In this paper the frequency stability of small-signal high-speed amplifier circuits using Bode criterion is analysed theoretically. In particular, the inverting and non-inverting amplifiers employing current-feedback operational amplifiers are under review. Based on the analysis of the operational principle, the equations for complex transfer functions of both circuits and formulas for the related electrical parameters are obtained. Moreover, using these formulas recommendations for a stable operation are given. As well, design procedure of the amplifiers with resistive and capacitive load is suggested. The efficiency of the proposed procedure and recommendations are verified by simulation modelling and experimental testing of sample electronic circuits.

Keywords: analogue circuits; high-speed amplifiers; operational amplifiers; CFOA; stability analysis; Bode plots; frequency response; analogue simulation

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1. Introduction

The small-signal high-speed (with bandwidth > 1 MHz) amplifiers are essential building blocks for video amplifiers, RF/IF amplifiers, high-speed A/D drivers and D/A buffers (Analog Dev., MT-060, 2008; Jung, 2002; Tietze & Schenk, 2008). In the past 10 years current-feedback operational amplifiers (CFOAs) have been basically used as active building blocks for design of high-speed amplifiers. As a kind of the monolithic operational amplifiers (op amps) family, the CFOAs have been realized to overcome the finite gain-bandwidth product of the conventional voltage-feedback operational amplifiers (VFOAs) (Jassim, 2013; Seifart, 2003). However, the CFOA-based amplifier circuits are less understood and documented in comparison with the amplifiers using VFOAs.

Stability analysis is an essential part in the design process of the analogue circuits, especially for the high-speed amplifiers, that use CFOAs. The practical interpretation of the sustainability definition is: ‘An amplifier circuit is stable if all voltages and currents are reduced to zero when the input voltages and currents are zero’. Otherwise, in lack of an input signal unexpected oscillations (or self-oscillations) can occur at the output, which is unacceptable. In the theory of electronic circuits, various criteria of stability are designed. For the analysis of analogue circuits, the most commonly used are the criteria of Nyquist (Seifart, 2003) and Bode (Laker & Sansen, 1994; Nagaria, Gopi Krishna, & Singh Rakesh, 2008).

In the analysis of the Bode criteria, the behaviour of the magnitude and phase frequency characteristics or Bode plots are investigated. The Bode plots can be drawn directly from experimental data or computer simulations.

For the amplifier circuits it is assumed that the open-loop transfer function of the CFOAs is stable and their logarithmic a.c. transfer characteristic monotonously decreased with an increase in the frequency ω of the input signal. For this case the Bode criterion says: the closed-loop system containing op amp with negative feedback is stable only if the a.c. transfer characteristic of the open-loop system crosses the x-axis (0 dB) before the linear phase characteristic has reached −180°.

A relatively large number of books, publications and company application reports are devoted to the theory and the design of the amplifier circuits employing CFOAs (Jassim, 2013; Jung, 2002; Kamath, 2014; Mancini, 2001, 2002; Palumbo, 1997; Pandiev, 2012; Safari & Azhari, 2012; Schmid, 2003; Seifart, 2003; Texas Instruments, 2002; Tietze & Schenk, 2008). The authors’ attention is focused on the structure and principle of operation at d.c. and large frequency range (Jung, 2002; Safari & Azhari, 2012; Seifart, 2003; Tietze & Schenk, 2008). However, for the analysis of basic amplifier circuits a simplified model of the CFOA is used. The attention of the author in Jassim (2013) is focused on the design technique, employing CMOS CFOA. Some results on the behaviour of the non-inverting amplifier are given, which also confirm the
The organization of the paper is as follows: in Section 2 the structure and relation between the input and the output voltages and currents for a linear high frequency model of the CFOA used in this work are presented; in Sections 3 and 4 the principles of operation of the inverting and non-inverting amplifiers at low and high frequencies are described; also in Section 4 recommendations for improving the frequency stability are defined, based on the obtained results; the proposed design procedure is given in Section 5; to illustrate the proposed theoretical analyses and the procedure, in Section 6 examples of studying the frequency stability of the inverting and non-inverting amplifiers at several voltage gains and various CFOAs are given. Finally in Section 7, the concluding remarks are given.

2. Current-feedback operational amplifiers

The most common CFOA (or transimpedance amplifier) is equivalent to a (positive second-generation current conveyor) CCII + plus an output voltage buffer. These op amps have a high impedance non-inverting input, a low-impedance inverting input \( x \), a current output \( z \) and a voltage output \( o \). In some of the CFOAs the port \( z \), between the first stage (CCII+) and the second stage (voltage follower), is defined as an external pin. The port \( o \) is the output of the voltage buffer, where the output resistance \( r_o \) is very low (several ohms magnitude). The linear model of the CFOA used in this work is presented in Figure 1.

Figure 1. A linear model of the real CFOA.
3. Principle of operation and basic analysis  
(approximately up to 50 MHz)

The objects of study are the inverting and non-inverting amplifier circuits, employing CFOAs.

3.1. An inverting amplifier

The schematic structure of the high-speed inverting amplifier, employing CFOA, is shown in Figure 2. In this circuit is introduced a parallel negative feedback through the resistors \( R_T \) and \( R_N \). The resistor \( R_F \) is used for compensation of the input bias current of the CFOA. If the circuit in Figure 2 is used as a video line driver, the best frequency response can be obtained by the addition of small resistances \( R \) and \( C \) in, \( + \) and \( - \) respectively and \( C \) out, \( + \) and \( - \) respectively.

In this circuit \( C_{ph} \) is an example (Figure 2) is simplified. Including the external elements, we obtain the a.c. equivalent circuit of the analysed inverting amplifier (Figure 3). The \([Y]\)-matrix of the circuit was composed using the well-known formulas (Boyanov & Shoikova, 1989), and after some transformations (using the condition \( r_0 \ll r_L \)) we obtain the following expression for the transfer function:

\[
A_U(s) = \frac{U_o(s)}{U_i(s)} = \frac{-R_F/R_N}{R_F C_1 \left(1 + \frac{r_o}{R_F} \right) \left(1 + \frac{r_o}{R_F} + \frac{r_l}{R_F} \right)}
\]

\[
= \frac{H}{s + j\omega_p}
\]  
(2)

Comparison of the left and right sides of Equation (2) results in the following formulas:

\[
H = A_{U0} = \frac{-R_F/R_N}{R_F C_1 \left(1 + \frac{r_o}{R_F} \right) \left(1 + \frac{r_o}{R_F} + \frac{r_l}{R_F} \right)}
\]  
(3a)

and

\[
\omega_p = \frac{1 + \frac{r_o}{R_F} \left(1 + \frac{r_l}{R_F} \right) \left(1 + \frac{r_o}{R_F} + \frac{r_l}{R_F} \right)}{R_F C_1 \left(1 + \frac{r_o}{R_F} \right) \left(1 + \frac{r_o}{R_F} + \frac{r_l}{R_F} \right)}
\]  
(3b)

where \( H \) is the d.c. voltage gain and \( \omega_p \) is the pole frequency, defining working frequency bandwidth \( f_{-3dB} \) of the circuit.

Therefore, for \( r_{in} \approx 0, \ r_0 \ll R_L \) and \( R_F \ll r_L \), the working frequency bandwidth \( f_{-3dB} \approx 1/(2\pi R_F C_1) \) of the amplifier depends only on the internal capacitance \( C_1 \) and the external feedback resistor \( R_F \). Moreover, the bandwidth \( f_{-3dB} \) does not depend on the resistance \( R_N \), setting the voltage gain of the circuit. The possibility for independent adjustment of gain and bandwidth is one of the main advantages of the amplifiers with CFOAs in comparison to those realized with VFOAs.

Based on the transfer function (2) for the module and the phase is obtained

\[
|A_U| = \frac{A_{U0}}{\sqrt{1 + f/f_0}^2} \text{ and } \varphi_{A_U} = 180^\circ - \arctan \left( \frac{f}{f_0} \right)
\]

3.2. A non-inverting amplifier

The electronic circuit of the non-inverting amplifier is shown in Figure 4. In this circuit the CFOA is with serial negative feedback through the resistors \( R_F \) and \( R_N \). The resistor \( R_F \) is used for compensation of the input bias current of the op amp, and the resistors \( R_T \) and \( R_o \) are used...
frequency bandwidth. In these cases, the resistance $r_{in}$ has values from several ohms to several tens of ohms. For example, the resistance $r_{in}$ is equal to $8 \Omega$ for the op amp AD8009 (from Analog Devices), while for the op amp AD844 (from Analog Devices) $r_{in}$ is $50 \Omega$. For relatively large gains and low frequencies (up to 50 MHz) according to formula (3b), the value of the feedback resistor $R_F$ is obtained by $R_F = r_{in}(f_t/f_{3dB}) - r_{in}A_{U0}$, where $f_t = 1/(2\pi r_{in}C_1)$ is the transit frequency of the CFOA.

The module and the phase for the non-inverting amplifier are given as

$$|A_U| = \frac{A_{U0}}{\sqrt{1 + \left(\frac{f}{f_p}\right)^2}}$$

and

$$\phi_{A_U} = \arctan\left(\frac{f}{f_p}\right).$$

From the analysis of the obtained formulas for the inverting and non-inverting amplifiers at low frequencies (i.e. $f \ll f_p$), the phase shift $\phi_{A_U}$ for the non-inverting amplifier is approximately equal to $0^\circ$ and for the inverting amplifier is $-180^\circ$, respectively. For higher frequencies (i.e. $f > f_p$) the gain is decreased to $|A_U| = A_{U0} \cdot (f_p/f)$ (the slope of $|A_U|$ is approximately equal to $-20 \text{ dB/dec}$). Furthermore, $|\phi_{A_U}| \approx 90^\circ$, which ensures the stable operation of the circuits (the phase margin is greater than $45^\circ$).

The results of these analyses can be used to study amplifier circuits with CFOAs, whose operating frequency bandwidth is up to 50 MHz. In these cases, the frequency response is maximally flat in the pass-band. Sample CFOAs suitable for applications up to 50 MHz are THS6184 (from Texas Instruments), LT1256 (from Linear Technology) and AD844 (from Analog Devices).

4. Analysis at high frequencies of the amplifier circuits

At higher frequencies ($> 50 \text{ MHz}$), analyses of the transfer function of the input network, and thus of the overall transfer characteristic of the systems of the amplifiers affect...
two capacitances \( C_{P} \) and \( C_{N} \). \( C_{P} \) is the capacitance to non-inverting input, which is formed by the capacitance \( C_{in}^{+} \)— Figure 2 plus the mounting capacitance (i.e. \( C_{P} = C_{in}^{+} + C_{M} \), where \( C_{M} \) is the parasitic board capacity with values usually up to 3 pF (Texas Instruments, 2002)). \( C_{N} \) is the capacitance to inverting input, which is formed by \( C_{in}^{-} \)and the mounting capacitance (i.e. \( C_{N} = C_{in}^{-} + C_{M} \)). Also the transfer characteristic of the amplifier circuits is affected by the load capacitance \( C_{L} \), connected in parallel to the resistance \( R_{L} \). In the following two subsections of the paper, the effects of \( C_{P} \), \( C_{N} \) and \( C_{L} \) on the frequency response of the inverting and non-inverting amplifiers are examined separately.

4.1. Effect of \( C_{P}, C_{N} \) and \( C_{L} \) on the frequency response of the non-inverting amplifier

The analysis of the circuit in Figure 4 is performed according to the method of the nodal voltages. The CFOA is replaced by the linear model, given in Figure 1. The transfer function (using the condition \( r_{in} \ll r_{1} \)) at \( Z_{L} = R_{L} || (1 / s C_{L}) \) (the load capacitance is a parallel connection of resistor \( R_{L} \) and parasitic capacitance \( C_{L} \)) and using the condition \( r_{o} \ll r_{1} \) can be found by

\[
A_{U}(s) \approx \frac{1}{s + \left( 1 \left/ \left( r_{1} || R_{P} \right) \right/ C_{P} \right)} \frac{s C_{R} r_{o} + C_{R} r_{o}}{r_{in}^{+} + R_{P} C_{R} r_{o}} \left( s + \frac{1 + \frac{r_{o}}{C_{R}}}{C_{R} C_{R} + C_{R} r_{o}} \right) \left( s + \frac{1 + \frac{r_{o}}{C_{R}}}{C_{R} r_{o}} \right)
\]

\[
= \frac{r_{in}^{+} + R_{P} C_{R} r_{o}}{s^{2} + \frac{1}{\frac{r_{o}}{C_{R} C_{R} + C_{R} r_{o}} + \frac{1 + \frac{r_{o}}{C_{R}}}{C_{R} C_{R} + C_{R} r_{o}}}} \left( s + \frac{1 + \frac{r_{o}}{C_{R}}}{C_{R} C_{R} + C_{R} r_{o}} \right)
\]

\[
= H \frac{s + \omega_{p, in}}{s^{2} + \omega_{p, in}^{2} + \frac{1}{C_{P} r_{P} o} + \frac{1}{C_{P} o^{2}}},
\]

where \( R_{P} = R_{0} + R_{P} \) is the equivalent resistance to the non-inverting input and \( R_{0} \) is the internal resistance of the input voltage source.

The equalization of the left and right sides of Equation (6) results in the following formulas for the basic parameters:

\[
H = \frac{r_{in}^{+} + R_{P} C_{R} r_{o}}{r_{in}^{+} + R_{P} C_{R} C_{N} \left( 1 + \frac{r_{o}}{C_{R}} + \frac{1}{C_{R} C_{R} + C_{R} r_{o}} \right)} \left( s + \frac{1 + \frac{r_{o}}{C_{R}}}{C_{R} C_{R} + C_{R} r_{o}} \right)
\]

\[
= \frac{r_{in}^{+} + R_{P} C_{R} r_{o}}{1 + \frac{r_{o}}{C_{R} C_{R} + C_{R} r_{o}} C_{R} C_{N} \left( 1 + \frac{r_{o}}{C_{R}} + \frac{1}{C_{R} C_{R} + C_{R} r_{o}} \right)} \left( s + \frac{1 + \frac{r_{o}}{C_{R}}}{C_{R} C_{R} + C_{R} r_{o}} \right)
\]

\[
= \frac{r_{in}^{+} + R_{P} C_{R} r_{o}}{r_{in}^{+} + R_{P} C_{R} r_{o}} \left( s + \frac{1 + \frac{r_{o}}{C_{R}}}{C_{R} C_{R} + C_{R} r_{o}} \right)
\]

is the transmission coefficient;

\[
\omega_{p, in} = \frac{1}{\left( r_{in}^{+} || R_{P} \right) C_{P}}
\]

is the pole angular frequency related to the effect of the resistance \( R_{P} \) and capacitance \( C_{P} \) to the non-inverting input;

\[
\omega_{z} = \frac{1 + R_{F} / R_{N}}{R_{F} C_{N} + r_{o} C_{1}}
\]

is the zero angular frequency related to the effect of the resistance \( R_{F} \) and capacitance \( C_{N} \) to inverting input of the CFOA;

\[
\omega_{p} = \frac{1}{\sqrt{R_{F} C_{N} \left( 1 + \frac{r_{o}}{C_{R}} + \frac{1}{C_{R} C_{R} + C_{R} r_{o}} \right)}}
\]

is the pole angular frequency (self-oscillating frequency or undamped natural frequency) and

\[
Q_{p} = \frac{R_{F} C_{N} \left( 1 + \frac{r_{o}}{C_{R}} + \frac{1}{C_{R} C_{R} + C_{R} r_{o}} \right)}{\sqrt{R_{F} C_{1} \left( 1 + \frac{r_{o}}{C_{R}} + \frac{1}{C_{R} C_{R} + C_{R} r_{o}} \right)}};
\]

is the quality factor of the circuit.

After substitution of \( s = j \omega \) in formula (6), based on the obtained general complex function, the module and the phase shift can be found by using

\[
|A_{U}(j \omega)| = A_{U}(\omega)
\]

\[
= \frac{\omega_{p, in}}{s + \omega_{p, in}^{2} + \frac{1}{C_{P} r_{P} o} + \frac{1}{C_{P} o^{2}}}
\]

\[
= \frac{\omega_{p, in}^{2}}{s^{2} + \omega_{p, in}^{2} + \frac{1}{C_{P} r_{P} o} + \frac{1}{C_{P} o^{2}}}
\]

\[
= \frac{\omega}{\omega_{p, in}} - \frac{\omega}{\omega_{p, in}} - \frac{\omega}{\omega_{p, in}}
\]

(12a)

(12b)

To compensate the effect of the parasitic capacitance \( C_{P} \), capacitor \( C_{P} \) can be placed in parallel with \( R_{P} \) so that \( C_{P} \gg C_{P} \) (Stoianov, 2000). The modified transfer function of the input network with capacitor \( C_{P} \) in parallel with \( R_{P} \) is

\[
T_{in}(s) = \frac{r_{in}^{+}}{r_{in}^{+} + R_{P} C_{R} r_{o}} \left( s + \frac{1 + \frac{r_{o}}{C_{R}}}{C_{R} C_{N} \left( 1 + \frac{r_{o}}{C_{R}} + \frac{1}{C_{R} C_{R} + C_{R} r_{o}} \right)} \right)
\]

(13)

If \( C_{P} \gg C_{P} \) and \( R_{P} \ll r_{in}^{+} \) then

\[
\frac{1 + \frac{r_{o}}{C_{R}}}{C_{R} C_{N} \left( 1 + \frac{r_{o}}{C_{R}} + \frac{1}{C_{R} C_{R} + C_{R} r_{o}} \right)} \approx 1
\]

and \( u_{i} \approx u_{j} \).

In this compensation method, the working frequency bandwidth is not narrowed and also the spikes are not reduced.

At the compensated effect of the capacitance \( C_{P} \), the analysis of formulas (6), (12a) and (12b) shows that the transfer function is characterized by a double pole (or a
complex pole) with angular frequency equal to \( \omega_p \) and one real zero with angular frequency \( -\omega_z \). However, for \( \omega = 0 \) the voltage gain has a value \( A_L(0) = H(\omega_z/\omega_p) = 1 + (R_F/R_N) \), while for much higher frequencies the gain decreases to zero. Therefore, the transfer characteristic of the non-inverting amplifier circuit is a low-pass type. For \( Q_p \geq 0.707 \) at a frequency equal to \( \omega_p \), the denominator tends to zero and the voltage gain theoretically increases towards infinity. As a result, peaking of the frequency characteristic occurs and the phase shift between the input and output signal increases rapidly to 180°. Moreover, the circuit of the non-inverting amplifier becomes unstable. In the cases where \( Q_p < 0.707 \) and \( \omega_p \ll \omega_z \), the frequency response monotonically decreases and the phase shift decreases to \( -180° \) (the phase margin is less than 45°). With increase in the frequency, the first component in the formula (12b) tends to be 90°, while the second component tends to be \( -180° \). From the theory of electronic circuits (Seifart, 2003), it is known that for \( \omega \geq 10\omega_p \) the phase shift increases to \( -180° \). According to formula (12b), the stable operation of the amplifier can be produced at the condition that the difference between \( \omega_p \) and \( \omega_z \) does not exceed 10 times. At \( \omega_z < \omega_p \) ringing in the output signal occurs, which can cause unstable operation.

Therefore, for the working bandwidth \( \omega_{-3dB} \), where \( |A_L(j\omega)| \) decreases with 3 dB (or \( 1/\sqrt{2} \)), the following is obtained:

\[
\omega_{-3dB} = \frac{\omega_p^2}{\omega_z} \sqrt{1 + \left( \frac{1}{2Q_p} \right) \omega_p^2 + \left( 1 + \frac{1}{2Q_p} \right) \omega_z^2 + \frac{1}{Q_p^2}} \quad (14)
\]

At \( Q_p = 1/\sqrt{2} \) formula (14) is simplified and yields

\[
\omega_{-3dB} = \frac{\omega_p^2}{\omega_z} \sqrt{1 + \sqrt{1 + \frac{\omega_z^4}{\omega_p^4}}} \quad (15)
\]

For \( \omega_z \gg \omega_p \), \( \omega_{-3dB} \approx \omega_p \). The value of the module of the general complex function (12a) at \( \omega = \omega_p \) is found by the formula

\[
A_L(\omega_p) = Q_p H \sqrt{\frac{\omega_z^2 + \omega_p^2}{\omega_p^2}}.
\]

Based on the analysis of the formulas for \( \omega_p, \omega_z \) and \( Q_p \) for the chosen resistance \( R_F \) and CFOA at \( Q_p < 0.707 \), in the frequency response ringing can occur under the condition

\[
R_N > \frac{R_F}{k_{\omega} - 1},
\]

where

\[
k_{\omega} = \sqrt{\frac{R_F C_N}{r_m C_1}} \sqrt{1 + \frac{r_o}{R_L} + \frac{C_L r_o}{C_N r_m}}
\]

is the coefficient of peaking of the output signal.

Figure 6. Magnitude and phase Bode plots for the loop gain of a voltage follower employing CFOA.

The magnitude and phase Bode plots for the loop gain of a voltage follower using CFOA are shown in Figure 6. A voltage follower or buffer with op amp is obtained from the non-inverting amplifier by removing the resistor \( R_N \). This circuit is analogous to the emitter (source) follower employing one transistor. The Bode plots are constructed by summing the corresponding logarithmic characteristics of all their sections, realizing zeros and poles:

\[
L(\omega) = 20 \log A_L(\omega) = \sum_{i=1}^{n} L_i(\omega) \quad \text{and} \quad \phi_A(\omega) = \sum_{i=1}^{n} \phi_i(\omega),
\]

where \( L_i(\omega) \) are the individual logarithmic magnitude characteristics and \( \phi_i(\omega) \) are the individual phase characteristics.

Formulas (12a) and (12b), and the corresponding Bode plots for \( \omega_z < \omega_p \), given in Figure 6, show the following:

1. In the voltage follower operation mode \( (A_L(0) = 1) \), there is always causes peaking in the frequency response, the coefficient \( k_{\omega} \) is greater than unit. By increasing the d.c. voltage gain \( A_L(0) \), the frequency \( \omega_z \) increases. As a result, the amount of peaks in the frequency response decreases;

2. The phase margin of the voltage follower and the non-inverting amplifier is greater than 45°. However, the stability of the circuit is worse. For a given frequency of the input signal, at which the asymptotes \( L_2 \) and \( L_3 \) intersect, the rate of change in voltage gain is about 40 dB/dec. The peaks in the magnitude plot lead to an increase in the amplitude of the output signal, which may adversely affect the next stages of the electronic devices and systems.
To compensate the effect of the capacitances $C_N$ and $C_L$, the following recommendations can be used:

1. Reduce the value of $C_N$ by removing ground or power plane around the circuit trace to the inverting input;
2. Reduce the value of the feedback resistor $R_F$.
3. As a result, the operating frequency bandwidth is expanded – formula (14);
4. Reduce the value of $C_L$ by minimizing the length of output cables;
5. For amplifiers the condition

$$A_{U0} > \frac{\sqrt{R_F C_N}}{1 + \frac{r_{in} C_L}{1 + \frac{r_{in} C_L}{R_F C_N}}}$$

has to be kept, then $\omega_p < \omega_z$. As well as it has to maintain the $Q_p$ to be less than $1/\sqrt{2}$, which can be achieved at $(R_F C_L / 2) > r_{in} C_N + r_{in} C_L$.

Furthermore, the obtained transfer function of the inverting amplifier shows that $\omega_p \gg \omega_z$, as it always satisfies the condition

$$\frac{r_o}{R_F} \sqrt{\frac{r_{in} C_L}{R_F C_N}} \frac{1}{1 + \frac{r_{in}}{R_L}} \ll 1.$$

For video line drivers (working up to 100 MHz) to isolate the output terminal from the load capacitance, a small resistor $R_o$, between the output of the op amp and the load, can be connected (Mancini, 2001; THS3091 – datasheet, 2014). Thus, the working bandwidth is narrowed by forming the additional pole frequency of $T_{out} = 1/[(R_o || R_L) C_L]$ and the d.c. transmission coefficient of the output network at $R_o = R_L$ (where $R_L$ is the characteristic impedance of the cable), yielding $T_{out, dc} = R_L / (R_o + R_L)$. Furthermore, the phase margin decreases at higher frequencies.

4.2. Effect of $C_p$, $C_N$ and $C_L$ on the frequency response of the inverting amplifier

The analysis of the inverting amplifier (Figure 2) is also performed according to the method of the nodal voltages, when the op amp is replaced by the linear model, presented in Figure 1. For the corresponding transfer function (using the condition $r_{in} \ll r_{iA}$) is found

$$A_U(s) = \frac{1}{s^2 + \frac{1}{C_N r_{in}} + \frac{1}{C_L r_{in}}} \left(1 + \frac{R_o}{\sqrt{R_F R_L}} \frac{1}{R_{in} C_L} \right)$$

for amplifiers employing CFOAs. The schematic design for the circuits in Figures 2 and 4 is based on the following sequence:

1. Technical specification. The circuit elements are calculated using predefined: amplitude of the input voltage $U_{in}$ or amplitude of the input voltage source $e_{iA}$ with internal resistance $R_G$; input resistance $R_{in}$; amplitude of the output voltage $U_{out}$; load resistance $R_L$ and capacitance $C_L$; parasitic board capacitance $C_M$; output resistance $R_{oA}$; cut-off frequency $f_{3db}$; relative error $e_{io}$ [%] defined by the input offset current and voltage, temperature drift $\epsilon_{Dio}$ [%] of the $e_{io}$ within temperature range $\Delta T$ and minimum value of a signal-to-noise (SN) ratio [dB].

2. An electronic circuit is selected. An object of an analysis and design are the amplifier circuits shown in Figures 2 and 4. The inverting circuit (Figure 2) provides smaller input resistance, while the non-inverting circuit (Figure 4) is with greater input resistance. Furthermore, the inverting circuit reduces the influence of the parasitic input capacitance at high frequencies (the inverting input $x$ is a virtual ground – $u_{ix} \approx 0$).
The op amp is selected. The main advantages of the CFOAs are the greater slew rate and the wider bandwidth compared to the VFOAs. The higher value of the slew rate is associated with a higher consumption current in a dynamic mode of operation. In order to produce low value of the power dissipation, most of the CFOAs work with supply voltages less than ±5 V. Then, at high frequencies, without additional output power stage, most CFOAs can get a maximum output current greater than 20 mA. The op amp is selected according to the following conditions:

- Maximum output voltage $U_{om} \geq U_{RL}$ ($U_{om}$ is the maximum output voltage of the op amp);
- The power supply voltage $V_{CC} = -V_{EE}$ is selected higher than the maximum output voltage $U_{om}$, as saving the condition $V_{CC_{min}} < V_{CC} < V_{CC_{max}}$;
- Maximum output current $I_{o,max} > I_L$, where $I_L = U_{RL}/R_L$;
- Small-signal bandwidth $f_1 > (5 \ldots 10)f_{-3dB}$, where $f_1$ is the cut-off frequency at voltage gain equal to 1;
- Slew rate $SR_{CFOA} > 2\pi f_{-3dB} U_{RL}$.

(4) The value of the equivalent quality factor of the frequency response is obtained:

$$Q_p = \sqrt{\omega_{-3dB}} \frac{r_{m} C_N \left(1 + \frac{\alpha_m}{r_L} + \frac{\alpha_m}{r_m} \right)}{1 + \frac{\alpha_m}{r_L}}.$$  

(5.1) For low-frequency (up to 50 MHz) amplifiers, the value of the feedback resistor $R_F$ is calculated:

- For the inverting amplifier $R_F = r_1 (f_1/f_{-3dB}) - (1 + |A_{U0}|)r_{in}$, where $f_1 = 1/(2\pi f_1 C_1)$ is the transit frequency of the CFOA and $|A_{U0}| = U_{RL}/U_{im}$ is the voltage gain of the circuit;
- For the non-inverting amplifier $R_F = r_1 (f_1/f_{-3dB}) - r_{in} A_{U0}$;

(5.2) At higher frequencies (> 50 MHz) and $\omega_p < \omega_2$, the value of the feedback resistor $R_F$ is calculated:

$$R_F = r_1 \frac{f_2 N}{f_{-3dB}} \frac{1}{1 + \frac{r_c}{r_m}} \left[1 - \frac{1}{2Q_p^2}\right] + \sqrt{\left(1 - \frac{1}{2Q_p^2}\right)^2 + 1},$$

where $f_2 N = 1/(2\pi r_{in} C_N)$.

(6) The value of the gain resistor $R_N$ is calculated:

- For the inverting amplifier $R_N = R_T/|A_{U0}|$
- For the non-inverting amplifier $R_N = R_T/(A_{U0} - 1)$.

The calculated values for the resistors $R_F$ and $R_N$ according to the aforementioned formulas have to be consistent with the values from the datasheet of the chosen CFOA.

(7) The value of the compensation resistor $R_F$ is determined as $R_F = R_N|/R_F$.

(8) The phase margin is calculated: $\varphi_{om} = 180^\circ - |\varphi_{om}(\omega_1)|$, where $\omega_1 = 2\pi f_1$.

For the inverting amplifier, the phase shift is

$$\varphi_{om}(\omega) = 180 - \arctan \frac{\omega \alpha_m}{Q_p (\omega_0^2 - \omega^2)}.$$  

The obtained value for the phase margin has to be greater than 45°. Otherwise, other op amps with smaller values of the parasitic capacitances should be chosen.

(9) The input impedance $Z_A$ is calculated:

- For the inverting amplifier: $Z_A \approx R_N$;
- For the non-inverting amplifier:

$$Z_A \approx \frac{r_m + R_p'}{\sqrt{1 + (f/f_{-3dB})^2}},$$

where $f_{-3dB} = 1/2\pi (r_m + R_p') C_p$ is the cut-off frequency of the input electrical network.

At low frequencies ($f \ll f_{-3dB}$) $Z_A \approx r_m + R_p'$.  

(10) The output resistance $R_{out}$ is calculated: $R_{out} \approx r_o/\beta A_{d0}$, where $\beta = R_N/(R_N + R_F)$ is the negative feedback coefficient and $A_{d0} \approx r_1/r_m$ is the d.c. open-loop voltage gain of the chosen CFOA.

(11) The output offset voltage of the circuit is calculated. First, the output offset voltage for room temperature $-25^\circ$ (using condition $R_F = R_T|/R_F$) is calculated: $U_{o,err} = (1 + R_F/R_{in})U_{io} - R_{F}|/R_{in}$, where $U_{io}$ is the input offset voltage and $I_{o}$ is the input offset current of the chosen CFOA. For video drivers the resistor $R_F$ is removed and a resistor $R_T$ in parallel to the non-inverting input of the amplifier is connected. In this case the output offset voltage is $U_{o,err} = (1 + R_F/R_{in})[U_{io} - (R_{C}|/R_{in})I_{B} + (R_F/R_N)I_{B}^o]$. Then the relative error $\varepsilon_{io} = (U_{o,err}/U_{RL})100\%$ is compared to the value given in step $N_{o}$.  

(12) The output offset voltage drift is calculated. First, the output offset voltage drift is calculated: $\Delta U_{o,err} = (1 + R_F/R_{in}) \Delta U_{io}(T) - R_{F}|/R_{in}$ for the given temperature range $\Delta T$. Then the relative error $\varepsilon_{io} = (\Delta U_{o,err}/U_{RL})100\%$ is compared to the value given in step $N_{o}$. 1. If the result does not satisfy the specification, a more precise op amp or performing new calculations for the resistances with lower values can be chosen.

13. The SN ratio is calculated. First, the resulting noise voltage density at the amplifier’s output is calculated: $\tilde{S}_{U_{out}} = \sqrt{\sum_i S_{I_{U_i}}}^2$ for $i = 1, 2, \ldots$, where $S_{I_{U_i}}$ is the individual noise components.
SN = \frac{U_{o,\text{eff}}}{U_{oN}} = \frac{U_{o,\text{eff}}}{S_{U,\text{out}} \sqrt{B_{eq}}},

where \( B_{eq} = 1.57f_{-3\text{dB}} \) is the bandwidth of the circuit multiplied by the correction factor of \( \pi/2 = 1.57 \) and \( U_{o,\text{eff}} \) is the output effective value. The obtained value for the SN is compared to the value given in step V9 of the procedure. If the result does not satisfy the specification, other op amps with lower voltage and current noise can be chosen.

6. Verification check, experimental testing and discussions

To verify the theoretical analysis and the proposed design procedure, in this section examples of studying the frequency stability of the inverting and non-inverting amplifiers at several voltage gains are given. A wide bandwidth (\( B_{1} > 300 \text{ MHz} \)) CFOA type AD8011 (AD8011 datasheet, 2014) and high-current (\( I_{o,\text{max}} > 200 \text{ mA} \)) CFOA type THS3091 (THS3091 datasheet, 2014) are chosen as active building elements for the investigated electronic circuits. The THS3091 uses an 8-pin SOIC and the 8-pin SOIC with PowerPAD™ (Texas Instruments Incorporated, Dallas, Texas 75243, USA) packages. The package type PowerPAD™ is designed so that the thermal pad is exposed on the bottom side of the integrated circuit (IC). The thermal coefficient for the PowerPAD packages is substantially improved over the basic SOIC.

The verification check of the op amps is performed by Cadence OrCAD® (Cadence Design Systems, Inc., San Jose, CA 95134, USA), using OrCAD PSpice® program with AD8011AN PSpice macro-model (version 1.0) (AD8011 SPICE macro-model, 2014) and THS3091 PSpice macro-model (THS3091 PSpice Model, 2014). The AD8011AN model simulates the input offset voltage and current (offsets will not vary with input common-mode voltage), small-signal closed-loop gain and phase versus frequency, output current limiting and output voltage limiting, slew rate, step response performance (slew rate is based on 10–90% of step response), quiescent current at operating point, noise effects, input impedance and output impedance. The values of the modelled parameters at \( V_S = \pm 5 \text{ V} \), \( R_L = 1 \text{ k\Omega} \) and \( T_A = 27^\circ \text{C} \) are as follows: \( U_{o,\text{io}} = 2 \text{ mV} \), \( I_{o,\text{io}}^+ = 5.125 \mu\text{A} \), \( I_{o,\text{io}}^- = 5.125 \mu\text{A} \), \( r_{o,\text{io}}^+ = 517 \text{k\Omega} \), \( C_{o,\text{io}}^+ = 1 \text{ pF} \), \( r_{o,\text{io}}^- = 50 \text{ \Omega} \), \( C_{o,\text{io}}^- = 2.3 \text{ pF} \), \( r_t = 1.27 \text{ M\Omega} \), \( C_t = 1.52 \text{ pF} \) (or \( f_t = 82.5 \text{ kHz} \)), \( f_s \approx 670 \text{ MHz} \) (at \( A_{t,\text{o}} = 1 \)), \( S_{U,\text{io}} = 7.53 \text{nV}/\sqrt{\text{Hz}} \) (at \( f = 10 \text{ kHz} \)), \( U_{o,\text{im}} = \pm 4 \text{ V} \), \( SR > 1000 \text{ V}/\mu\text{s} \), \( I_{o,\text{max}} = 60 \text{ mA} \) and \( r_o = 22 \Omega \).

The THS3091 model simulates input offset voltage, input bias currents, small-signal closed-loop gain and phase versus frequency (bandwidth is high in gains of \(+1 \text{ V/V} \) and \(+2 \text{ V/V} \) and low at higher gains), output voltage limiting, slew rate, step response performance (slew rate is correct at 2 V step), settling time, quiescent current, noise effects, output impedance and loading effects. The values of the modelled parameters at \( V_S = \pm 15 \text{ V} \), \( R_L = 100 \text{ \Omega} \) and \( T_A = 27^\circ \text{C} \) are as follows: \( U_{o,\text{io}} = 0.9 \text{ mV} \), \( I_{o,\text{io}}^+ = -4.5 \mu\text{A} \), \( I_{o,\text{io}}^- = -3.5 \mu\text{A} \) (or \( I_{o,\text{io}} = 4 \mu\text{A} \) and \( I_{o,\text{io}} = 1 \mu\text{A} \)), \( r_{o,\text{io}}^+ = 1.1 \text{ M\Omega} \), \( C_{o,\text{io}}^+ = 1.2 \text{ pF} \), \( r_{o,\text{io}}^- = 32 \Omega \), \( C_{o,\text{io}}^- = 1.4 \text{ pF} \), \( r_t = 848 \text{k\Omega} \), \( C_t = 0.8 \text{ pF} \) (or \( f_t = 234 \text{ kHz} \)), \( f_s \approx 240 \text{ MHz} \) (at \( A_{t,\text{o}} = 1 \)), \( S_{U,\text{io}} < 6 \text{nV}/\sqrt{\text{Hz}} \) (at \( f \approx 10 \text{ kHz} \)), \( U_{o,\text{im}} \approx \pm 3.1 \text{ V} \), \( SR > 1000 \text{ V}/\mu\text{s} \) (at \( R_f = 1.21 \text{k\Omega} \) and \( A_{t,\text{o}} = 2 \)) and \( r_o = 100 \Omega \).

Table 1. Comparison between calculated parameters and simulation results for \( A_{t,\text{o}} = +5, R_L = 50 \text{ \Omega} \) and \( C_L = 20 \mu\text{F} \).

| Parameter | Calculated results | Simulation results |
|-----------|--------------------|--------------------|
| \( A_{t,\text{o}} = +5, f_{-3\text{dB}} = 100 \text{ MHz} \), \( R_f = 1.96 \text{k\Omega} \) and \( R_N = 490 \Omega \) (\( R_f = 1.96 \text{k\Omega} \pm 1\% \) and \( R_N = 487 \Omega \pm 1\%) \), \( U_{o,\text{im}} = 100 \text{ mV} \), \( R_{G} = 50 \text{ \Omega} \) and \( R_T = 50 \text{ \Omega} \) | \( U_{o,\text{err}} = 19.40 \text{ mV} \) | \( U_{o,\text{err}} = 19.32 \text{ mV} \) |
| \( A_{t,\text{o}} \) | 5 | 5 |
| \( Q_p \) | 0.429 | 0.5 |
| \( f_{-3\text{dB}} \) | – | 100.2 \text{ MHz} |
| \( \phi_{im} \) | 130.5° | 126.4° |
| \( A_{t,\text{o}} = +5, f_{-3\text{dB}} = 120 \text{ MHz} \), \( R_f = 1.32 \text{k\Omega} \) and \( R_N = 330 \Omega \) (\( R_f = 1.33 \text{k\Omega} \pm 1\% \) and \( R_N = 332 \Omega \pm 1\%) \), \( U_{o,\text{im}} = 100 \text{ mV} \), \( R_{G} = 50 \text{ \Omega} \) and \( R_T = 50 \text{ \Omega} \) | \( U_{o,\text{err}} = 16.12 \text{ mV} \) | \( U_{o,\text{err}} = 16.05 \text{ mV} \) |
| \( A_{t,\text{o}} \) | 5 | 5 |
| \( Q_p \) | 0.54 | 0.52 |
| \( f_{-3\text{dB}} \) | – | 121.2 \text{ MHz} |
| \( \phi_{im} \) | 120° | 116° |
| \( A_{t,\text{o}} = +5, f_{-3\text{dB}} = 150 \text{ MHz} \), \( R_f = 845 \Omega \) and \( R_N = 211 \Omega \) (\( R_f = 845 \Omega \pm 1\% \) and \( R_N = 210 \Omega \pm 1\%) \), \( U_{o,\text{im}} = 100 \text{ mV} \), \( R_{G} = 50 \text{ \Omega} \) and \( R_T = 50 \text{ \Omega} \) | \( U_{o,\text{err}} = 13.69 \text{ mV} \) | \( U_{o,\text{err}} = 13.64 \text{ mV} \) |
| \( A_{t,\text{o}} \) | 5 | 5 |
| \( Q_p \) | 0.602 | 0.650 |
| \( f_{-3\text{dB}} \) | – | 140.6 \text{ MHz} |
| \( \phi_{im} \) | 110° | 106° |
The verification check of the models for AD8011AN and THS3091 is performed by comparing the simulation results to the datasheet typical parameters of the real op amps.

In Table 1 the calculated parameters and the simulation results for three values of the working bandwidth – 100, 120 and 150 MHz – are presented. The voltage gain is chosen with value equal to $+5$ and $R_{L}||C_{L} = 50 \Omega||20 \text{pF}$.

Figure 7. Module of the complex transfer function at $R_{F} = 1 \text{k}\Omega$ and d.c. voltage gain $+1$, $+2$ and $+6$, respectively.

Figure 8. Phase shift of the complex transfer function at $R_{F} = 1 \text{k}\Omega$ and d.c. voltage gain $+1$, $+2$ and $+6$, respectively.
Based on the proposed design procedure, values for the passive components and values for the basic dynamic parameters were found. The maximum error between the calculated values of the electrical parameters and the simulation results is not higher than 10%. Moreover, an error of 10% is quite acceptable considering the tolerances of the technological parameters.

After implementation of a verification check by computer simulations, a combination of simulation and experimental study on various circuits of inverting and non-inverting amplifiers was performed. The tested circuits were implemented on a FR4 PCB laminate with surface-mount device passive components for the resistors and capacitors.

The study of the electronic circuits is performed in two stages. The first stage of computer simulations and experimental study is implemented for the inverting and non-inverting amplifier circuits using CFOA AD8011ARZ, biased with ±5 V supplies. The values of the chosen passive components for the investigated non-inverting amplifier circuit (Figure 4) are as follows: (1) \( R_F = 1 \, k\Omega \) and \( R_N \to \infty \) at \( A_{U0} = 1 \); (2) \( R_F = 1 \, k\Omega \) and \( R_N = 1 \, k\Omega \) at \( A_{U0} = 2 \); (3) \( R_F = 1 \, k\Omega \) and \( R_N = 200 \, \Omega \) at \( A_{U0} = 6 \) with tolerances ±1%. The resistor \( R_T \) is chosen equal to 51 \( \Omega \) with tolerance ±1%. The coefficient \( k_\omega \) of ringing of the output signal, according to the formula given in Section 4.1, is 5.53. The a.c. transfer characteristics of the circuits are obtained experimentally by using network analyser HP4195A. To measure the output signal, an active probe type HP41800A with input impedance 100k\( \Omega \)/3pF is used. For the a.c. sweep analysis, the frequency is swept from 100 kHz to 1 GHz by decades, with 100 points per decade. The input voltage source is chosen with amplitude −10 dBm or 70.8 mV (with initial phase shift equal to zero), to ensure amplitude of the output voltage, not higher than 500 mV at \( A_{U0} = 6 \). The simulation and experimental results for the module and phase shift at three values of the d.c. voltage gain for the non-inverting amplifier are plotted in Figures 7 and 8. As can be seen for low frequencies approximately up to 10 MHz, the gains (Figure 7) are with constant value and are frequency independent. At \( A_{U0} \) equal to 1 and 2 in the form of the frequency response causes peaking and for the voltage follower the amplitude reaches almost 15 dB. Moreover at gains 1 and 2, the module of the transfer function decreases with greater speed, such as for the frequency equal to 500 MHz reaches value equal to −10 dB. The difference between the simulation and the experimental results is due to the influence of the additional parasitic poles determined by the inertial intermediate stages of the CFOA. The simulated values of the zero-pole pair at \( A_{U0} = 1 \) are \( f_z = 30 \, \text{MHz} \) and \( f_p = 250 \, \text{MHz} \), respectively.

For \( A_{U0} = 2 \), \( f_z = 60 \, \text{MHz} \) and \( f_p = 250 \, \text{MHz} \). At the voltage gain equal to 6, \( f_z = 180 \, \text{MHz} \) and \( f_p = 250 \, \text{MHz} \), the amount of peaking is small, because the frequencies are close and as a result, the transfer characteristic monotonically decreases to unity. The phase margin (Figure 8) for the three gains is greater than 45°, which means that the

![Figure 9](image-url)  
Figure 9. Alternating current transfer characteristics at \( R_F = 500 \, \Omega \) and d.c. voltage gain +1, +2 and +6, respectively.
Figure 10. Alternating current transfer characteristics at d.c. voltage gain $-1$, $-2$, $-5$ and $-10$, respectively.

Figure 11. Alternating current transfer characteristics at $R_F = 1000\ \Omega$ and $C_L$ with values 10, 20 and 47 pF.
amplifier circuits are stable according to the Bode criterion. At gains equal to 1 and 2, the phase shift between the input and the output signal is positive, because \( \omega_z < \omega_p \). This additional phase shift has a value less than 50°, which does not affect the stability of the circuits.

To verify the results of the theoretical analysis at \( R_T = 500 \Omega \), new non-inverting amplifiers were implemented with the following passive components: (1) \( R_T = 1 \mathrm{k}\Omega \) and \( R_N = 1 \mathrm{k}\Omega \) at \( A_{\mathrm{UD}} = 1 \); (2) \( R_T = 1 \mathrm{k}\Omega \) and \( R_N = 499 \Omega \) at \( A_{\mathrm{UD}} = 2 \); (3) \( R_T = 1 \mathrm{k}\Omega \) and \( R_N = 200 \Omega \) at \( A_{\mathrm{UD}} = 5 \) and (4) \( R_T = 499 \Omega \) and \( R_N = 51 \Omega \) at \( A_{\mathrm{UD}} = 10 \). All resistors were chosen with tolerance \( \pm 1\% \). The resistor \( R_T = 51\Omega \pm 1\% \) is used for gains \(-1\), \(-2\) and \(-5\). The a.c. transfer characteristic at the four values of the voltage gains is presented in Figure 10. As can be seen, all frequency characteristics are not ringing \( (Q_p < 0.707) \) and decreased monotonically with increase in the frequency of the input signal. Furthermore, the zero frequency \( f_z \) is much greater than the frequency of the double pole. The phase shift between the input and the output signal varies from +180° to −180°, as for the frequency approximately equal to 400 MHz the phase shift is 0°. For gains \(-1\) and \(-2\) the phase margin is greater than 45°. At gains equal to \(-5\) and \(-10\) for frequencies greater than 240 MHz, the phase margin is less than 45° and the circuits are unstable. The phase shift of the output signal varies approximately up to −90°. This phase shift is determined by the influence of additional parasitic poles in the frequency response of the op amp.

The second series of computer simulations and experimental study was performed for the non-inverting amplifier circuits (with \( R_T = 1 \mathrm{k}\Omega \), \( R_N = 250 \Omega \) and \( A_{\mathrm{UD}} = 5 \)) at complex (active-capacitive) load employing CFOA THS3091D (using an 8-pin SOIC with PowerPAD™), biased with \( \pm 15 \mathrm{~V} \) supplies. The a.c. transfer characteristic at amplitude of the output voltage equal to 4 V and for three values (10, 20 and 47 pF) of the capacitive load of the amplifier is plotted in Figure 11. The \( R_C \) was chosen equal to 100 \( \Omega \), as the maximum output current is 40 mA. For frequencies up to 30 MHz, the module is with constant value, approximately equal to 5.0 (or \( \approx 14 \mathrm{~dB} \)). At capacitive load equal to 10 and 20 pF, the quality factor, according to formula (11), is equal to 0.5 and 0.71, respectively. In these cases, the peaks are not observed and the transfer characteristic decreases monotonically with an increase in the frequency of the input signal. At capacitive load equal to 50 pF, the quality factor becomes larger than 0.707 and in the form of transfer characteristic ringing occurs. Furthermore, the phase shift between the input and the output signal increases and at frequency equal to 330 MHz it becomes 135°. At further increase of the frequency, the phase margin becomes smaller than 45°, which decreases the stability of the amplifier.

### 7. Conclusion

In this paper, a study of the frequency stability analysis for high-speed inverting and non-inverting amplifiers using CFOAs has been presented. Based on the analysis of the principle of operation, equations for the complex transfer functions are obtained, as well as recommendations for improving the stability at complex load and design procedure are defined. The proposed procedure can be useful for the analysis and design of high-speed amplifier circuits employed in various analogue and mixed-signal circuits, such as video amplifiers, line drivers and analogue switches. The results obtained by the theoretical analyses are validated through simulation and experimental testing of sample electronic circuits using monolithic op amps AD8011 and THS3091. The maximum error between calculated values and the simulation results for the basic electrical parameters is not higher than 10%.

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### Disclosure statement

No potential conflict of interest was reported by the authors.

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