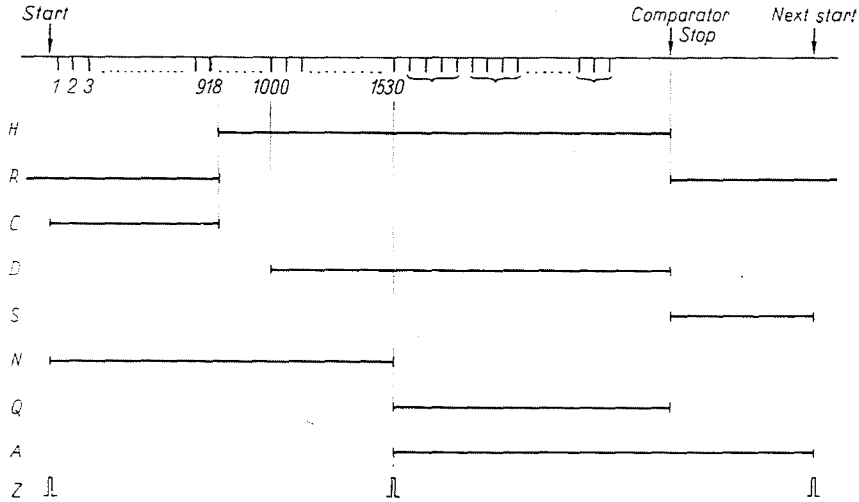

Fig. 5. Operation sequence of the switches shown in Fig. 4

The clock may be connected directly to the counter via switch N or via the quartering-thirding (Q/T) circuit. This circuit passes regularly every fourth pulse: the linearizing matrix, however, can instruct it to pass every third one, for the sake of the best tracking of the nonlinear $F(r)$ curve.

The counter is double-utilized. It controls the charging period and continues the counting after this, even during the first part of the discharge. After counting down 530 pulses (which correspond to 0 dB, see Table 1), the quartering-thirding circuit is inserted and the counter controls directly the display.

Fig. 5 shows the operation sequence. Before the start the room temperature source resistor is connected to the transistor to be measured. The start command initiates the 9.18 s long charging period. At the end of this, R and C open and H closes. The switching of the source resistors may cause transients, therefore the discharge period begins only 0.82 s later. This means that in fact the waveform shown in Fig. 1 is rather a trapezoid than a triangle.

5.3 s later N opens, the counter is reset by Z. Q closes and the display (Nixie tubes) is activated by A. From now, every four (or three) clock pulses increase the displayed value by 0.1 dB. The unity counter drives the tenths of decibels, the ten counter the decibels from 0 to 5, thus the maximum value is 5.9 dB. This range and accuracy is sufficient for testing modern silicon transistors.

Finally the comparator stops the whole process which takes all in all 17 s in average. The display remains active to the next start command.

Conclusion

Accurate but as rapid as possible low frequency noise measurement raises complex problems. The use of primary noise standard results in non-linear system equation. This problem may be overcome by using digital linearizing technique. It is believed that this principle may have more general uses.

Acknowledgement

G. HIDAS and A. KARDOS contributed to this project with many useful remarks. The permission of publication is acknowledged to the Industrial Research Institute for Electronics "HIKI".

Appendix

One of the main factors influencing the accuracy is the hot to room temperature ratio T_h/T_r. The temperature of the hot source can be regulated using a sensor element (R_h) connected into a Wheatstone bridge which controls the heating power of the oven. Contrary to the usual, the reference arm of the bridge also consists of a temperature sensing resistor (R_r) instead of a fixed value one. This tracks the room temperature.

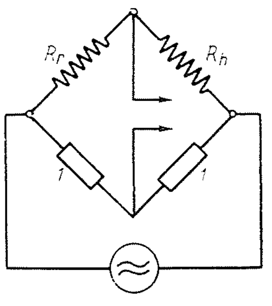

Fig. 6. Temperature sensing bridge

As a first approximation let us suppose that both resistors have linear $R(T)$ curves. Then

$$R_r = R_1[1 + \alpha_1(T_r - T_0)] \tag{A.1}$$

$$R_h = R_2[1 + \alpha_2(T_h - T_0)] \tag{A.2}$$

where R_1, R_2 are nominal values at T_0, α_1 and α_2 are temperature coefficients. According to Fig. 6 in equilibrium $R_r = R_h$ and

$$\frac{R_2}{R_1} = \frac{1 + \alpha_1(T_r - T_0)}{1 + \alpha_2(T_h - T_0)} \tag{A.3}$$

When T_r changes by ΔT_r, from (A.1) it follows that

$$\Delta R_r = R_1 \alpha_1 \Delta T_r \tag{A.4}$$

and similarly

$$\Delta R_h = R_2 \alpha_2 \Delta T_h \tag{A.5}$$

Supposing again an equilibrium, $\Delta R_r = \Delta R_h$ and

$$\frac{\alpha_1}{\alpha_2} = \frac{R_2}{R_1} \frac{\Delta T_h}{\Delta T_r} \tag{A.6}$$

results.

To fulfil the condition $T_h/T_r = \text{const.}$

$$\frac{\Delta T_h}{\Delta T_r} = \frac{T_h}{T_r} \tag{A.7}$$

is needed. Combining (A.3), (A.6) and (A.8) we get

$$\frac{\alpha_1}{\alpha_2} = \frac{T_h}{T_r} \frac{1 + \alpha_1(T_r - T_0)}{1 + \alpha_2(T_h - T_0)} \tag{A.8}$$

from which α_1 and consequently the resistor material to be used for R_r may be determined.

For estimating the order of magnitude of α_1 let $T_0 = T_r$ and suppose that the $R(T)$ curves, considered already linear, start at the origin of the $R - T$ plane, i.e. $R = 0$ when $T = 0$. Then $\alpha_2 = 1/T_1 = 1/T_r$ and using (A.8)

$$\alpha_1 = \frac{1}{T_r} = \alpha_2 \tag{A.9}$$

results. Therefore the same idealized material should be used in both arms of the Wheatstone-bridge and $R_2/R_1 = T_r/T_h$.

In fact, the $R(T)$ relationship is rather of the form

$$R = R_0[1 + \alpha(T - T_0) + \beta(T - T_0)^2 + \ldots] \tag{A.10}$$

where $\beta < 0$ and the higher order terms generally may be neglected. In this case

$$\frac{R_2}{R_1} < \frac{T_r}{T_h} \tag{A.11}$$

and in (A.4)...(A.6) $\mathrm{d}R/\mathrm{d}T$ should be used instead of alphas. Combining (A.6) and (A.7) we finally get

$$\left.\frac{\mathrm{d}R}{\mathrm{d}T}\right|_{T_r} = \frac{R_2}{R_1}\frac{T_h}{T_r}\left.\frac{\mathrm{d}R_h}{\mathrm{d}T}\right|_{T_h}. \tag{A.12}$$

Since $R_2T_h/R_1T_r < 1$, the room temperature resistor may have smaller temperature coefficient than the hot one. This can be realized when the hot sensor is made from platinum and the cold one partly from platinum, partly from a temperature invariant resistance material.

Summary

The low frequency, relatively wideband, accurate noise measurement of electronic devices involves several difficulties. A primary noise standard should be used: however, it results in nonlinear system equation. In addition, the display of the logarithmic noise figure has been needed for practical reasons. These nonlinearities can be compensated using a digital linearizing technique. The logarithmic noise figure between 0 and 5.9 dB is displayed in 0.1 dB steps. The overall accuracy is of the order of 0.1 dB.

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