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STRUCTURAL METHODS FOR INCREASING THE SPEED OF OPERATIONAL AMPLIFIERS

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Structural methods for improving the speed-of-response of operation amplifiers (OA) are discussed along with ways to reduce the settling time of an OA with a parallel negative feedback loop. Promising block diagrams for high-speed OA's are analyzed, specifically for OA current-to-voltage converters, and their maximal potentialities are estimated with regard to increasing the cutoff frequency and reducing the settling time of their output voltage.

Operational amplifiers (OA) are employed widely in a great variety of arrangements. In a number of cases the dynamic properties of an OA determine the speed-of-response of the equipment in which they are utilized.

The most important parameter of an OA, which characterizes its speed-of-response, is the settling time τ_δ of the output voltage as determined with a step-type input signal. It is helpful to minimize τ_δ in practically all OA applications, but especially in that broad class of arrangements such as converters for the type of data (digital-analog and analog-digital) and commutators.

In this connection the objective of the present article is to analyze new OA block diagrams that can reduce the settling time as well as to estimate their maximal speed-of-response potentialities.

If it is assumed that the transfer function for an open feedback loop $K_0(s)$ (s is a complex variable) corresponds to a first-order inertial element, then the transient process in the OA for a step-type input signal decays exponentially with an equivalent time constant of $1/\omega_{co}$ (the cutoff frequency is determined from the condition $|K_0(j\omega_{co})| = 1$), and the settling time for a given relative error is equal to

$$\tau_\delta = \frac{1}{\omega_{co}} \ln \frac{1}{\delta}.$$

It is evident that to reduce τ_δ in this case one must increase ω_{co}. In a real amplifier circuit the maximal value of ω_{co} is limited by parasitic high-frequency poles ω_{pi} that occur in $K_0(j\omega)$ on account of nonideal amplifier elements, a load capacity, and mounting capacities. If the values of ω_{pi} are remote enough from ω_{co}, then they can be replaced by a single equivalent pole

$$\omega_p \approx \left(\sum_{i=1}^{n} \frac{1}{\omega_{pi}}\right)^{-1},$$

that introduces at the frequency ω_{co} the same phase shift as the actual set ω_{pi} and also affects the transient process in approximately the same way.

With ω_p taken into account the transient process will have the form

$$a(t) = \frac{U_{out}(t)}{U_{out}(\infty)} = 1 + h_1 e^{s_1 t} + h_2 e^{s_2 t}, \tag{1}$$

where

$$s_{1,2} = -\frac{\omega_{co} k_1}{2}\left(1 \pm \sqrt{1 - \frac{4}{k_1}}\right), \quad k_1 = \frac{\omega_p}{\omega_{co}},$$

$$h_1 = -\frac{s_2}{s_2 - s_1}, \quad h_2 = \frac{s_1}{s_2 - s_1}.$$

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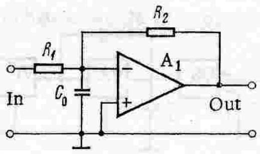

Fig. 1

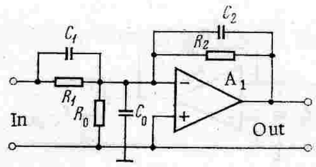

Fig. 2

It follows from Eq. (1) that if $k_1 \gg 4$, then

$$a(t) \approx 1-\left(1-\frac{2}{k_1}\right)^{-1}\left\{\left(1-\frac{1}{k_1}\right)\exp\left[-\omega_{co}\left(1+\frac{1}{k_1}\right)t\right]-\frac{1}{k_1}\exp\left[-k_1\omega_{co}\left(1-\frac{1}{k_1}\right)t\right]\right\},$$

and τ_δ is even somewhat smaller than in the case of the ideal element of the first order with the same ω_{co}. It can be shown that τ_δ has a minimum when $k_1 = 4$, where $s_1 = s_2 = -2\omega_{co}$ and

$$a(t) = 1-(1+2\omega_{co}t)\exp(-2\omega_{co}t).$$

Here τ_δ for $\delta = (1.0$ to $0.01)\%$ is about 1.5 times less than when $\omega_p \to \infty$.

When $k_1 < 4$, an oscillatory transient process occurs, but when $k_1 \geq 2$, the value of τ_δ does not exceed the values for $\omega_p \to \infty$. A reduction of k_1 to a value less than 2 results in an abrupt increase of the oscillation and the value of τ_δ. Considering these characteristics and also the need to have some margin it is advisable to choose $\omega_{co} = \omega_p/4$. In this case it is permissible to have a load capacity such that ω_p is reduced by half.

Besides the existence of a pole ω_p there may be in the transfer function of a real amplifier what are known as dipoles. A dipole represents a pair composed of a zero and a nearby pole that occur on account of a mismatching error in the correcting networks. In this case if the frequency mismatch of a zero (ω_i') and a pole (ω_i) is small and their frequencies are many times less than ω_{co}, then the transient process can be approximated by the following relation:*

$$a(t) \approx \left(1+\sum_{i=1}^{n}\frac{\omega_i-\omega_i'}{\omega}\right)^{-1}\left\{[1-\exp(-\omega_{co}t)]+\sum_{i=1}^{n}\frac{\omega_i-\omega_i'}{\omega_{co}}[1-\exp(-\omega_i t)]\right\}. \tag{2}$$

From Eq. (2) it follows that at first $a(t)$ is rapidly damped with a small time constant $1/\omega_{co}$ owing to the first term with the square brackets, and then the damping is slowed down abruptly owing to the large time constants $1/\omega_i$.

However, if

$$\left(1+\sum_{i=1}^{n}\frac{\omega_i-\omega_i'}{\omega_{co}}\right)^{-1}\sum_{i=1}^{n}\frac{\omega_i-\omega_i'}{\omega_{co}} \leqslant \delta, \tag{3}$$

then the presence of the dipole has practically no effect on the value of τ_δ. In the opposite case τ_δ increases sharply.

As seen from Eqs. (1) and (2), in order to achieve the maximal speed-of-response it is good practice to synthesize the OA circuit with the feedback network taken into account so that $K_0(j\omega)$ has an attenuation of 20 dB/decade over the frequency range from $\delta\omega_{co}$ to $4\omega_{co}$. In this case one can have dipoles in the low-frequency region when the condition of Eq. (3) is met.

The greatest difficulty in realizing such a transfer function when using an OA in circuits with parallel feedback (Fig. 1) occurs because the capacity C_0 at the OA's input results in the appearance of a relatively low-frequency pole $1/T_0$, where $T_0 = C_0R_1R_2/(R_1 + R_2)$, in $K_0(j\omega)$. Here, in order to avoid an oscillatory transient process one must sharply reduce ω_{co} to a value approximately equal to $4/T_0$ and be satisfied with $\tau_\delta \geq 40T_0$ ($\delta = 0.01\%$). For example, if $R_1 = R_2 = 10$ kΩ and $C_0 = 30$ pF, then $\tau_\delta = 6$ μsec. To reduce τ_δ one must either reduce R_2 (which is not efficient) or employ correcting sections in the feedback network, as shown in Fig. 2, which provides for the equality $C_0R_0 = C_1R_1 = C_2R_2$. In this case it is quite possible to compensate fully for the pole, but the matching must be very accurate, compensation is achieved only for a fixed transfer factor ($K_p = R_2/R_1$), it is impractical for a current signal (for example, in a digital-analog converter), and, besides this, the introduction of R_0 increases the zero drift.

* The error in a determination of τ_δ by this formula for systems of an order no greater than five does not exceed 7%.

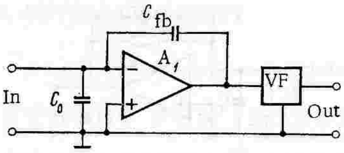

Fig. 3

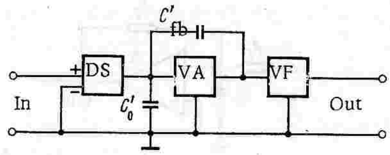

Fig. 4

To overcome these difficulties a specialized OA can be assumed that has a very low input impedance in the high-frequency region, i.e., that operates as a current-to-voltage converter (OA-CV) and not as a voltage-to-voltage converter like the usual OA. In the ideal case the OA-CV should have input (z_{in}) and output (z_{out}) impedances that are equal to zero with an infinitely high transfer impedance $Z_{tr} = U_{out}/I_{in}$ and an unlimited frequency range. It is not difficult to see that the transfer function for the open feedback loop of such an OA-CV does not depend on C_0 and the transfer factor of the computing amplifier since the shunting action on z_{in} of the resistance R_1 and of the capacity C_0 at the input is negligible.

We will consider two possible design principles for an OA-CV, and we will estimate their potentialities.

One of the possible block diagrams for an OA-CV is shown in Fig. 3. It consists of an input inverting amplifier A_1 supplied with feedback through C_{fb}, and an output voltage follower VF. If it is assumed that in the frequency range under discussion the transfer function of the VF is unity and the transfer function of A_1 is equal to

$$K_1(j\omega)=k_0\left(1+j\frac{\omega}{\omega_1}k_0\right)^{-1}$$

(where ω_1 is the frequency of unity gain for A_1), then when ω_1/k_0

$$z_{in}\approx\frac{1}{\omega_1 C_{fb}\left(1+j\frac{\omega}{\omega_1}\right)}.$$

From this it is seen that over a frequency range from about $4\omega_1/k_0$ to $\omega_1/4$ it can be said that $z_{in} = R_{in} = 1/\omega_1 C_{fb}$. With a large ω_1 the value of R_{in} can be made very small. For example, with $C_{fb} = 2$ pF and $f_1 = \omega_1/2\pi = 800$ MHz we will obtain $R_{in} \approx 100\ \Omega$. If such an OA-CV operates in the circuit of Fig. 1, then its frequency $\omega_{co} = 1/C_{fb}R_2$, and the pole due to C_0 corresponds to $\omega_p = \omega_1 C_{fb}/(C_{fb} + C_0)$. It is advisable to make the maximal value of ω_{co}, as noted above, equal to $\omega_p/4$, i.e.,

$$\frac{1}{C_{fb}R_2}=\frac{\omega_1 C_{fb}}{4(C_{fb}+C_0)}.$$

When R_2, C_0, and ω_1 are given, it is necessary to make

$$C_{fb}=\frac{2}{\omega_1 R_2}\left(1+\sqrt{1+\omega_1 C_0 R_2}\right), \tag{4}$$

which provides for

$$\omega_{co}=\frac{\omega_1}{2\left(1+\sqrt{1+\omega_1 C_0 R_2}\right)}, \tag{5}$$

$$\tau_\delta=\frac{2}{\omega_1}\left(1+\sqrt{1+\omega_1 C_0 R_2}\right)\ln\frac{1}{\delta}. \tag{6}$$

Compared with the ordinary circuit without C_{fb} the value of ω_{co} is increased and the value of τ_δ is decreased by the factor $2\omega_1 C_0 R_1 R_2/(R_1 + R_2)(1 + \sqrt{1+\omega_1 C_0 R_2})$. As an example, when $R_1 = R_2 = 10$ kΩ and $C_0 = 30$ pF, $f_1 = 800$ MHz without C_{fb} and it is found that $f_{co} \approx 0.28$ MHz with $\tau_\delta \approx 6$ μsec ($\delta = 0.01\%$), but with $C_{fb} = 1.6$ pF we obtain $f_{co} \approx 10$ MHz and $\tau_\delta = 160$ nsec, i.e., characteristics that are about 38 times better.

The transfer function of a real amplifier A_1 inevitably contains a high-frequency pole $\omega_{p1} < \omega_1$. As has been demonstrated by modeling, the frequency of unity gain for the contour formed by the amplifier A_1 and a capacitor C_{fb}, which is equal to

$$\omega_{p1}=\frac{\omega_1 C_{fb}}{C_{fb}+C_0}+4\omega_{co}$$

should not exceed $0.5\omega_{p1}$ (otherwise it becomes too oscillatory). Therefore, with Eq. (5) taken into account we find that if

$$\omega_{p1} \geqslant 8\omega_{co} = \frac{4\omega_1}{1+\sqrt{1+\omega_1 C_0 R_2}}, \tag{7}$$

then the relations presented above, especially Eq. (6), hold well. With smaller values of ω_{p1} (for the example presented with $f_p < 80$ MHz) the values of C_{fb} or R_2 should be increased so that $C_{fb}R_2 = 8/\omega_{p1}$. In this case

$$\omega_{co} = \omega_{p1}/8, \quad \tau_\delta = \frac{8}{\omega_{p1}} \ln\frac{1}{\delta}. \tag{8}$$

The amplifier A_1 can be realized in accord with the block diagram shown in Fig. 4, where DS is a differential stage having a high-impedance output, VA is a voltage amplifier composed of an emitter follower on the input and a common-emitter push-pull stage, VF is a voltage follower, C_0' is the sum of the output capacity of the DS and the output of the VA, while C_{fb}' is the capacity of the local feedback.

Keeping in mind that the amplifier's input capacity should be a part of C_0, then DS ought to have a minimal input capacity and for this a cascaded structure is advisable or the use of input repeaters with the DS.

As a calculation shows and as confirmed by an experimental study, when transistors having $f_T = 1.5$ GHz and $C_c = 1$ pF are used in A_1, it is possible to provide $f_{p1} = \omega_{p1}/2$ around 1600 MHz. In this case the maximal potentialities of an OA-CV with such a structure where $R_1 = R_2 = 5$ kΩ and $C_{fb} = 1.6$ pF are characterized by the following values: $f_{co} \approx 20$ MHz and $\tau_\delta = 56$ nsec for $\delta = 0.1\%$, and $\tau_\delta = 80$ nsec for $\delta = 0.01\%$.

A virtue of this kind of OA-CV is the feasibility – owing to the small value of R_{in} – of changing K_p by R_1 (i.e., the ratio R_2/R_1) with practically no increase of τ_δ. The minimal value of R_1 can be made equal to

$$R_1 = 4R_{in} \approx \frac{4}{\omega_1 C_{fb}'} = \frac{2R_2}{1+\sqrt{1+C_0 R_2 \omega_1}}.$$

Therefore

$$0 \leqslant K_p \leqslant (1+\sqrt{1+\omega_1 C_0 R_2})/2.$$

For this example K_p can be chosen in the range from 0 to 20 while keeping $\tau_\delta \approx 56$ nsec. (Remember that when K_p is increased in the ordinary OA's, an increase of τ_δ proportional to $1 + K_p$ is unavoidable.)

It is permissible to change C_0 to within certain limits. If C_0 is doubled versus the original value assumed for a calculation, then an oscillation will appear in the transient process owing to the proximity of ω_{01} to ω_{co}, but (for $\delta \le 1\%$) τ_δ has practically no increase. A reduction in C_0 also has an insignificant effect on τ_δ when the conditions of the inequality (7) are fulfilled, but in this case the frequency ω_{01} will increase. When there is a pole ω_{p1} in the transfer function $K_1(j\omega)$ smaller than ω_1, the maximal value of ω_{01} should not exceed $0.5\omega_{p1}$. Considering the restrictions indicated, the value of C_0 must be chosen in the interval

$$2C_{0,0} \geqslant C_0 \geqslant C_{fb}\left(\frac{2\omega_1}{\omega_{p1}} - 1\right),$$

where $C_{0,0}$ is the initial value of C_0 adopted when calculating C_{fb}.

We note that the introduction of C_{fb} from the output of an OA without a VF reduces the maximal potentialities of the circuit because in this case the load capacity on the OA's output results in a reduction of the frequency band, not only in the contour formed by the amplifier and basic feedback but in the contour with C_{fb}, thereby making it more oscillatory and requiring a reduction in ω_{co} and an increase in τ_δ.

Another design principle for an OA-CV was proposed in [1], namely, to connect on the input of the OA (Fig. 5) a current follower (CF) (loaded by $R_0' \approx R_2$) that has a very small input impedance achieved by using a common-base type of stage as shown in Fig. 6, where A_1 is an auxiliary OA which forms a low-frequency channel that is needed to obtain a small zero drift.

The transfer function of the CF per the circuit of Fig. 6 for the conditions $\omega_{11}C_1'R_1 \gg 1$, $R_1' \gg r_e'$, and $\omega_{11}C_1'R_Bk_0 \gg 1$ can be represented in the following form:

$$K_{cf}(j\omega) = \frac{\alpha z_0' R_0 k_0 (1+j\omega/\omega_1')(1+j\omega/\omega_2')}{(R_1'+r_e')(R_0+R_2)(1+j\omega/\omega_1)(1+j\omega/\omega_2)} \tag{9}$$

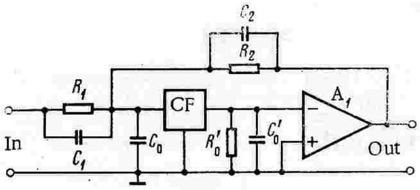

Fig. 5

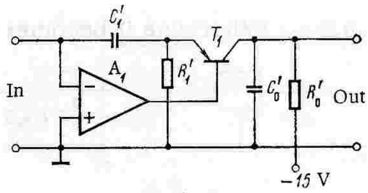

Fig. 6

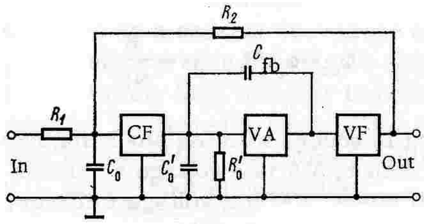

Fig. 7

where $R_0 = R_1R_{in1}/(R_1 + R_{in1})$; k_0, R_{in1}, and ω_{11} are, respectively, the gain, input resistance, and frequency for unity gain of A_1; α and r_e' are, correspondingly, the current gain and the input resistance of transistor T_1 in a common-base circuit,

$$\omega_1' \approx \frac{1}{C_1'R_1'}, \quad \omega_1 \approx \frac{1}{(C_1'R_B+1/\omega_{11})k_0}, \quad z_0' = \frac{R_0'}{1+j\omega C_0'R_0'}$$

$$R_B = R_0R_2/(R_0+R_2), \quad \omega_2' \approx \omega_{11}(1-1/C_1'R_1'\omega_{11}),$$

$$\omega_2 \approx \omega_{11}(1+1/\omega_{11}C_1'R_B)/(1+r_e'/R_B).$$

A disadvantage of the circuit in Fig. 5 is that the capacity C_0 causes a pole to appear in $K_0(j\omega)$ at the frequency $1/R_0'C_0'$. To compensate for it, it is necessary to employ C_1 and C_2 which must be matched with C_0 as was done in the circuit of Fig. 2. This can be avoided by the block diagram shown in Fig. 7 (the designations are similar to Figs. 4 and 5, but $R_0' \gg R_2$).

The transfer function for the contour R_2 of the amplifier in the circuit of Fig. 7 can be obtained if in Eq. (9) z_0' is replaced by the equivalent load resistance and $K_{CF}(j\omega)$ is multiplied by the transfer function of the output portion of the circuit, then

$$K_0(j\omega) \approx \frac{\alpha R_0'R_0R_2k_0(1+j\omega/\omega_1')(1+j\omega/\omega_2')}{(R_0+R_2)(R_1'+r_e')R_1'(1+j\omega/\omega_1)(1+j\omega/\omega_2)(1+j\omega/\omega_3)(1+j\omega/\omega_4)},$$

where $\omega_3 \approx 1/R_0'C_{fb}'k_2$; R_0' is the equivalent resistance formed by the parallel connection of R_{out} of the CF and R_{in} of the VA; k_2 is the gain of the VA at dc, $\omega_4 \approx \omega_{12}C_0'/(C_0' + C_{fb})$, and ω_{12} is the frequency of unity gain for the VA.

To provide the minimal value of τ_δ, i.e., to meet the requirements presented above for $K_0(j\omega)$, it is necessary to choose parameters such that $\omega_1' \approx \omega_3$, $\omega_2' \approx \omega_2 \approx \omega_{11} \approx 1/C_1'r_e'$, and $\omega_4 \geq 4\omega_{co}$ (ω_{co} is the cutoff frequency for the basic feedback circuit through R_2). In addition, so that the dipoles occurring because of a mismatch in these frequencies should not increase τ_δ, it is advisable to ensure that $\omega_3 \ll \omega_{co}\delta$, and that the error in choosing ω_{11} is no greater than $\delta\omega_{11} \leq \delta\,\omega_{co}/\omega_{11}(1 + R_B/r_e')$.

The computation of the optimal value for C_{fb} (that provides the lowest value of τ_δ) involves expressions similar to Eqs. (4)-(6) and (8). But for those same values of ω_{12} and ω_{p1} a value of τ_δ can be obtained which is half that for the structure of Fig. 4. An advantage of the structure in Fig. 7 (versus that of Fig. 4) is the possibility of varying R_1 and C_0 within wider limits. The maximal value of R_1 here is not restricted, but the minimal should be no less than $3r_e'$. With $R_2 = 10$ kΩ it is possible to have a maximal value of $K_p = 100$ ($R_1 = 100\,\Omega$) without increasing τ_δ. The value of C_0 can be varied from zero to a value C_{0max} that is limited by the resonance effect on the input network which is due to the equivalent inductance of the input impedance ($z_{in.\,CF}$) of the CF (a common-base stage). It can be shown that at a high frequency

$$z_{in.\,CF} \approx r_e'(1+j\omega L/r_e'),$$

where $L = r_b/\omega_T$, ω_T being the frequency of unity current gain for the transistor in the CF. In this case the introduction of C_0 is equivalent to the appearance of an added pole in $K_0(j\omega)$ at the frequency

$$\omega_p \approx \sqrt{\frac{r_e'^2}{4L^2}+\frac{1}{LC_0}}-\frac{r_e'}{2L}=\frac{r_e'}{2L}\left(\sqrt{1+\frac{4L}{C_0 r_e'^2}}-1\right).$$

Granting this, it is possible to find C_{0max} from the following condition: $\omega_p = 4\omega_{co}$ (ω_{co} is the cutoff frequency of the amplifier over the basic feedback loop through R_2), i.e.,

$$C_{0max} \leqslant \frac{1}{4\omega_{co}(4\omega_{co}L+r_e')}.$$

For $f_T = \omega_T/2\pi = 1.5$ GHz, $r_b = 60\Omega$, $r_e' = 30\Omega$, and $f_{co} = 16$ MHz it is permissible to have $C_0 = 0$ to 74 pF. In this case with $R_2 = 5$ kΩ and $R_1 = (0.1$ to $100)$ kΩ we will have $\tau_\delta \approx 90$ to 100 nsec ($\delta = 0.01\%$).

These estimates were obtained on the assumption that all OA stages operate in the linear region during transient processes. It is usually reckoned that during a step in the input signal (i.e., for a signal having an infinite buildup rate) an amplifier of necessity operates in the nonlinear region because the maximal rate V_m of the signal variation at its output is limited. Of course in such a case τ_δ is increased significantly.

However, in reality the maximal rate of change with a step on the input does not exceed the value

$$I_{in}/C_{fb} = U_{out.max}/R_2C_{fb} = U_{out.max}\omega_{co}$$

(it is assumed that $U_{out}(\infty) \leq U_{out.max}$). Therefore, if an OA's circuit is designed so that its V_m exceeds or equals $U_{out.max}\omega_{co}$, then there is no restriction on the rate and no increase in τ_δ. As indicated by computation and confirmed experimentally, the realization of the circuit in Fig. 7 using push-pull amplifier stages with transistors having $C_c \leq 1.0$ pF and $f_T \geq 1.5$ GHz when the total current consumption is 10 to 15 mA can provide a $V_m \approx$ (2000 to 3000) V/μsec. In this case the OA's operation has no limit by V_m: with $U_{out.max} = 10$ V, f_{co} is up to 50 MHz, i.e., for $\tau_\delta = 21$ nsec ($\delta = 0.1\%$) and $\tau_\delta = 30$ nsec ($\delta = 0.01\%$). These values are typical of the maximal potentialities for the speed-of-response of an OA for which the block diagram is given in Fig. 7.

CONCLUSION

For circuits with parallel negative feedback in which there is a substantial capacity at the OA's inverting input it is advisable to employ special-purpose amplifiers that are designed to be current-to-voltage converters. It is then possible to increase the cutoff frequency and reduce the settling time by an order of magnitude in comparison with the usual OA voltage-to-voltage converters.

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