SOME DEFINITIONS AND INTERPRETATION PROBLEMS OF ELECTRONICS

József ZOLTAI

Department of Measurement and Instrumentation Engineering
Technical University of Budapest
H-1521 Budapest, Hungary

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Abstract

In connection with the writing of a text-book (HAINZMANN - VARGA - ZOLTAI, 1992) it became necessary to make clear some basic notions. The paper selects a few of them, presenting the definition and interpretation of the main quality describing parameters of oscillators and DC voltage regulators.

Keywords: oscillators, voltage regulators, stabilization factors, quality describing parameters.

Introduction

When designing oscillators or voltage regulators it is important to know whether the selected circuit arrangement will correspond to the specified requirements. In this respect the so-called quality describing parameters give good bases if their definition and interpretation is clearly known. In the basic special literature only the definition formulas are sometimes given without a proper interpretation (TSZMK, 1977), or the application of the quality describing parameters is done without the explanation of their nature (HERPY, 1973). This paper tries to make clear both the definition and the interpretation of such quality describing parameters: first the amplitude and frequency stabilization factors of oscillators, then the voltage stabilization quality factor, the output resistance and temperature coefficient of voltage regulators.

Stabilization Factors of Oscillators

Sinusoidal oscillators can be made in many different connection structures. According to a very frequent but not exclusive solution an amplifier is equipped with a negative and a positive feedback. One of them is amplitude dependent, thus taking care of amplitude stabilization, and the other is frequency dependent, thus accomplishing the task of frequency stabilization. In *Fig. 1* the gain of the amplifier is denoted by A, the β is the amplitude dependent negative feedback factor and B denotes the frequency dependent positive feedback factor. The listed transfer factors are generally complex quantities and the general condition of the oscillation can be formulated with them according to the following equation:

$$H = A^* B = \frac{A}{1 + A\beta} B = 1 , \tag{1}$$

where H is the loop-gain in the positive feedback loop consisting of B and A^* and

$$A^* = \frac{A}{1 + A\beta}$$

is the closed loop gain of the amplifier with the negative feedback alone.

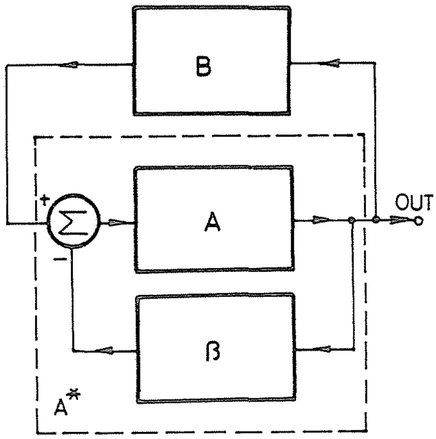

Fig. 1. General block diagram of oscillators

The general condition of the oscillation can be resolved into an amplitude condition:

$$|A^*|\ |B| = 1 \tag{2}$$

and a phase condition:

$$\phi_{A^*} + \phi_B = 0 \pm n2\pi , \tag{3}$$

where $n = 0, 1, 2, \ldots$

Assume that because of some reason (e.g. temperature change, aging, etc.) the magnitude and the phase of the loop gain of the positive feedback loop changes. This will cause amplitude and frequency change in the oscillation. The oscillator can be regarded to be the better, the smaller are the changes of the amplitude and frequency, respectively. The amplitude and frequency stability of the oscillators is characterized by two quality describing parameters:

– the *amplitude stabilization factor*:

$$\boxed{\sigma_a = -\frac{\partial H}{\partial \hat{V}_{\text{out}}}\Big|_{\hat{V}_{\text{out0}}} \hat{V}_{\text{out0}}} \tag{4}$$

– and the *frequency stabilization factor*:

$$\boxed{\sigma_f = -\frac{\partial \phi_H}{\partial \omega}\Big|_{\omega_{\text{osc}}} \omega_{\text{osc}}} \tag{5}$$

In these definitions the output signal of the oscillator has been supposed to be a voltage whose original nominal peak value was $\hat{V}_{\text{out0}}$ and oscillation (angular) frequency was ω_{osc}. The ϕ_H denotes the phase of the loop gain. (The negative signs are included in the definitions to result positive numbers for σ_a and σ_f.)

Consider the simple case, when A and β are frequency independent real transfer factors and B depends on the frequency (it is complex), further let β be amplitude-dependent. The required characteristic of the closed loop gain of the amplifier with the negative feedback is shown by *Fig. 2* (monotonic decrease with increasing amplitude). The figure illustrates that in case of a steeper slope of the gain a given change of the magnitude of the loop gain ΔH causes a smaller amplitude change.

Fig. 3 shows the phase plot of the transfer factor B. This figure illustrates that the steeper the phase response, the smaller the frequency change will be due to some phase error $\Delta\phi_H$ of the loop.

If we look for stability describing parameters in case of changes outside the positive feedback loop, the application of the above definitions (4 and 5) is clumsy. E.g. if the changes are in the branch of the amplifier, the original definitions can be used only by modifying the loop gain: instead of $H = BA^*$ the expression of the loop gain should be taken now $H = (B-\beta)A$, thus the modified definitions of the modified stability describing parameters are

– for the *modified amplitude stabilization factor*:

$$\sigma_a' = -\frac{\mathrm{d}[(B-\beta)A]}{\mathrm{d}\hat{V}_{\text{out}}}\Big|_{\hat{V}_{\text{out0}}} \hat{V}_{\text{out0}} \,, \tag{6}$$

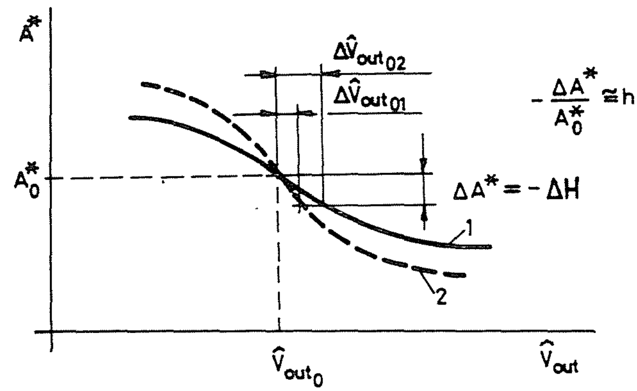

Fig. 2. Illustration to the amplitude stabilization mechanism

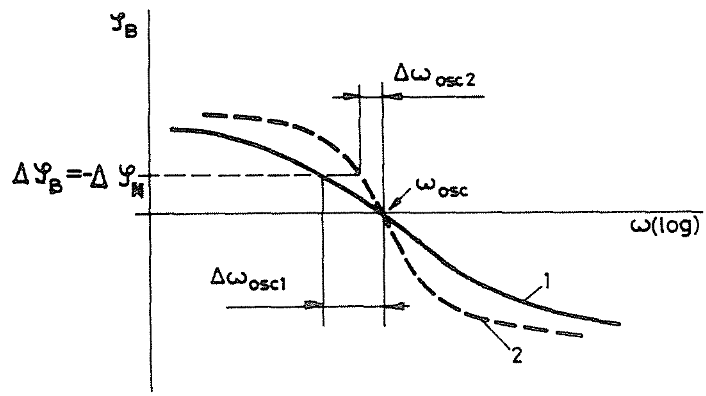

Fig. 3. The effect of the phase slope on the frequency stability

– for the *modified frequency stabilization factor*:

$$\sigma'_f = -\frac{\mathrm{d}\phi_{(B-\beta)A}}{\mathrm{d}\omega}\Big|_{\omega_{osc}} \omega_{osc} \; . \tag{7}$$

It is not possible to apply any definition in the form of the derivative of some loop gain or its phase if changes arising in the branch of the negative feedback are considered.

Introduce in general, for modeling the amplitude error h and the phase error ϕ, an error member with the transfer factor

$$h = (1 + h)e^{j\phi} \; . \tag{8}$$

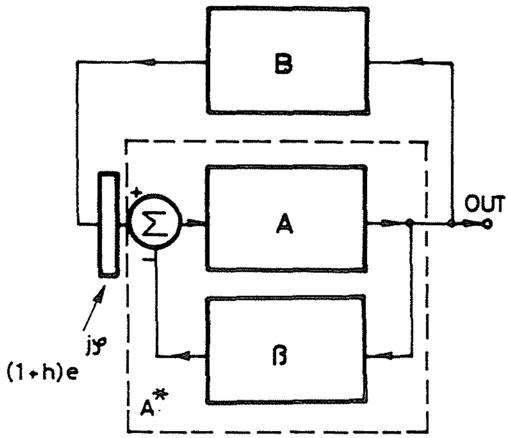

Fig. 4. Inserting an error member into the positive feedback loop

Place the error member first in the positive feedback loop (*Fig. 4*). The general mechanism of amplitude stabilization is this: because of the increased loop gain by $(1+h)$ the amplitude of the oscillation is growing (if $h > 0$), but this growth decreases the amplitude dependent gain A^* of the amplifier with negative feedback, thereby decreasing the loop gain of the positive feedback loop until the condition of oscillation will be fulfilled again. For the relative change $\Delta H/H$ of the loop gain, which decreases to its $(1+h)$th value, the following can be written:

$$H\frac{1}{1+h} = H + \Delta H = H\left(1 + \frac{\Delta H}{H}\right) , \tag{9}$$

thus in case of $|h| \ll 1$:

$$\frac{\Delta H}{H} = -h . \tag{10}$$

As we have seen, the amplitude of the oscillation is the more stable, the smaller is the change of the amplitude belonging to the change of the loop gain, therefore the limit value of the ratio

$$-\frac{\dfrac{\Delta H}{H}}{\dfrac{\Delta \hat{U}_{\text{out}}}{\hat{U}_{\text{out0}}}} \tag{11}$$

was introduced as the definition of σ_a (see *Eq.* (4)). Applying relationship (10), a new and general definition can be formulated, which will be valid without modification for any position of the error member: in the branch of the positive feedback or of the amplifier or of the negative feedback. The general definition is this:

$$\boxed{\sigma_a = \lim_{h\to 0} \frac{h}{\frac{\Delta \hat{V}_{\text{out}}}{\hat{V}_{\text{out}0}}}} \tag{12}$$

The general process of frequency stabilization is similar to that of the amplitude stabilization. The feedback loop produces a phase shift, equal in magnitude to the arisen parasitic phase shift ϕ but opposite signed, as a result of the change of the oscillation frequency, thus the condition of oscillation will be fulfilled again. If the compensating change of the phase shift of the loop gain is $\Delta\phi_H$, then:

$$\Delta\phi_H = -\phi \; . \tag{13}$$

The more stable the frequency, the smaller the necessary frequency change is, therefore the limit value of the ratio

$$-\frac{\Delta\phi_h}{\frac{\Delta\omega_{\text{osc}}}{\omega_{\text{osc}0}}} \tag{14}$$

was introduced as the definition of the frequency stabilization factor (see *Eq.* (5)). Applying relationship (13), a new and general definition can be formulated, which will be valid without modification for any position of the error member. The general definition is this:

$$\boxed{\sigma_f = \lim_{\phi\to 0} \frac{\phi}{\frac{\Delta\omega_{\text{osc}}}{\omega_{\text{osc}0}}}} \tag{15}$$

According to *Fig. 5.a* and *b* the error member can be inserted in the branch of the amplifier or of the negative feedback. In these cases the stabilization process becomes double-staged.

If it is in the branch of the amplifier, in the first stage the negative feedback reduces the errors h and ϕ to their $(1 + H_n)$th value:

$$h' = \frac{h}{1 + H_n} \quad \text{and} \quad \phi' = \frac{\phi}{1 + H_n} \; , \tag{16}$$

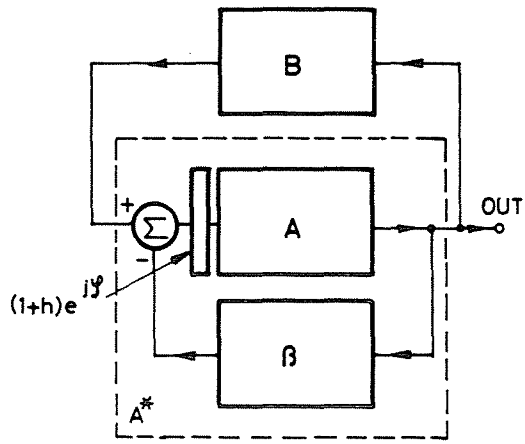

Fig. 5a. The error member in the branch of the amplifier

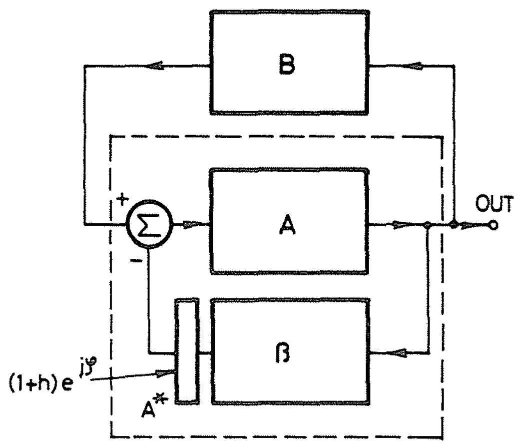

Fig. 5b. Inserting the error member into the negative feedback loop

where H_n is the loop gain of the negative feedback loop ($H_n = A\beta$). In the second stage the h', ϕ' errors are affected by the already described mechanisms of the amplitude and frequency stabilization, respectively. Thus the resulting amplitude and frequency stabilization factors (σ'_a and σ'_f) are $(1 + H_n)$ times greater than the stabilization factors valid for the positive feedback loop alone:

$$\sigma'_a = (1 + H_n)\sigma_a \tag{17}$$

and

$$\sigma'_f = (1 + H_n)\sigma_f \ . \tag{18}$$

The σ'_a and σ'_f can be called stabilization factors belonging to the amplifier itself (they are also called modified stabilization factors).

If the error member is placed in the branch of the negative feedback (*Fig. 5*), the stabilization process is this: the negative feedback has little effect on the amplitude and the phase errors, thus they will be transferred to the amplitude and phase error of A^* almost unchanged (if $|H_n) \gg 1$):

$$h'' = h\frac{H_n}{1+H_n} \cong h \quad \text{and} \quad \phi'' = \phi\frac{H_n}{1+H_n} \cong \phi \ . \tag{19}$$

This means that in such cases the original stabilization factors can be used again ($\sigma''_a \cong \sigma_a$ and $\sigma''_f \cong \sigma_f$).

Quality Characteristics of DC Voltage Regulators

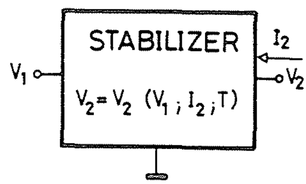

Fig. 6. The V_2 output of a voltage regulator (or stabilizer) depends on the input voltage (V_1), the load current (I_2) and the temperature (T)

As illustrated in *Fig. 6*, the output voltage of a DC voltage regulator (V_2) is a multi-variable function depending on the input voltage (V_1), the load current (I_2) and the temperature (T):

$$V_2 = V_2(V_1, I_2, T) \ . \tag{20}$$

If the listed quantities change, the output voltage also changes. Its total differential is

$$\Delta V_2 = \frac{\partial V_2}{\partial V_1}\Delta V_1 + \frac{\partial V_2}{\partial I_2}\Delta I_2 + \frac{\partial V_2}{\partial T}\Delta T \ . \tag{21}$$

Passing over to relative changes:

$$\frac{\Delta V_2}{V_2} = \underbrace{\frac{\partial V_2}{\partial V_1}\frac{V_2}{V_1}\frac{\Delta V_1}{V_1}}_{\frac{1}{Q_f}} + \underbrace{\frac{\partial V_2}{\partial I_2}\frac{I_2}{V_2}\frac{\Delta I_2}{I_2}}_{R_{ki}\frac{-1}{R_t}} + \underbrace{\frac{\frac{\partial V_2}{\partial T}}{V_2}\Delta T}_{\alpha_{uT}} \ . \tag{22}$$

According to the above relationship the main quality parameters can be defined as follows:

– *Voltage stabilization quality factor*:

$$Q_f = \frac{\partial V_1}{\partial V_2}\frac{V_2}{V_1} \cong \left.\frac{\frac{\Delta V_1}{V_1}}{\frac{\Delta V_2}{V_2}}\right|_{\substack{I_2=\text{constant}\\ T=\text{constant}}} \quad [-] \tag{23}$$

– *Output resistance*:

$$R_{\text{out}} = \frac{\partial V_2}{\partial I_2} \cong \left.\frac{\Delta V_2}{\Delta I_2}\right|_{\substack{V_1=\text{constant}\\ T=\text{constant}}} \quad [\text{ohm}] \tag{24}$$

– *Temperature coefficient*:

$$\alpha_{vT} = \frac{1}{V_2}\frac{\partial V_2}{\partial T} \cong \left.\frac{\frac{\Delta V_2}{V_2}}{\Delta T}\right|_{\substack{V_1=\text{constant}\\ I_2=\text{constant}}} \quad \left[\frac{1}{{}^0C}\right] \tag{25}$$

Our investigations are going to cover these quality parameters, although in the practice other, partly equivalent quality describing parameters are also used: e.g. the line regulation, the ripple rejection, the load regulation, the efficiency, etc. (Analog Devices, 1992).

Keeping in view the connection diagramme of Fig *7.a*, let us draw the ac equivalent circuit of the regulator and by means of this determine the main quality parameters.

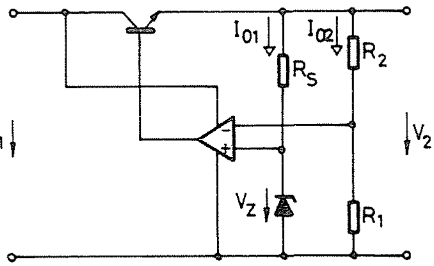

Fig. 7a. Series pass-transistor voltage regulator

In the equivalent circuit of *Fig. 7.b* the transistor is represented by its y-parameter (or g-parameter) model. The differential mode gain of the error amplifier is assumed to be frequency-independent:

$$A_{vd} \neq A_{vd}(f)\,. \tag{26}$$

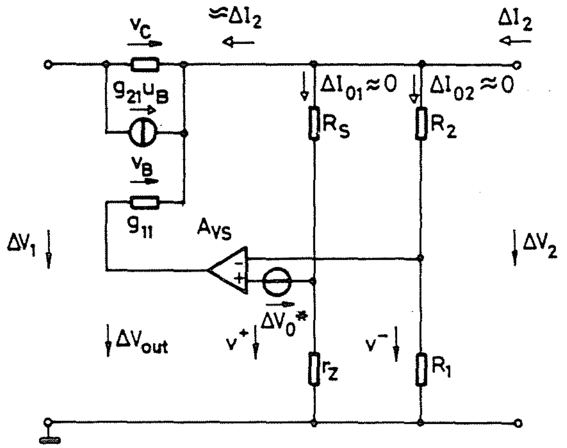

Fig. 7b. AC equivalent circuit of the regulator

Another assumption is that the current change of the Zener diode and the output divider can be neglected:

$$\Delta I_{01} \quad \text{and} \quad \Delta I_{02} \ll \Delta I_2 \ . \tag{27}$$

The voltage source at the input of the error amplifier is modelling three different effects.

– The finite supply voltage rejection ratio of the error amplifier. Since the voltage change ΔV_1 is also the supply voltage change, this component can be expressed by

$$S_{SV}\Delta U_1 \ , \tag{28}$$

where S_{SV} is the sensitivity to supply voltage changes (it is the reciprocal value of SVRR, the supply voltage rejection ratio).

– The offset voltage of the error amplifier (ΔV_0), assuming that at some reference temperature offset zero adjustment has been made, thus the offset is the consequence of a temperature change. E.g. if the current drift is negligible ΔV_0 can be calculated with the voltage drift v_d:

$$\Delta V_0 = v_d \Delta T \ . \tag{29}$$

– The temperature dependence of the Zener voltage:

$$\Delta V_z = V_z \alpha_{vz} \Delta T \ . \tag{30}$$

With all these quantities:

$$\Delta V_0^* = S_{SV}\Delta V_1 + (v_d + V_z\alpha_{vz})\Delta T \ . \tag{31}$$

The output voltage change of the error amplifier is:

$$\Delta V_{\text{out}} = A_{vd}(v^+ - v^-) = A_{vd}\left(\Delta V_0^* + \Delta V_2\left[\frac{r_z}{r_z + R_s} - \frac{R_1}{R_1 + R_2}\right]\right) = $$
$$= A_{vd}\Delta V_0^* - A_{vd}\beta_v\Delta V_2 \ , \qquad (32)$$

where the feedback transfer factor was introduced:

$$\beta_v = \frac{R_1}{R_1 + R_1} - \frac{r_z}{r_z + R_s} \ . \qquad (33)$$

The following loop equations can be written:

$$v_c = \Delta V_1 - \Delta V_1 \qquad (34)$$

and

$$v_b = \Delta V_{\text{out}} - \Delta V_2 \ . \qquad (35)$$

The change of the output current:

$$-\Delta I_2 = v_B(g_{11} + g_{21}) + v_C g_{22} \ . \qquad (36)$$

Considering that $g_{11} \ll g_{21}$; $1 \ll A_{vd}\beta_v$ and $g_{22} \ll g_{21}A_{vd}\beta_v$, rearranging *Eqs.* (28) . . . (36) we get:

$$\Delta V_2 = \left(\frac{g_{22}}{A_{vd}\beta_v g_{21}} + \frac{S_{SV}}{\beta_v}\right)\Delta V_1 + \frac{1}{g_{21}A_{vd}\beta_v}\Delta I_2 + \frac{v_d + \alpha_{vz}V_z}{\beta_v} \ . \qquad (37)$$

Comparing this to the general formula of ΔV_2, expressed from the *Eq.* (22):

$$\Delta V_2 = \frac{1}{Q_f}\frac{V_2}{V_1}\Delta V_1 + R_{\text{out}}\Delta I_2 + \alpha_{vT}V_2\Delta T \ , \qquad (38)$$

the calculation formulas of the quality parameters can be read out:

$$\frac{1}{Q_f} = \frac{V_1}{V_2}\left(\frac{g_{22}}{A_{vd}\beta_v g_{21}} + \frac{S_{SV}}{\beta_v}\right) \ , \qquad (39)$$

$$R_{\text{out}} = \frac{1}{g_{21}A_{vd}\beta_v} \qquad (40)$$

and

$$\alpha_{vT} = \frac{u_d + \alpha_{vz}V_z}{\beta_v V_2} \ . \qquad (41)$$

Transforming the equation of R_{out} a little considering that $A_{vd}\beta_v = H$, the loop-gain of the error amplifier, we get:

$$R_{\text{out}} = \frac{\frac{1}{g_{21}}}{H} \cong \frac{R_{\text{out}}^{FC}}{1+H}, \tag{42}$$

which proves that the output resistance of the regulator is in connection with the output resistance of the common collector stage formed by the series pass-transistor. If the base of the transistor were grounded, its output resistance would be

$$R_{\text{out}}^{FC} = \frac{1}{g_{21}}, \tag{43}$$

but this is reduced by the error amplifier's voltage feedback to its $(1+H)$th value (since $H \gg 1$, $H \cong 1+H$).

Recollecting the definition of Q_f:

$$Q_f = \frac{\frac{\Delta V_1}{V_1}}{\frac{\Delta V_2}{V_2}} = \frac{V_2 \Delta V_1}{V_1 \Delta V_2}, \tag{44}$$

that is

$$\frac{1}{Q_f} = \frac{U_1}{U_2}\frac{\Delta U_2}{\Delta U_1}, \tag{45}$$

it is evident that the term between the braces on the right side of equation (39) gives $\Delta V_2/\Delta V_1$. The first part of that term after some transformation:

$$\frac{g_{22}}{A_{vd}\beta_v g_{21}} = \frac{\frac{1}{A_{vd}\beta_v g_{21}}}{\frac{1}{g_{22}}} \cong \frac{\frac{1}{A_{vd}\beta_v g_{21}}}{\frac{1}{g_{22}} + \frac{1}{A_{vd}\beta_v g_{21}}} = \frac{R_{\text{out}}}{\frac{1}{g_{22}} + R_{\text{out}}}, \tag{46}$$

since

$$\frac{1}{g_{22}} \gg \frac{1}{A_{vd}\beta_v g_{21}} = \frac{1}{H g_{21}}. \tag{47}$$

According to this, the regulator allows the input voltage changes to appear at the output like an attenuator, where the parts of the attenuator are the collector-emitter resistance of the pass transistor $1/g_{22}$ and the output resistance of the regulator (see *Fig. 8*).

The second part of the term between the braces in *Eq.* (39) can be interpreted after a little rearrangement:

$$\frac{S_{SV}}{\beta_v} = \frac{S_{SV} A_{vd}}{\beta_v A_{vd}} = \frac{\frac{\Delta V_{\text{out}}^*}{\Delta V_1}}{H} \cong \frac{\frac{\Delta V_{\text{out}}^*}{\Delta V_1}}{1+H}, \tag{48}$$

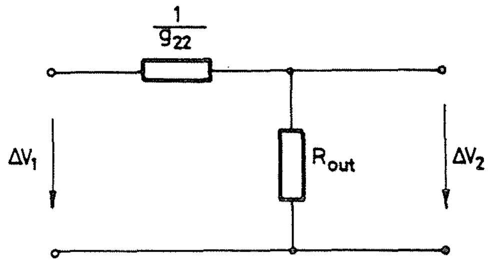

Fig. 8. For input voltage changes the regulator behaves like an attenuator

where $S_{SV}A_{vd}\Delta V_1 = \Delta V^*_{\text{out}}$ is, corresponding to the definition of the supply voltage rejection ratio, the error signal at the output of the error amplifier due to an input voltage change of ΔV_1. *Eq.* (48) expresses that its relative value $\Delta V^*_{\text{out}}/\Delta V_1$ is reduced by the negative feedback to an extent in accordance with the measure of the feedback $(1 + H)$.

Since in the *Eq.* (41)

$$\beta_v V_2 = \left(\frac{R_1}{R_1 + R_2} - \frac{r_z}{r_z + R_s}\right) V_2 \cong \frac{R_1}{R_1 + R_2} V_2 = V_z \; , \tag{49}$$

it can be rewritten like this:

$$\alpha_{vT} = \frac{v_d}{V_z} + \alpha_{vz} \; , \tag{50}$$

meaning that the temperature dependence of the reference voltage (α_{vz}) will be transferred into the temperature dependence of the regulator and the drift voltage of the error amplifier contributes to this with its relative value in terms of the Zener voltage.

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