NOISE IN DIFFERENTIAL AMPLIFIERS

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1. Introduction

Due to the revolutionary spreading of integrated technology, in the last few years, bipolar transistors and field effect transistors are utilized more and more in differential amplifiers. In operational amplifiers the excellent DC-properties of the symmetrical configuration, while in tuned amplifiers the low feedback of the differential amplifier are utilized. In both fields of application, differential amplifiers are also used as input stages. This fact necessitates to investigate the noise problems of differential amplifiers, taking the special configurations developed in integrated circuits into consideration.

2. Symbols

$\overline{e_b^2}$	$r_{bb'}$ resistance noise
$\overline{e_e^2}$	shot noise of the emitter circuit
β	common emitter small signal current gain
q	electrone charge
k	Boltzmann constant
T	absolute temperature
Δf	actual differential bandwidth
I_e	continuous (d. c.) emitter current
i_e	the alternating component of continuous (d. c.) emitter current
I_{CB0}	inverse collector-base current
$\overline{e_s^2}$	substrate-collector conductance noise
$\overline{e_c^2}$	collector bulk resistance noise
$\overline{i_b^2}$	current distribution noise
$\overline{i_{cb0}^2}$	saturation current noise
α	common base dynamic current gain
A	common base static large signal current gain
f_α	cut-off frequency of common base dynamic signal current gain

3. Transistor noise

The physical noise-equivalent model of a bipolar transistor is well-known from e.g. [1], [2], [3], [4], [5]. In case of planar technology transistors used in integrated circuits this model is modified by the presence of the substrate and a high, collector-side series resistance [1], [5].

Starting from previous results and from the equivalent circuit model in Fig. 1, the noise of bipolar transistor differential amplifiers had been investigated.

$$\overline{e_b^2} = 4kTr_{bb'}\,\Delta f \qquad \overline{i_b^2} = 2q(A - |\alpha|^2)I_e\,\Delta f$$

$$\overline{e_e^2} = 4kT\frac{r_e}{2}\Delta f \qquad i_{cb0}^2 = 2q\,I_{CB0}\,\Delta f$$

$$\overline{e_c^2} = 4kT\,r_c\,\Delta f \qquad \alpha = \frac{\alpha_0}{1 + j\dfrac{f}{f\alpha}}$$

$$\overline{e_s^2} = 4kT\,r_s\,\Delta f \qquad f_\alpha \cong \frac{1}{2\pi\,r_e\,C_c}. \tag{1}$$

The equivalent circuit model gives no information concerning the behaviour in the flicker range, this will be discussed later on.

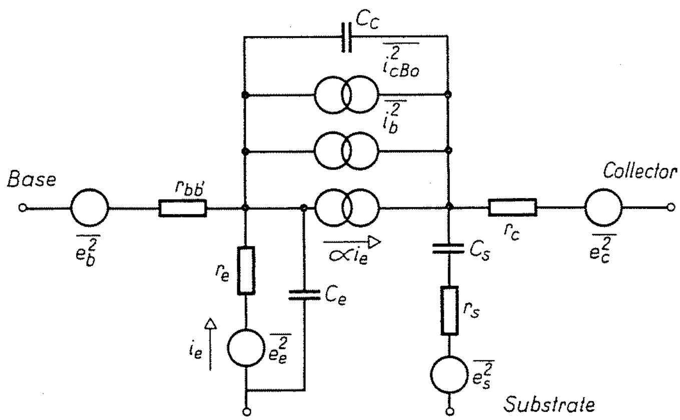

Fig. 1

In case the noise of the differential amplifier is investigated on basis of the complete equivalent circuit model of the transistor, a badly organized, complex relation results. It is thus expedient to simplify the equivalent circuit model in Fig. 1, so that only the really important effects should come to light.

Approximations:

1. $f \ll f_a$, thus $\dfrac{1}{2\pi fC_e} \gg r_e$

2. $\dfrac{1}{2\pi fC_s} \gg |Z_t|$, where Z_t is the collector-side loading impedance

3. $r_c \ll Re\,[Z_t]$ (2)

The resulting simplified equivalent circuit model is transformed in a way that the transistor noise is characterized by noise-current and noise-voltage sources reduced to input and by correlation factor interpreted between them [3]. Due to the difficulties to handle the correlation factor, the current and voltage generators will be divided into totally correlated and not correlated members. The resulting equivalent circuit model T is seen in Fig. 2.

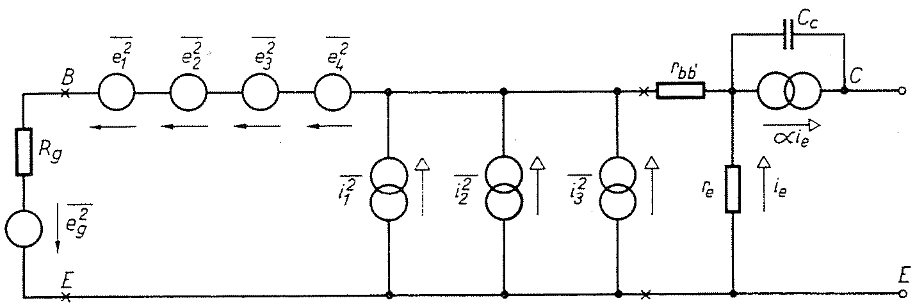

Fig. 2

Using symbols in Fig. 1 and (1), the individual generators are:

$$\overline{e_1^2} = \overline{e_e^2}\left|\frac{\alpha + j\omega\, r_{bb'}\, C_c}{\alpha - j\omega\, r_e\, C_c}\right|^2 \cong \overline{e_e^2}$$

$$\overline{e_2^2} = \overline{i_b^2}\left|\frac{r_e + r_{bb'}}{\alpha - j\omega\, r_e\, C_c}\right|^2 \cong \overline{i_b^2}\,\frac{(r_e + r_{bb'})^2}{|\alpha|^2}$$

$$\overline{e_3^2} = \overline{i_{CB0}^2}\left|\frac{r_e + r_{bb'}}{\alpha - j\omega\, r_e\, C_c}\right|^2 \cong \overline{i_{CB0}^2}\,\frac{(r_e + r_{bb'})^2}{|\alpha|^2}$$

$$\overline{e_4^2} = \overline{e_b^2}\left|\frac{\alpha - j\omega\, r_e\, C_c}{\alpha - j\omega\, r_e\, C_c}\right|^2 = \overline{e_b^2}$$

$$\overline{i_1^2} = \overline{e_e^2}\left|\frac{j\omega C_c}{\alpha - j\omega\, r_e\, C_c}\right|^2 \cong 0$$

$$\overline{i_2^2} = \overline{i_b^2}\left|\frac{1}{\alpha - j\omega\, r_e\, C_c}\right|^2 \cong \frac{\overline{i_b^2}}{|\alpha|^2}$$

$$\overline{i_3^2} = \overline{i_{CB0}^2}\left|\frac{1}{\alpha - j\omega\, r_e\, C_c}\right|^2 = \frac{\overline{i_{CB0}^2}}{|\alpha|^2}\,. \tag{3}$$

The similar suffix current and voltage sources are totally correlated, hence their instant values change proportionally to each other maybe with a time lag. When writing up the equations the approximation $r_{bb'}$; $r_e < \alpha/\omega C_c$ was applied.

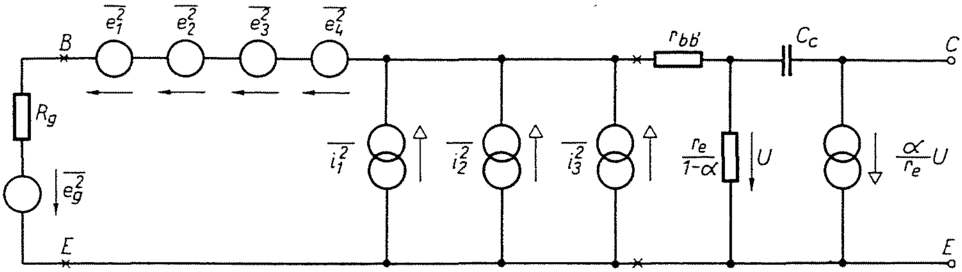

Fig. 3

The transistor of the equivalent circuit model T in fig. 2 will be considered below as noiseless and describable by any equivalent circuit model; from the point of noise characterization these are equivalent. As an example, the equivalent circuit model π is shown in Fig. 3.

4. The noise of constant current sources

Prior to the noise investigation of differential amplifiers, the quantity of noise from the current generating device or stage of a differential transistor pair is to be determined. The usual constant current source solutions are seen in Fig. 4. The noise investigation is restricted to a medium frequency range (above the flicker range but well below the cut-off frequencies), neglecting also the effect of the feedback capacitance; thus the results become comprehensive.

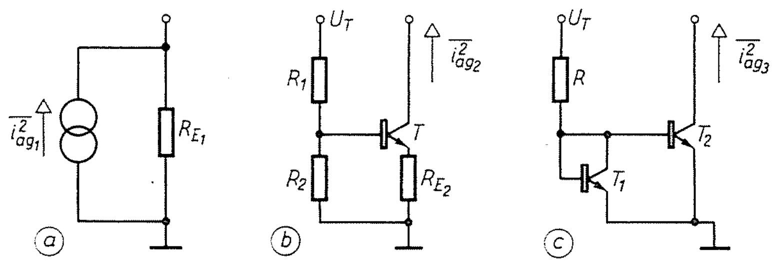

Fig. 4

In case of a simple resistance the evolving noise current can be indicated without difficulty (see Fig. 4a):

$$\overline{i_{ag1}^2} = \frac{4kT\,\Delta f}{R_{E1}}. \tag{4}$$

In case the current source is a transistor stage (see Fig. 4b), the resultant collector noise current can be determined on basis of Fig. 2. Neglecting the noise of the saturation current:

$$i_{ag2}^2 = 4kT\,\Delta f \frac{\alpha^2}{[(R_{E2}+r_e)+(1-\alpha)(r_{bb'}+R_B)]^2}\cdot$$
$$\cdot\left[R_{E2}+r_{bb'}+R_B+\frac{r_e}{2}+\frac{(R_B+R_{E2}+r_e+r_{bb'})^2}{2\beta r_e}\right], \tag{5}$$

where $R_B = R_1 \times R_2$.

By reducing R_B, the noise current diminishes, but a practical limit is imposed by the increase in base voltage divider current. In a given band the noise behaviour can be improved also by capacitively short circuiting the base voltage divider. If R_B is low and also the condition $R_{E2} \gg r_e$; $r_{bb'}\,|1-\alpha|$ is met, then:

$$i_{ag2}^2 \cong 4kT\,\Delta f\,\alpha^2\left(\frac{1}{R_{E2}}+\frac{1}{2\beta r_e}\right). \tag{6}$$

The resultant noise current can be reduced by increasing the resistance R_{E2} and applying a higher current gain factor transistor, at the given (d.c.) operating point.

For a circuit diagram as in Fig. 4c, the value of the resultant noise current is, in general, rather intricate to determine, therefore it is initially supposed that $R \gg r_{e1}$; r_{e2}; $r_{bb'}$ and $r_{bb'1} = r_{bb'2} = r_{bb'}$ and thus

$$i_{ag2}^2 = 4kT\,\Delta f \frac{\alpha_2^2}{\left[1-\alpha_2+\dfrac{r_{e2}+(1-\alpha_2)\,r_{bb'}}{r_{e1}+(1-\alpha_1)\,r_{bb'}}\right]^2}\left\{\frac{2r_{bb'}+\dfrac{r_{e1}+r_{e2}}{2}}{[r_{e1}+r_{bb'}(1-\alpha_1)]^2}+\right.$$
$$\left.+\left[1+\frac{r_{e2}+r_{bb'}}{r_{e1}+r_{bb'}(1-\alpha_1)}\right]^2\frac{1}{2\beta_2\,r_{e2}}+\left[1+\frac{r_{e1}+r_{bb'}}{r_{e1}+r_{bb'}(1-\alpha_1)}\right]^2\frac{1}{2\beta_1\,r_{e1}}\right\}. \tag{7}$$

If the (d.c.) operating points and the transistors are perfectly equal, then $r_{e1} = r_{e2} = r_e$ and $\alpha_1 = \alpha_2 = \alpha$:

$$i_{ag3}^2 = 4kT\,\Delta f\left[\frac{\alpha}{2-\alpha}\right]^2\left\{\frac{r_e+2r_{bb'}}{[r_e+r_{bb'}(1-\alpha)]^2}+\right.$$
$$\left.+2\left[1+\frac{r_e+r_{bb'}}{r_e+r_{bb'}(1-\alpha)}\right]^2\frac{1}{2\beta r_e}\right\}. \tag{8}$$

In case of a higher current gain factor the noise current is of lower intensity. With the condition of $\alpha \cong 1$,

$$i_{ag2}^2 = 4kT\,\Delta f\frac{1}{r_e}\left[1+2\frac{r_{bb'}}{r_e}+\frac{\left(2+\dfrac{r_{bb'}}{r_e}\right)^2}{\beta}\right]. \tag{9}$$

5. Recording the complete noise equivalent circuit model

It is obvious that a differential amplifier can be characterized by two transistors, two load impedances, a current source and the impedances terminating the two inputs. In case the behaviour is investigated from the point of noise, the internal impedance of the "current source" may be assumed to be negligible beside the input impedance of the commoned emitter point.

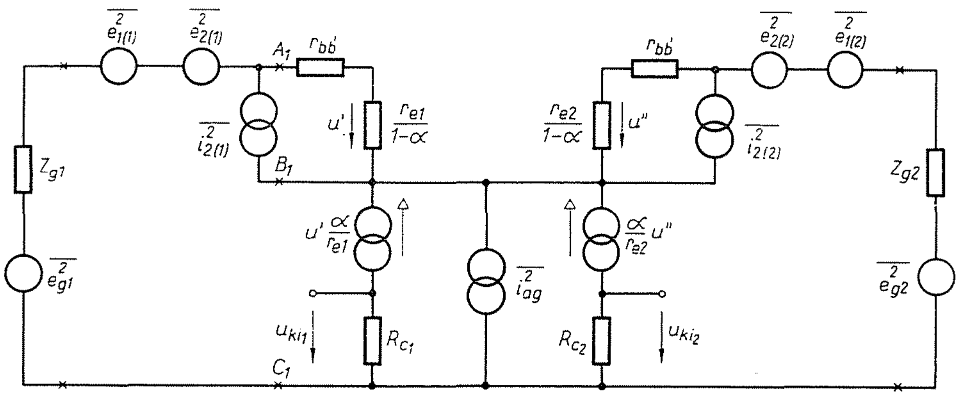

Fig. 5

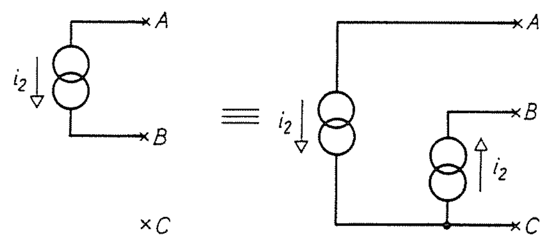

Fig. 6

Further approximations: the $r_{bb'}$ and the current gain factor of the two transistors is equal, the effect of saturation current noise is neglected and the feedback capacitance is omitted. To achieve a simpler form of the equation, the noise of the current source is characterized by an equivalent resistance, where:

$$R_e = \frac{4kT\,\Delta f}{\overline{i^2_{ag}}}. \tag{10}$$

In this case the entire noise equivalent circuit model is formed according to Fig. 5.

To simplify computations, current sources i_2 can be replaced by two, perfectly correlated generators (Fig. 6).

For the output voltages:

$$U_{ki1} = -\frac{\alpha}{r_{e1}} U' \qquad U_{ki2} = -\frac{\alpha}{r_{e2}} U''$$

$$U_{ki1} - U_{ki2} = -\alpha \left(\frac{U'}{r_{e1}} - \frac{U''}{r_{e2}} \right). \tag{11}$$

Let us assume that the impedances terminating the inputs are real.

The control of the differential amplifier may be reduced to three types: similar to e_{g1}, the voltage sources series connected with input No. 1., similar to e_{g2}, the voltage sources series connected with input No. 2., and similar to i_{ag} current sources feeding a common emitter point, carry out the control.

Introducing the symbol $S = \dfrac{\alpha}{r_{e1} + r_{e2} + (2r_{bb'} + R_{g1} + R_{g2})(1-\alpha)}$ the following transfer characteristics can be determined:

$$\frac{U_{ki1}}{e_{g1}} = -SR_{c1}, \qquad \frac{U_{ki2}}{e_{g1}} = SR_{c2},$$

$$\frac{U_{ki1}}{e_{g2}} = SR_{c1}, \qquad \frac{U_{ki2}}{e_{g2}} = -SR_{c2},$$

$$\frac{U_{ki1}}{i_{ag}} = SR_{c1}[r_{e2} + (R_{g2} + r_{bb'})(1-\alpha)],$$

$$\frac{U_{ki2}}{i_{ag}} = SR_{c2}[r_{e1} + (R_{g1} + r_{bb'})(1-\alpha)] \tag{12}$$

6. Noise figure determination

For this purpose resultant characteristics, $\overline{u_{ki1}^2}$; $\overline{u_{ki2}^2}$ and $\overline{(u_{ki1} - u_{ki2})^2}$ respectively, have to be determined, due to different noise sources (depending on the character of the output) and these have to be related to the value resulting from the noise of source impedance.

Let the internal resistance of the useful signal source be R_{g1}.

Introduce symbol $D = (A - |\alpha|^2)/|\alpha|^2$.

Now, three noise figures can be interpreted, depending on the output of the stage: a symmetrical and two asymmetrical ones:

$$F_{sz} = 1 + \frac{R_{g2}}{R_{g1}} + \frac{2r_{bb'}}{R_{g1}} + \frac{r_{e1} + r_{e2}}{2R_{g1}} +$$

$$+ \frac{\left\{r_{e1} + r_{bb'} + R_{g1} + \frac{r_{e2} R_{c1} - r_{e1} R_{c2} + (1-\alpha)\left[R_{c1}(R_{g2} + r_{bb'}) - R_{c2}(R_{g1} + r_{bb'})\right]}{R_{c1} R_{c2}}\right\}^2}{2R_{g1} r_{e1}} D +$$

$$+ \frac{\left\{r_{e2} + r_{bb'} + R_{g2} + \frac{r_{e1} R_{c2} - r_{e2} R_{c1} + (1-\alpha)\left[R_{c2}(R_{g1} + r_{bb'}) - R_{c1}(R_{g2} + r_{bb'})\right]}{R_{c1} R_{c2}}\right\}^2}{2R_{g2} r_{e2}} D +$$

$$+ \frac{\{r_{e2} R_{c1} - r_{e1} R_{c2} + (1-\alpha)\left[R_{c1}(R_{g2} + r_{bb'}) - R_{c2}(R_{g1} + r_{bb'})\right]\}^2}{R_e R_{g1}(R_{c1} + R_{c2})^2}, \tag{13}$$

$$F_{a1} = 1 + \frac{R_{g2}}{R_{g1}} + \frac{2r_{bb'}}{R_{g1}} + \frac{r_{e1} + r_{e2}}{2R_{g1}} +$$

$$+ \frac{[r_{e1} + r_{e2} + r_{bb'} + R_{g1} + (1 - \alpha)(R_{g2} + r_{bb'})]^2}{2R_{g1} r_{e1}} D +$$

$$+ \frac{|\alpha|^2 (r_{bb'} + R_{g2})^2}{2R_{g1} r_{e2}} D + \frac{[r_{e2} + (R_{g2} + r_{bb'})(1 - \alpha)]^2}{R_{g1} R_e}, \tag{14}$$

$$F_{a2} = 1 + \frac{R_{g2}}{R_{g1}} + \frac{2r_{bb'}}{R_{g1}} + \frac{r_{e1} + r_{e2}}{2R_{g1}} +$$

$$+ \frac{[r_{e1} + r_{e2} + r_{bb'} + R_{g2} + (1 - \alpha)(R_{g1} + r_{bb'})]^2}{2R_{g1} r_{e2}} D +$$

$$+ \frac{|\alpha|^2 (r_{bb'} + R_{g1})^2}{2R_{g1} r_{e1}} D + \frac{[r_{e1} + (R_{g1} + r_{bb'})(1 - \alpha)]^2}{R_{g1} R_e}. \tag{15}$$

a) Let us examine first the noise figure evolution at medium frequency (assuming $\alpha = \alpha_0 = A$), if $R_{c1} = R_{c2} = R_c$; $r_{e1} = r_{e2} = r_e$; $R_{g1} = R_{g2} = R_g$ (for current stability) and $1 - \alpha_0 \ll 1$.

In this case the symmetrical noise figure pertaining to the asymmetrical input control:

$$F_{sz} = 2\left[1 + \frac{r_{bb'}}{R_g} + \frac{r_e}{2R_g} + \frac{(r_e + r_{bb'} + R_g)^2}{2\beta_0 r_e R_g}\right], \tag{16}$$

This value is just the double of the noise figure valid for the common emitter stage delivered by the *Nielsen* formula [4], (a noise figure loss by 3 dB)

If the differential amplifier is controlled at both inputs (i.e. both R_{g1} and R_{g2} are "useful source resistances"), the noise figure coincides with that of a single common emitter stage.

In case of an asymmetrical output, the situation is of similar character, though the fifth and sixth term in (14) and (15) differ, and there is a surplus, the seventh term due to the current source noise, which now differs from zero as against the symmetrical case.

b) As a second and practically interesting case let us investigate the one, when $R_{g2} \ll R_{g1}$ but the other conditions in item a) are fulfilled. In this case

$$F_{sz} = 1 + \frac{R_{g2}}{R_{g1}} + \frac{2r_{bb'}}{R_{g1}} + \frac{r_e}{R_{g1}} + \frac{\left[r_e + r_{bb'} + R_{g1} + \frac{R_{g2} - R_{g1}}{2(1+\beta_0)}\right]^2}{2\beta_0 r_e R_{g1}} +$$

$$+ \frac{\left[r_e + r_{bb'} + R_{g2} + \frac{R_{g1} - R_{g2}}{2(1+\beta_0)}\right]^2}{2\beta_0 r_e R_{g1}} + \frac{(R_{g2} - R_{g1})^2}{4R_e(1+\beta_0)^2 R_{g1}},$$

$$F_{sz} = 1 + \frac{2r_{bb'}}{R_{g1}} + \frac{r_e}{R_{g1}} + \frac{(r_e + r_{bb'} + R_{g1})^2}{2\beta_0 r_e R_{g1}} +$$

$$+ \frac{(r_e + r_{bb'})^2}{2\beta_0 r_e R_{g1}} + \frac{R_{g1}}{4R_e(1+\beta_0)^2}, \tag{17}$$

hence the value of the noise figure may be by about 3 dB, lower than according formula (16).

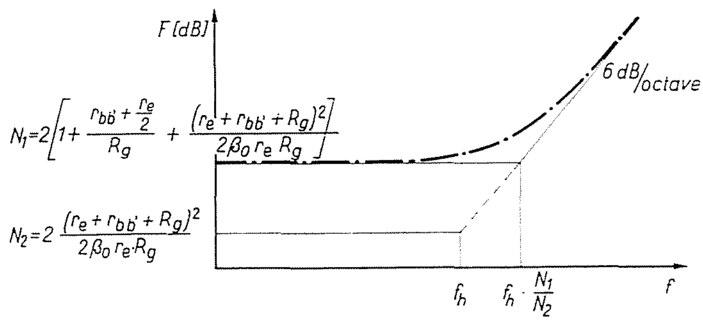

Fig. 7

In case of an asymmetrical output the situation is similar except that the noise of the current source is of a higher importance.

c) Supposing that in the investigated frequency range the common emitter current gain factor is well above 1, and that the feedback capacitance effect may be neglected, the frequency dependence of the noise figure can be traced back to the variation of the D factor. Using approximation $A = \alpha_0$, the value of D will have an ascent of 6 dB/octave, the pole frequency of D

is $f_h \cong f_\alpha / \sqrt{1+\beta_0}$. Accordingly, at high frequency the noise figure increases by a slope of 6 dB/octave. Completing the investigation under item a) by a frequency dependence (Fig. 7):

$$F_{sz} = 2\left[1 + \frac{r_{bb'}}{R_g} + \frac{r_e}{2R_g} + \frac{(r_e + r_{bb'} + R_g)^2}{2R_g\, r_e} D\right]. \tag{18}$$

7. The effect of feedback capacitance

If $\omega C_c > 1/r_e$, the noise figure is also influenced by C_c, especially for low (d.c.) operating point currents. For the sake of simplicity only the case of symmetrical input and output will be considered as this is the most common in high frequency application.

Applying the data from (3), the noise figure is evidently:

$$F_{sz} = 1 + \frac{r_{bb'}}{R_g} + \frac{r_e}{R_g}\left|\frac{\alpha + j\omega(r_{bb'} + R_g)\, C_c}{\alpha - j\omega\, r_e C_c}\right|^2 + \\ + \left|\frac{r_e + r_{bb'} + R_g}{\alpha - j\omega\, r_e C_c}\right|^2 \frac{A - |\alpha|^2}{2r_e R_g}. \tag{19}$$

The qualitative evaluation of formula (19) gives the expected result that feedback reduces the relative value of noise components resulting both from current distribution and emitter current (i.e., the noise figure). In case of multi-stage amplifiers, this reduction is not completely unequivocal, as C_c reduces also the power gain, thus the effect of the noise of subsequent stages may be important. A detailed investigation of this phenomenon is beyond the scope of the present study.

8. Noise and noise figure at low frequency

The equivalent circuit diagram in Fig. 1 characterizes the transistor noise only in the so-called shot noise range. According to measurements, other two noise effects emerge in the low frequency range, determinative for the transistor noise figure, below a given cut-off frequency. These two noise types are the so-called flicker noise and the so-called burst noise with characteristic impulse wave form [7], [8], [9]. From the physical point of view both can be attributed to the base emitter junction, acceptably exact data are, however, available only from measurements, as the real causes of these noises are not yet cleared up [9], [10]. Besides, the burst noise may not be equally intense for transistors of the same type even some may be perfectly noiseless. In the following, this latter case will be ignored. Fig. 8 shows the equivalent circuit model, valid in low frequences, where, according to [8]:

$$\overline{i_f^2} = KI_B^{\gamma} \cdot f^{\varepsilon} \Delta f \cong KI_B \cdot f^{-1} \Delta f$$

$$\gamma \cong 1 \text{ and } \varepsilon \cong -1 \tag{20}$$

In this approximation the r.m.s. value of the current belonging to the flicker noise source, is directly proportional to the base current and inversely

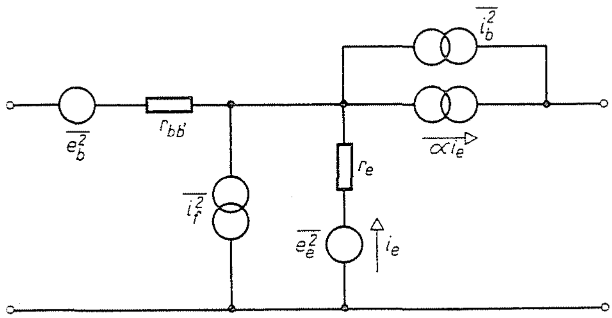

Fig. 8

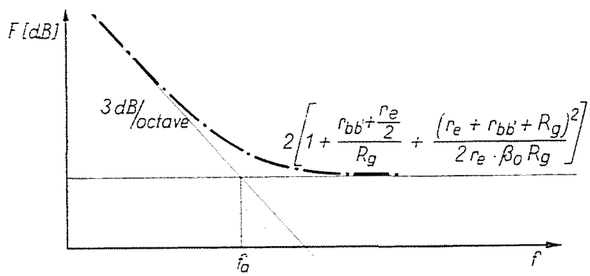

Fig. 9

proportional to the frequency [7], [8], while the noise current is totally independent from the voltage and current of the other noise generators.

In a symmetrical input and symmetrical output arrangement, if $R_{g1} = R_{g2} = R_g$; $r_{e1} = r_{e2} = r_e$; $R_{c1} = R_{c2} = R_c$, the narrow band noise figure can be written as:

$$F_{sz} = 2\left[1 + \frac{r_{bb'} + \frac{r_e}{2}}{R_g} + \frac{(r_e + r_{bb'} + R_g)^2}{2\beta_0 r_e R_g} + \frac{(r_{bb'} + R_g)^2}{R_g} K' I_B f^{-1}\right] \tag{21}$$

where:
$$K' = \frac{K}{4kT}$$

The frequency dependence of the noise figure is seen in Fig. 9.

The bottom cut-off frequency value is from (21):

$$f_a = \frac{K' I_B (r_{bb'} + R_g)^2}{R_g + r_{bb'} + \frac{r_e}{2} + \frac{(r_e + r_{bb'} + R_g)^2}{2\beta_0 r_e}} \,. \tag{22}$$

Applying the expression (22), the noise figure simplifies into:

$$F_{sz} = 2\left[1 + \frac{r_{bb'} + \frac{r_e}{2}}{R_g} + \frac{(r_{bb'} + r_e + R_g)^2}{2\beta_0 R_g r_e}\right]\left[1 + \frac{f_a}{f}\right]. \tag{23}$$

9. The relationship between the noise figure and the bandwidth

Until the noise power spectral density as a function of frequency, is constant and not function of frequency, the resultant noise figure is also frequency-independent. In the low frequency and high frequency ranges, the noise figure depends, however, also on the noise bandwidth of the system. The so-called broadband noise figure has an importance primarily in the low frequency range, as the relative bandwidth is great, in general, and thus a narrow band noise figure erroneous to describe the noise properties of the system.

In case of an ideal, infinite cut-off slope filter with frequency limits f_1 and f_2, the broad bandwidth noise figure can be calculated by applying relationship (23):

$$F_{sz} = 2\left[1 + \frac{r_{bb'} + \frac{r_e}{2}}{R_g} + \frac{(r_e + r_{bb'} + R_g)^2}{2\beta_0 r_e R_g}\right] \cdot \left[1 + \frac{f_a}{f_2 - f_1} \ln \frac{f_2}{f_1}\right] \tag{24}$$

if $f_2 > f_1$.

If the transmission is hindered by a low-pass RC filter of pole frequency f_2 and a high-pass RC filter of pole frequency f_1, the expression is modified into:

$$F_{sz} = 2\left[1 + \frac{r_{bb'} + \frac{r_e}{2}}{R_g} + \frac{(r_e + r_{bb'} + R_g)^2}{2\beta_0 r_e R_g}\right]\left[1 + \frac{2}{\pi} \frac{f_a}{f_2 - f_1} \ln \frac{f_2}{f_1}\right] \tag{25}$$

In both cases, pole frequency f_2 is lower than f_h in Fig. 7.

10. Practical viewpoints

a) If the differential amplifier is applied as a tuned amplifier, it has to be connected to the input by a transformer (tuned circuit). Let the secondary side impedance of the transformer be R_g and assuming small signal operation in case of a symmetrical drive, the noise figure is:

$$F' = \frac{1}{2} F \left(R_{g1} = \frac{R_g}{2}, \ R_{g2} = \frac{R_g}{2}\right), \tag{26}$$

while in case of an asymmetrical drive:

$$F' = F(R_{g1} = R_g \, , \ R_{g2} = 0) \, . \tag{27}$$

b) In case of broad bandwidth amplifiers, it is not expedient to apply a transformer as it is difficult to provide it with the suitable band width, and it can generate excess noises or even pick up external noises. In such cases the amplifier is mostly fed from asymmetrical output sources. From the point of bias stability, d.c. voltage amplifiers, should be supplied with equal resistances terminating the inputs ($R_g = R$), (Fig. 10).

$$F' = F(R_{g1} = R_g \, , \ R_{g2} = R) \, . \tag{28}$$

c) With RC-coupled amplifiers, high, d.c. loop gain can be brought about even in case there is no a.c. feedback (Fig. 11). Even the d.c. input terminals may not be identical, and from the point of a.c., highly different impedances

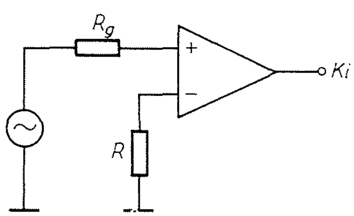

Fig. 10

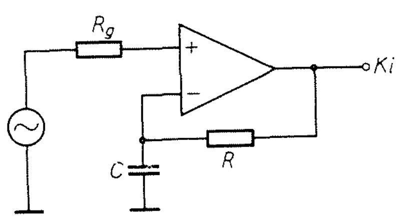

Fig. 11

may exist on the two sides. This fact enables the approximation of the noise figure of common emitter stages. In the case shown in Fig. 11:

$$F' = F(R_{g1} = R_g \, , \; R_{g2} = 0) \tag{29}$$

d) In feedback amplifiers the situation is similar save that the noise generated by the resistances of the feedback network has to be reckoned with. If the entire amplifier output resistance is negligible, then in most cases the resultant noise figure can be determined without difficulty.

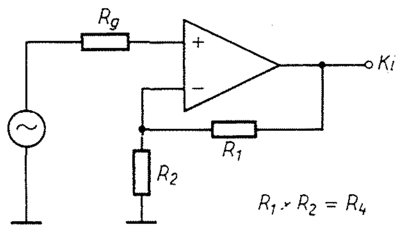

Fig. 12

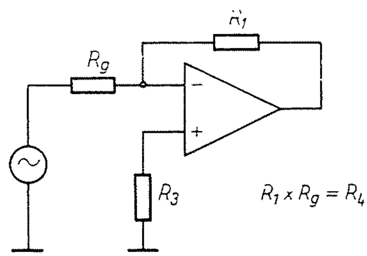

Fig. 13

For non-inverting amplifiers (Fig. 12):

$$F' \cong F(R_{g1} = R_g \, , \; R_{g2} = R_4) \tag{30}$$

For inverting amplifiers (Fig. 13):

$$F' \cong \left[1 + \frac{R_g}{R_1}\right] \cdot F(R_{g1} = R_4 \, , \quad R_{g2} = R_3) \tag{31}$$

e) The noise resistance of current sources (10) is generally negligible in cases a) and b) in Fig. 4. (4); (6).

The equivalent resistance of the two-transistor current source applied in monolithic integrated circuits may, however, influence the noise figure, especially in case of an asymmetrical output (7).

As an example optimum source resistance and minimum noise figure have been computed in three operating point of a sample-transistor, and complied in Tables 1 and 2 using the following data and assumptions:

$$r_{e1} = r_{e2} = r_e = \frac{U_T}{I_e}$$

$$r_{bb'1} = r_{bb'2} = r_{bb'} = 300 \text{ ohm}$$

$$\beta_{01} = \beta_{02} = \beta_0 = 300 . \tag{32}$$

Table 1

The current dependence of symmetrical and asymmetrical noise figures and optimal source resistances in case of the current source indicated in Figure 4c
$R_{g1} = R_{g2} = R_g$, and $R_{c1} = R_{c2} = R_c$, respectively

I_e	10 μA	100 μA	1 mA
r_e	2 kohm	260 ohm	26 ohm
F_{szopt}	2.13	2.22	2.66
$R_{gopt(sz)}$	50 kohm	8.2 kohm	2.25 kohm
$F_{a2\,opt} = F_{a1opt}$	2.26	2.53	4.13
$R_{gopt(a)}$	92 kohm	16.8 kohm	4.27 kohm
$R_e = 4kT\,\Delta f/i_{ag}^2$	885 ohm	23.3 ohm	0.305 ohm

Table 2

The current dependence of symmetrical and asymmetrical noise figures and optimal source resistances in case of the current source to be seen in Figure 4c
$R_{g1} = R_g$; $R_{g2} = 0$; $R_{c1} = R_{c2} = R_c$

I_e	10 μA	100 μA	1 mA
r_e	2.6 kohm	260 ohm	26 ohm
F_{szopt}	1.09	1.16	1.47
$R_{gopt(sz)}$	83.4 kohm	11.6 kohm	2.96 kohm
$F_{a2\,opt}$	1.18	1.4	2.73
$R_{gopt(a)}$	127 kohm	23.5 kohm	5.5 kohm
$R_e = 4kT\,\Delta f/i_{ag}^2$	885 ohm	23.3 ohm	0.305 ohm

The transistors of the current source were regarded as similar to the transistors of the long-tailed pair. It was assumed that the conditions in formula (8) are valid and also that the current of the source is two times that of one of the differential transistors. The current dependence of β, the flicker effect and the high frequency behaviour had been neglected.

On basis of the results obtained the following may be stated:

1. In case of asymmetric drive, the asymmetrical input termination is more favourable from the point of noise.

2. In case of high current and asymmetrical output the noise of the current source much influences the noise figure.

3. The optimum noise figure indicated for one transistor can be achieved only in case of symmetric drive and symmetric output.

4. From the point of noise the most favourable circuit configuration is the current source supplied only with resistance (see Fig. 4a).

5. The noise figure of the input stage in multi-stage, feedback amplifiers is minimal, if the terminating resistance on the non-driven input is of a low value. (See $R_1 \times R_2 = R_4$ and R_3, in Fig. 12 and 13, respectively.)

6. In monolithic integrated circuit amplifiers the bias current of the transistors of the input stage, as well as the type of the current source are given, thus the noise factor proper to the applied source resistance has made up. There is no possibility for optimation in such a case.

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Summary

The noise problems of differential amplifiers are discussed in case of symmetrica and asymmetrical input and output. New and more accurate results were obtained than in noise investigations [1] using simple, two-stage cascaded transistor amplifiers. Noises from different type current sources have been examined. The applied mode of treatment permits to determine the noise factors of transistorized differential amplifiers and the value of optimum source resistance in case of an arbitrary circuit configuration. Finally, some points of view are given for the circuit configuration of low-noise input stages.

References

1. SOLOMON, J. E.: Cascade Noie Figure of Integrated Transistor Amplifiers. Motorola Technical Information Note, AN-223, 1970.
2. BRUBAKER R.: Semi-conductor Noise Figure Consideration. Motorola Application Note, AN-421, 1970.
3. AMBRÓZY A.: Noise, noise measuring Engineers' training course lecture, 1968. (In Hungarian).
4. NIELSEN E. G.: Behavior of Noise Figure in Junction Transistors. Proc. IRE, Vol. 45, p. 957. June, 1957.
5. LYNN D.—MEYER C. S.—HAMILTON D. J.: Analysis and Design of Integrated Circuits. McGraw-Hill Book Company, 1967.
6. TELKES B.: Transistorized d.c. amplifiers in measuring techniques and automatics. (In Hungarian) Műszaki Könyvkiadó, Budapest, 1970.
7. CHENETTE E. R.: Low Noise Transistor Amplifiers Solid State Desigu, Vol. 5, No. 2, February, 1964.
8. PLUMB J. L.—CHENETTE E. R.: Flicker Noise in Transistors. IEEE Trans., on Electron Devices, ED-10, No. 5, September, 1968.
9. BRODERSEN A. J.—CHENETTE E. R.—JAEGER R. C.: Noise in Integrated Circuit Transistors. IEEE Journal of Solid State Circuits, SC-5, No. 2, April, 1970.
10. CONTI M.: Surface and Bulk Effect in Low Frequency Noise in PNP Planar Transistors, Solid State Electronics, No. 11, November, 1970.

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