SLEWING RATE IN OPERATIONAL AMPLIFIERS

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1. Introduction

The field of application of high gain operational amplifiers was considerably widened by the developments in integrated circuit technology and network configurations. Modern application technics treats operational amplifiers as universal circuit units and applies their advantageous properties to improve the parameters of a number of classic network solutions. The designers of integrated operational amplifiers try to approach the so-called "ideal operational amplifiers" with their constructed networks. The ideal operational amplifier will be a symmetrical input and asymmetrical output infinite differential and zero common mode gain unit, with an infinite bandwidth and input resistance, zero offset voltage, zero bias and offset currents as well as zero output resistance which, at the same time, does not bring about any accessory noise. Over and above the aforementioned properties let us mention another important one: in case of arbitrary large input and output signals the operation is still similar, that is, linear. The real networks are rather different from abstract "ideal operational amplifiers". The specifications in the data sheets reflect just these differences with the aid of operational parameters, limit-data and diagrams characterizing the operation.

One of the major hindrances of applying operational amplifiers is the finite bandwidth, in close connection with it the maximum possible slewing rate, and the frequency relation of the output voltage swing. [1, 2, 3, 5, 6, 7, 8, 9]. With the improved variety of the μA 702 and μA 709 type integrated networks, which can be regarded as classic ones, the solution of this problem was the main aim and it is to be expected that further improvements will follow in this very direction.

A suitable interpretation of the sphere of concepts is important not only for the designing production engineer but also for the user, as these very phenomena may influence in a major way the feasibility of the given tasks. The effect of maximum slewing rate manifests itself the most pregnantly in high loop gain amplifiers, in networks including the combination of operational amplifier—nonlinear units as well as in active filters.

In the following, the concept and calculation of maximum slewing rate will be discussed, introducing a simplified, non-linear equivalent circuit model of operational amplifiers.

2. The concept of slewing rate

By definition, slewing rate is the maximum speed of the voltage change recorded at the output of the amplifier in case a proper amplitude, ideal square wave function or step function signal is input in the amplifier, which just does not lead to an overdriven state of the stages of the amplifier — after the reproduction of the transients either in a positive or a negative direction [4].

In case of non-feedback amplifiers the above definitions are of no special importance. In such a case all stages of the amplifier operate almost linearily and thus the resulting slewing rate is proportional to the maximum speed of the output voltage change of a linear, low-pass, two-port weighting function.

In case of feedback amplifiers, one or more internal stages of the amplifier may, however get into an overdriven state in the course of transient responses and in the following this stage, or stages, determine the time function of the output signal. Slewing rate is thus a characteristic of the amplifier, related to the internal limitations of the active devices and the reactance determining the frequency response of the linear system. Thus, slewing rate is directly related with the value of the feedback amplifier compensating units and their position in the system.

The frequency-dependence of the maximum sinusoid output voltage swing can be traced back to similar physical reasons, but (neglecting here and now a trivial definition) it is an important data for the user in case of both feedback and non-feedback amplifiers.

3. Slewing rate in case of linearity

The two-port transfer function of the general low-pass contains n real and complex poles. The slewing rate SWR value is pressed by 1. in a general way.

$$SWR = \frac{d}{dt}\left[\mathfrak{L}^{-1}\left\{\frac{U_{0_M}}{p\left(1+\frac{p}{p_1}\right)\left(1+\frac{p}{p_2}\right)\dots\left(1+\frac{p}{p_n}\right)}\right\}\right]\Bigg|_{t=t_0} \qquad (1)$$

where U_{0_M} is the maximum static output voltage; $p_1, p_2 \dots p_n$ the poles of the transfer function and t_0 the least time where the second derivative of the right-hand-side term in brackets is zero.

The expression (1) immediately shows that in case of a single pole the output slewing rate is:

$$SWR = U_{0_M}\,\omega_1 \tag{2}$$

where ω_1 is the absolute value of the pole, i.e. the bandwidth of the amplifier.

The SWR value is delivered by (3) and (4) for the poles on the real axis and a complex conjugate pole pair, respectively (Figs 1 and 2). In the figures, the slewing rate data normalized to values $U_{0_M}\,\omega_1$, and $U_{0_M}\,\omega_0$, respectively, indicate the relationship between the slewing rate and the position of the poles.

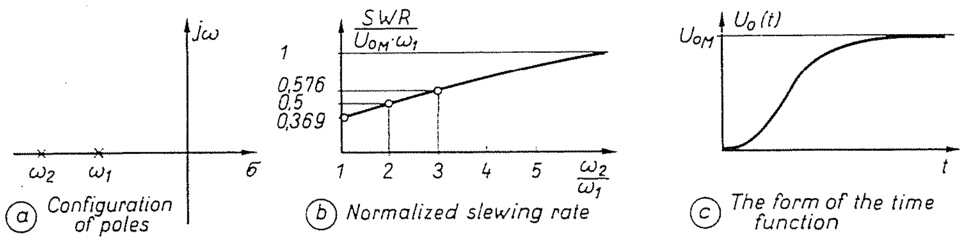

Fig. 1

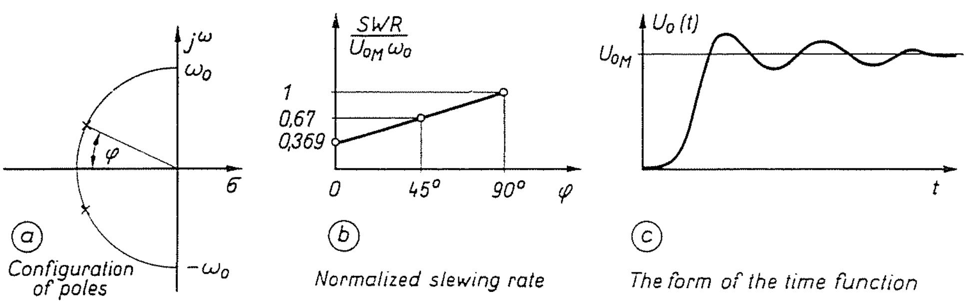

Fig. 2

$$SWR = U_{0_M}\,\omega_1 \left(\frac{\omega_1}{\omega_2}\right)^{\frac{1}{\frac{\omega_2}{\omega_1}-1}} \tag{3}$$

$$SWR = U_{0_M}\,\omega_0 \cdot e^{-\varphi\,\mathrm{ctg}\,\varphi} \tag{4}$$

where ω_0 is the absolute value of the complex roots, φ is the angle between the complex vector characterizing the roots and the negative real axis.

Relationship (3) demonstrates that the approximate relationship (5) for the resultant rise time of systems with several real roots, gives a rather rough approximation for SWR values.

$$t_{er} \cong \sqrt{t_1^2 + t_2^2 + \ldots + t_n^2}\,, \tag{5}$$

where $t_1; \ldots ; t_n$ are the independent rise times pertaining to the individual poles.

Higher degree two-ports can be investigated with the aid of (1), of minor importance for operational amplifiers.

4. The simplest non-linear equivalent circuit

The equivalent circuit model in Fig. 3 [4], symbolizes a three-stage, feedback amplifier. Two-ports A_1, A_2 and β are frequency-independent and can be separated from each other ideally. The maximum output voltage of amplifier A_2 is $\pm U_{0M}$, the middle stage is symmetrically current limited and it is supposed to have transfer-characteristic with a linear pole, according to Fig. 4.

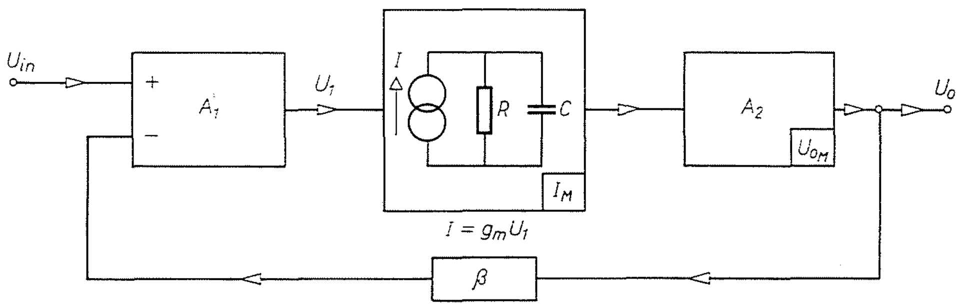

Fig. 3

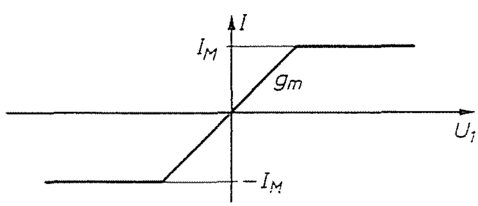

Fig. 4

The response of the system to the step function change is rather dependent on the input signal level. In case of small input signals, the system works in a linear mode of operation. In case of a sufficiently high control signal it can cause an overdriven state in the second stage at time $t = 0$, and during this period current I_M and capacitance C determine the slewing rate of the output voltage. After the transients the system will return to linear region, provided the stipulation by definition in item 2 is valid for the input signal amplitude.

With notations in Fig. 3, and supposing $A_2 RI_M > U_{0_M}$, always present in practice, the transfer functions of the linear system are the following:

$$U_0(p) = U_{\text{in}}(p) A_v \frac{1}{1+\frac{pRC}{1+A\beta}} \tag{6}$$

where $$A_v = \frac{A}{1+\beta A}; \qquad A = A_1 A_2 g_m R$$

$$I(p) = U_{\text{in}}(p) \frac{A_v}{A_2 R} \frac{1+pRC}{1+p\frac{RC}{1+\beta A}} \tag{7}$$

In case of step function input signal $U_{\text{in}}(p) = U_{\text{in}}/p$, the output time functions can be written in the following form:

$$U_0(t) = U_{\text{in}} A_v \left\{1 - \exp\left[-\frac{t}{RC}(1+\beta A)\right]\right\} \tag{8}$$

$$I(t) = U_{\text{in}} \frac{A_v}{A_2 R} \left\{1 + \beta A \exp\left[-\frac{t}{RC}(1+\beta A)\right]\right\} \tag{9}$$

If the maximum value of the current $I(t)$ exceeds the limit determined by the current-limit, the second stage will get into an overdriven state, relationships (6) and (7) are not valid any more. As $I(t)_{\max} = I(0)$, the limit here is:

$$U'_{\text{in}} \frac{A_v}{A_2 R}(1+\beta A) = I_M \tag{10}$$

$$U'_{\text{in}} = \frac{I_M R A_2}{A_v(1+\beta A)} = \frac{I_M}{g_m A_1} \tag{11}$$

The permissible maximum input signal (U'_{in}), still with no overdriving, is inversely proportional with the gain of the amplifier preceding the limited current stage.

During the time of saturation the time function of the output signal is:

$$U_0^*(t) = I_M R A_2 \left[1 - \exp\left(-\frac{t}{RC}\right)\right] \tag{12}$$

The complete time function is seen in Fig. 5. For the interval $t = 0$; $t = t_1$ the second stage is in an overdriven state. According to the former, the figure is valid if the condition $I_M RA_2 > U_{0_M} \geq A_v U_{\text{in}}$ is met.

By using relationship (12), time period t_1 can be computed in the following way:

$$U_0(t)=\frac{U_{\rm in}}{\beta}-\frac{I_M RA_2}{A\beta}=\frac{U_{\rm in}A_v}{\beta A}[1-a+A\beta] \qquad (13)$$

$$t_1=-\tau\ \ln\left[1-\frac{1+A\beta}{aA\beta}+\frac{1}{A\beta}\right]=-\tau\ \ln\left[1-\frac{1}{a}\right]\left[1+\frac{1}{A\beta}\right] \qquad (14)$$

$$a=\frac{I_M A_2 R}{U_{\rm in}A_v}>1; \qquad \tau=RC.$$

In Figs 6 and 7 the values of t_1 and $U_0(t_1)$ are plotted vs. a and $A\beta$, respectively.

From expression (12), the output slewing rate is [4]:

$$SWR=\frac{I_M A_2}{C}. \qquad (15)$$

From Fig. 5, it appears, without demonstration that the maximum rise rate is the highest in the overdriven region.

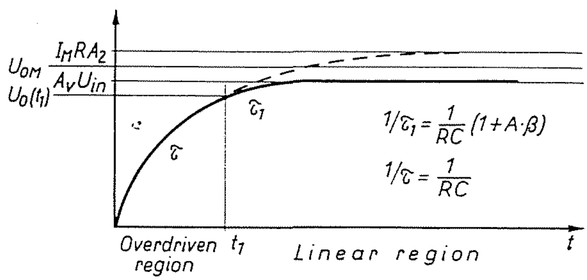

Fig. 5

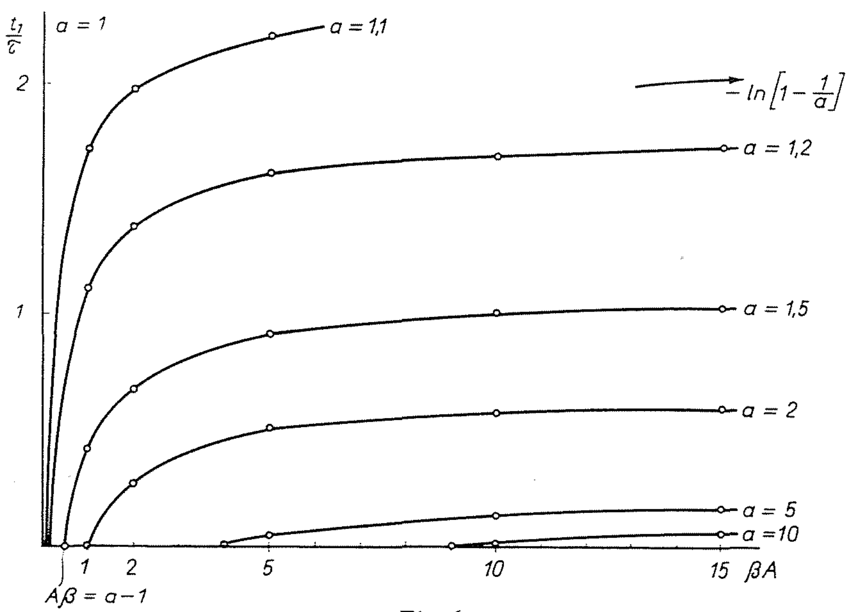

Fig. 6

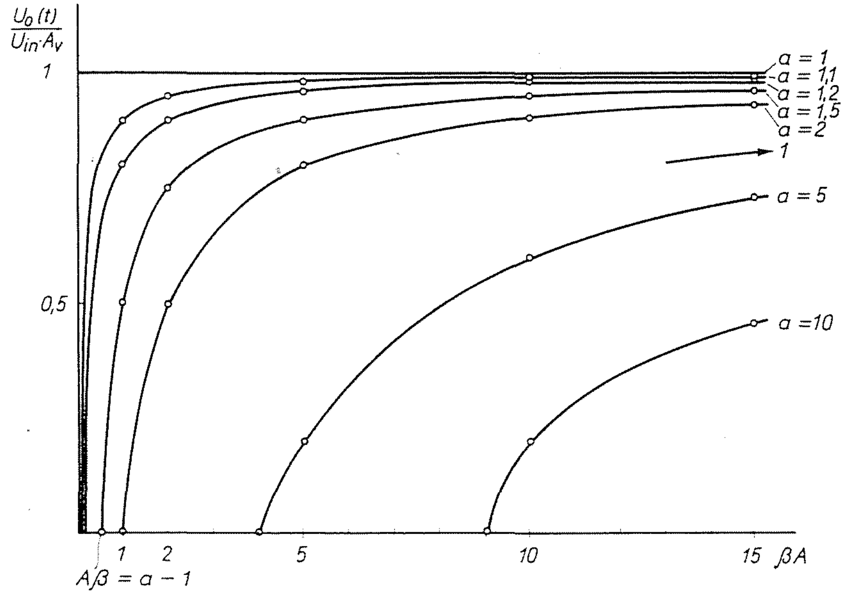

Fig. 7

Fig. 6, shows that in case of $U_{\text{in}} A_v > I_M A_2 R$, the system will never get into the linear region, as the second stage is in an overdriven state when the output signal corresponds the maximum output voltage. At a sufficiently high input signal, the complete rise time becomes independent from the input signal level and supposing that $U_{0_M} \ll I_M R A_2$, is approximately:

$$t_r \cong \frac{U_{0_M}}{SWR} \tag{16}$$

In such a case the rise section of the output signal is characteristically linear.

5. Non-linear equivalent circuit model with two poles

Multi-stage amplifiers have, in every case, several dominant high frequency poles, thus, A_1 and A_2 in the equivalent circuit model, (Fig. 3), can both be frequency-dependent. In this case it is much more complicated to investigate the feedback amplifier, as, increasing the feedback requires stability investigation of the linear system and necessity of compensation emerges.

Compensation is, in most cases, a shifting of the lowest frequency pole in a way that the linear transfer function of the feedback system fulfils certain conditions (for example, maximally flat or critically damped response). Compensation necessitates to build in external capacitances, the point of building in having a great influence on the slewing rate value.

In further investigations the system will be supposed to have two poles, hence the second pole ω_2 can be ordered to either twoport A_1 or A_2. Once again, feedback is independent from the frequency and compensation is carried out by increasing condenser C of the second stage, affecting the frequency of pole $\omega_1 = 1/RC$. The aim of compensation in our case is to bring about a maximally flat transfer function.

Concerning the feedback linear system, the following relation may be written up:

$$U_0(p) = U_{\text{in}} A_v \frac{1}{1 + p\dfrac{2\xi}{\omega_0} + \dfrac{p^2}{\omega_0^2}} \tag{17}$$

where:

$$\omega_0 = \sqrt{\omega_2\,\omega_1(1+\beta A)}\,; \qquad \xi = \frac{1}{2}\,\frac{\sqrt{\dfrac{\omega_1}{\omega_2}} + \sqrt{\dfrac{\omega_2}{\omega_1}}}{\sqrt{1+\beta A}}\,.$$

The condition of a maximally flat response is:

$$\frac{1}{\omega_1} = \frac{1}{\omega_2}\left[\sqrt{\frac{1+\beta A}{2}} + \sqrt{\frac{1+\beta A}{2} - 1}\right]^2. \tag{18}$$

If $\beta A \gg 1$, with a small neglection

$$RC \cong \frac{2}{\omega_2}\,\beta A\ . \tag{19}$$

Relationship (19) is suitable for computing the required compensating condenser. The response to step function of the closed, linear system is:

$$U_0(t) = A_v U_{\text{in}}\left[1 - e^{-\frac{\omega_0}{\sqrt{2}}t}\left(\sin\frac{\omega_0}{\sqrt{2}}t + \cos\frac{\omega_0}{\sqrt{2}}t\right)\right]. \tag{20}$$

The time function of current I is investigated in two cases depending on whether the second pole pertains to the first or the third stage.

a) The second pole is in stage A_2

$$I(t) = \frac{A_v U_{\text{in}}}{A_2 R}\left[1 + e^{-\frac{\omega_0}{\sqrt{2}}t}\left(\beta A \sin\frac{\omega_0}{\sqrt{2}}t + \beta A\ \cos\frac{\omega_0}{\sqrt{2}}t\right)\right]. \tag{21}$$

The maximum value of $I(t)$ occurs at the time $t = 0$.

$$I(t)_{\max} = I(0) = \frac{A_v U_{in}}{A_2 R} (1 + \beta A) \tag{22}$$

In case of a sufficiently high U_{in} voltage, the second stage will be in the over-driven state:

$$U'_{in} = \frac{I_M}{g_m A_1} \tag{23}$$

The result is thus similar to the one indicated in item 4. The maximum output slewing rate depends, however, on the value of loop gain, through relationship (19).

In a non-linear range, the time function of the output signal is also influenced by the frequency dependence of A_2:

$$U_0(t) = I_M R A_2 \left[1 - \frac{\omega_2 e^{-t\omega_1}}{\omega_2 - \omega_1} + \frac{\omega_1 e^{-t\omega_2}}{\omega_2 - \omega_1}\right] \tag{24}$$

Supposing that the loop gain is sufficiently high, that is, $\omega_2/\omega_1 \gg 1$

$$U_0(t) = I_M R A_2 [1 - e^{-t\omega_1}] \tag{25}$$

$$SWR = \frac{I_M A_2 R}{2\beta A} \omega_2 \tag{26}$$

After compensation, SWR is about inversely proportional to the loop gain.

The character of the time functions $U_0(t)$ and $I(t)$, appears from Fig. 8. Time t_1 can be computed also in this case by (14).

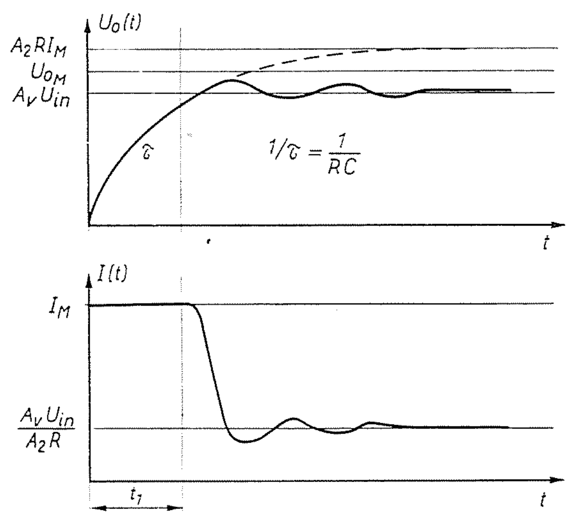

Fig. 8

b) In stage A_1, the second pole

$$I(t) = \frac{U_{\text{in}} A_v}{A_2 R} \left[1 - e^{-\frac{\omega_0}{\sqrt{2}} t} \left\{ \cos \frac{\omega_0}{\sqrt{2}} t + \right.\right.$$
$$\left.\left. + \left(1 - \sqrt{2 \frac{\omega_2}{\omega_1} (1 + \beta A)} \right) \sin \frac{\omega_0}{\sqrt{2}} t \right\} \right]. \tag{27}$$

The maximum value of current $I(t)$ occurs at the following point:

$$\operatorname{tg} \frac{\omega_0}{\sqrt{2}} t = \frac{- \frac{\omega_0 \sqrt{2}}{\omega_1}}{2 - \frac{\omega_0 \sqrt{2}}{\omega_1}} \cong + 1. \tag{28}$$

In case $A\beta \gg 1$, the neglection is permitted. The value of the maximum current is approximately proportional to the value of $A\beta$, in case of a high loop gain.

$$I(t)_{\max} = \frac{U_{\text{in}} A_v}{A_2 R} e^{-\frac{\pi}{4}} \sqrt{2} A\beta . \tag{29}$$

The limiting value of the input voltage

$$U'_{\text{in}} = \frac{I_M}{g_m A_1} \frac{e^{\frac{\pi}{4}}}{\sqrt{2}} = \frac{I_M}{g_m A_1} 1.56 . \tag{30}$$

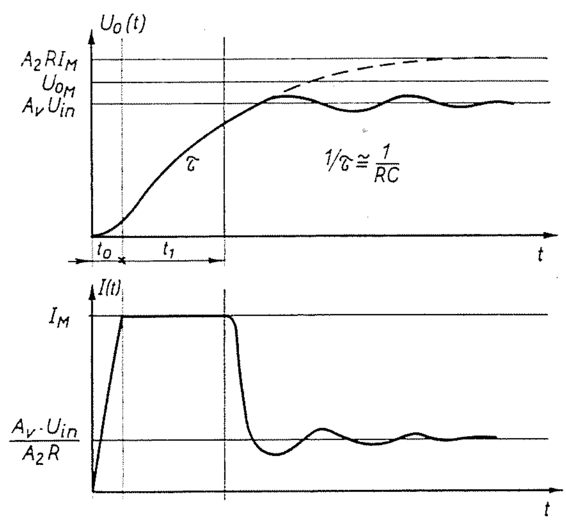

Fig. 9

The output signal can be divided into three stages, according to Fig. 9. At time t_0 the system is operating linearily, during time t_1, the second stage gets into an overdriven state, and finally the network returns to the linear region. Time t_0 can be approximated, by (31), provided $U_{\text{in}} \gg U'_{\text{in}}$. Approximating the time function of the current with a starting slope, at $t = 0$:

$$t_0 \cong \frac{I_M A_2 R}{A_v U_{\text{in}}} \frac{\omega_1}{\omega_0^2} = \frac{I_M}{A_1 g_m U_{\text{be}}} \frac{1}{\omega_2}. \tag{31}$$

If $\omega_2 \gg \omega_1$, then, in general, time t_0 is negligible as against t_1. This supposition involves that t_1 can be computed at a sufficient accuracy from (14) while the *SWR* value is given by (26) at a good approximation.

6. The limitations of the equivalent circuit models

The real amplifying systems are much more complex than the ones described above. The differences and the evolving problems are summarized in the following.

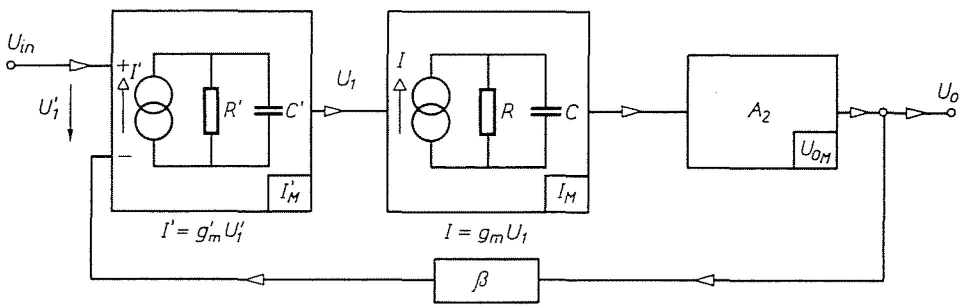

Fig. 10

a) In general, high gain systems consist of three amplification stages: two for voltage and one for output. The linear transfer function thus contains three active poles. Compensating means to form expediently the frequency characteristic of a three-pole feedback system. Supposing that the frequency of the third pole is much higher than that of the other two, the results obtained for the two-pole systems, can be generalized also for this one.

In such cases the position of the active poles of the feedback system is influenced but by the loop gain and the poles on the two lower frequencies. In more complex cases: numerical analysis is the general method of evaluation.

b) In multi-stage systems, in general, the output of all stages is limited. The above discussed symmetrical circuit limitation is typical for differential amplifiers. In general, several stages may be overdriven at the same time. As an example and by using former results, the operation of the system in Fig. 10, will be examined. The arrangement is an extended variety of the circuit seen in Fig. 3. According to our assumptions, compensation is done in the second stage, by modifying the frequency of the dominant pole. The amplifier A_1, similarly to stage two, can be characterized with a single linear pole and a current limited transfer function. The input voltage limits of the overdriven states of the individual stages are:

$$U'_{\mathrm{in}_1} = \frac{I'_M}{g'_m} \tag{32}$$

$$U'_{\mathrm{in}_2} = \frac{I_M}{g_m g'_m R'} 1.56 \tag{33}$$

as $A_1 = g'_m R'$ in (30). The first stage is overdriven first, if:

$$U'_{\mathrm{in}_1} < U'_{\mathrm{in}_2} \tag{34}$$

$$I'_M < \frac{I_M 1.56}{g_m R'} \tag{35}$$

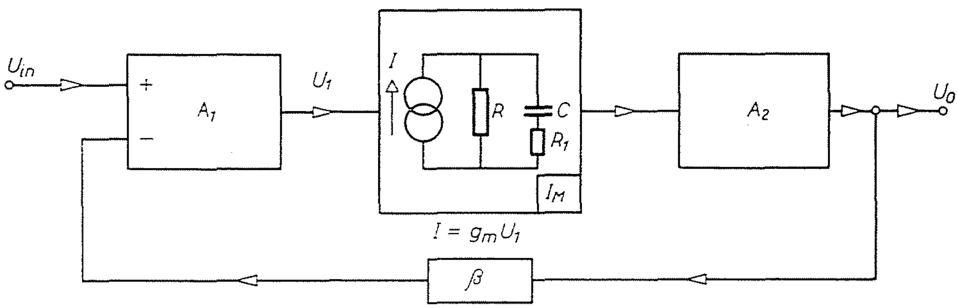

Fig. 11

c) Besides the symmetrical current limitation and depending on the circuit arrangement of the stage, asymmetrical current and voltage limitation is also frequent. Often (e.g. in sample and hold circuits) the output stage and the loading capacity determine the maximum slewing rate. The special arrangements require individual analyses.

d) Compensation does not mean in each case the increase of the capacity pertaining to the dominant pole. Compensation is often done by means of series connected R and C, adding one pole and one zero to the transfer function. This method may much increase the maximum slewing rate at about $t = 0$, in the response to the step function. This phenomenon will be illustrated on the circuit in Fig. 11. For the sake of simplicity let us assume that the loop gain is sufficiently high and A_1, A_2 and β can be regarded as frequency-independent, compared to the dominant pole of the system.

The responses of the linear system to the step function are:

$$U_0(t) = U_{in} A_v \left\{1 - \left[1 - \frac{(1+\beta A) R_1}{(1+\beta A) R_1 + R}\right] e^{-\omega_p t}\right\} \tag{36}$$

$$I(t) = \frac{U_{in} A_v}{A_2 R} \left\{1 + \left[\frac{(1+\beta A)(R_1+R)}{(1+\beta A) R_1 + R} - 1\right] e^{-\omega_p t}\right\} \tag{37}$$

where $\omega_p = \dfrac{1+\beta A}{(1+\beta A) R_1 C + RC}$, and let $\omega_p' = \dfrac{1}{C(R_1+R)}$.

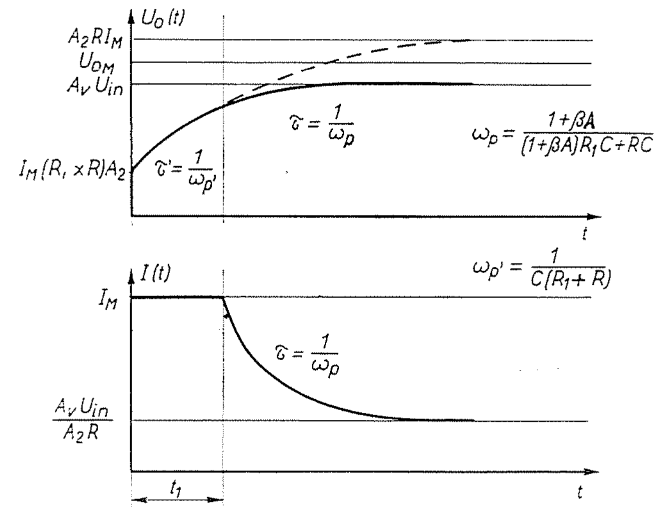

Fig. 12

The second stage will enter the non-linear region if the following inequality is fulfilled:

$$I(t)_{max} = \frac{U_{in} A_v}{A_2 R} \frac{(1+\beta A)(R_1+R)}{(1+\beta A) R_1 + R} > I_M \tag{38}$$

That is:

$$U_{in}' = \frac{I_M}{A_1 g_m} \frac{(1+\beta A) R_1 + R}{R_1 + R} \tag{39}$$

In the overdriven region the time function of the output signal is only determined by the second and third stage.

$$U_0(t) = I_M R A_2 \left[1 + \left(\frac{R_1}{R+R_1} - 1\right) e^{-\omega_p' t}\right] \tag{40}$$

The signal forms are indicated in Fig. 12.

7. The frequency dependence of the maximum output voltage swing

The low frequency output voltage swing is generally limited by the output stage or the driving stage. By increasing the frequency, one of the intermittent capacitive load stages may get to the limit of linear maximum output voltage swing, before overdriving the output stage. Thus the not clipped maximum sinusoid output signal is determined by this interior stage. In high loop gain systems this fact often limits the usility of the amplifier, although the small signal bandwidth of the amplifier may be many times that of the limit frequency of the maximum output voltage swing.

a) Systems with one pole [4].

For the network in Fig. 3, the maximum possible distortion free sinusoid output voltage, affected by the limitation in stage two, can be computed in the following way:

$$U_{0\,\max} = I_M \frac{R}{|1 + j\omega RC|} A_2 . \tag{41}$$

In case this is less than the maximum output voltage swing of the output stage, the maximum value of the output voltage is determined by the second stage.

If $1/\omega C \ll R$ is fulfilled, relationship (41) simplifies into:

$$U_{0\,\max} = \frac{I_M}{\omega C} A_2 . \tag{42}$$

The limit frequency of the dynamic range, i. e. the frequency where the level determined by the output stage and the intermediate stage is equal:

$$f_h = \frac{I_M A_2}{2\pi C\, U_{0_M}} . \tag{43}$$

Above this frequency the dynamic range decreases at a slope of 6 dB/octave, viz. hyperbolically, versus of frequency.

$$U_{0\,\max}(f) = U_{0_M} \frac{f_h}{f} . \tag{44}$$

b) System with two poles [4].

Using the expression (20) the power bandwidth considerations and the loop gain are related by:

$$U_{0\,\max} = \frac{I_M RA}{2A\beta} \frac{\omega_2}{\omega} . \tag{45}$$

Thus the maximum of the output voltage is inversely proportional to the loop gain of the compensated amplifier.

In case also A_2 is frequency-dependent and its cut-off slope is 6 dB/octave, the maximum output voltage swing decreases by 12 dB/octave, as a function of frequency. Decrease of the maximum output voltage swing of the output stage by 6 dB/octave over limit frequency f_h, leaves the condition (42) inaffected.

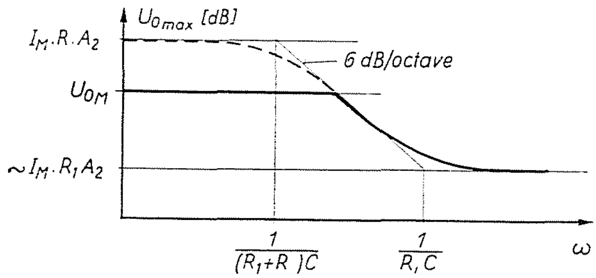

Fig. 13

In case of the series RC-member compensation in Fig. 11, the power bandwidth changes — according to the sense — in relation to the function in Fig. 13.

8. Results of the measurements

To prove the presented principles and computation methods, measurements had been carried out and, in case of a few commercial integrated circuits, the records in the data sheets had been applied.

a) The HIKI-made EH3 type integrated thin film operational amplifier consists of two differential amplifiers and one unity-gain output stage. Compensation is done in the second stage, by increasing the collector-base capacitances of the transistors by external components.

With the aid of the Miller-transformation, our hitherto results can be generalized also for collector-base capacitance:

$$SWR = \frac{I_M A^* A_2}{C[1 + A^*]} \cong \frac{I_M A_2}{C} \tag{46}$$

where I_M is the maximum current of the stage, before, A^* is the voltage gain of the stage containing collector-base capacitance, A_2 is the voltage gain of the output stage and $C[1 + A^*]$ is the Miller capacitance. With similar symbols, the limit frequency of the power bandwidth is:

$$f_h = \frac{I_M A^* A_2}{2\pi C[1 + A^*]\, U_{0_M}} \cong \frac{I_M A_2}{2\pi C\, U_{0_M}} \tag{47}$$

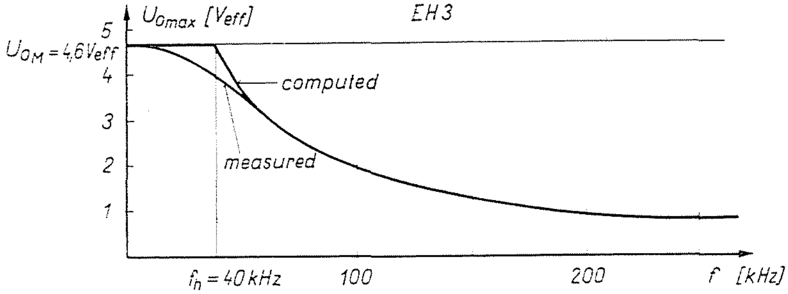

Fig. 14

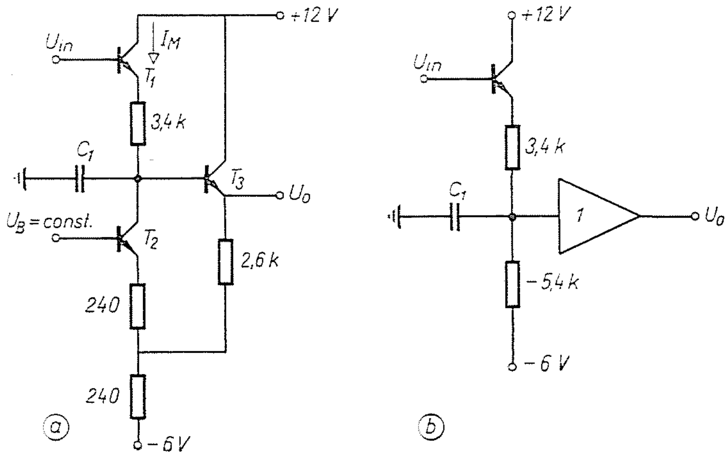

Fig. 15

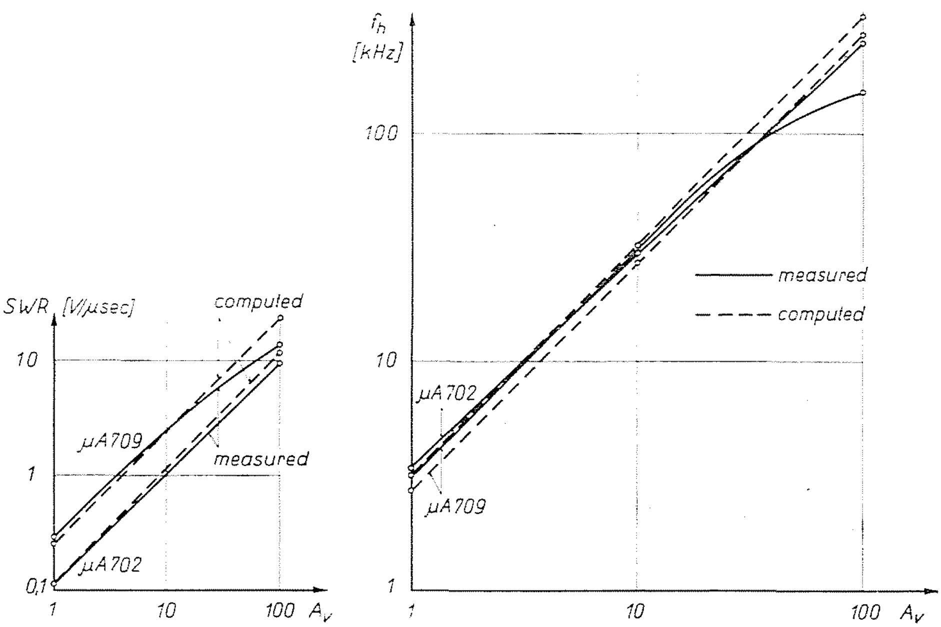

Fig. 16

Concerning the amplifier EH3, $A_2 = 1$; $I_M = 0.74$ mA, $U_{0_M} = 13$ V_{p-p}, and $C = 220$ pF. Calculated and measured SWR values are 3.3 and 2.2 V/μsec, respectively.

The frequency dependence of the power bandwidth is seen in Fig. 14. Beyond the limit frequency f_h, relationship (42.) indicates the maximum output voltage swing at a close accuracy.

b) In circuit type μA 702, [9, 10], compensating is done in the positive feedback output stage. The circuit diagram of the output stage and its equivalent circuit model, in case of compensation, is seen in Figs 15a and b C_1 is the compensating capacitance, $I_M = 1.1$ mA is the operating point current of transistor T_1. The catalogue data and the computed values are in good agreement (see Fig. 16).

c) In circuit type μA 709 [9, 10], compensation means to increase the feedback capacitance of the second amplifying and asymmetric conversion stage. The gain of the output stage $A_2 = 30$, the maximum current of the driving, current limited first stage is $I_M = 40$ μA. The measured and computed data vs. gain of the feedback system are shown in Fig. 16.

Comparison of the measured and computed results indicates that the assumed equivalent circuit models provide a good simulation of the real, physical phenomena, especially in the case of high loop gains. To the authors knowledge no publications on the investigation of these equivalent circuit models and the substantial evaluation of the results are found in the literature.

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Summary

Concept and calculation of maximum slewing rate, one of the major hindrances of applying operational amplifiers, are presented. A simplified non-linear circuit model of operational amplifiers helps to give some qualitative and quantitative relationships between the slewing rate and the small signal high frequency characteristics. The theoretical results are in excellent agreement with the recorded diagrams.

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