

Chapter 2

CFOAs: Merits, Demerits, Basic Circuits and Available Varieties

2.1 Introduction

Current feedback op-amps (CFOA) started attracting attention of the analog circuit designers and researchers when it was realized that one can design amplifiers exhibiting a characteristic which was the most significant departure from the characteristics exhibited by well-known VOA-based realizations in that CFOA-based circuits could realize variable-gain and yet constant bandwidth, as against the unavoidable gain-band-width-conflict in case of the VOA-based designs (as explained in Chap. 1). Furthermore, it was recognized that due to much higher slew rates of the order of several hundred to several thousand V/ μ s (which can be as large as 9,000 V/ μ s for modern CFOAs), as compared to a very modest 0.5 V/ μ s for the general purpose and most popular μ A741-type VOA, CFOAs could lead to circuits capable of operating over much wider frequency ranges than those possible with VOAs.

In this chapter, we focus on the merits and demerits of CFOAs; discuss the various basic analog circuits realizable with CFOAs and highlight a variety of commercially available IC CFOAs from the various leading IC manufacturers.

2.2 AD844: The CFOA with *Externally-Accessible Compensation Pin*

Although in view of the popularity of the CFOAs they have been manufactured as integrated circuits by a number of IC manufacturers, there are two varieties which are in use. There are CFOAs which are pin-compatible to VOAs and do not have externally accessible compensation pin. On the other hand, AD 844-type CFOA from Analog Devices [1] has the option that its compensation pin (number 5) is externally-accessible while still maintaining pin-capability with VOAs.

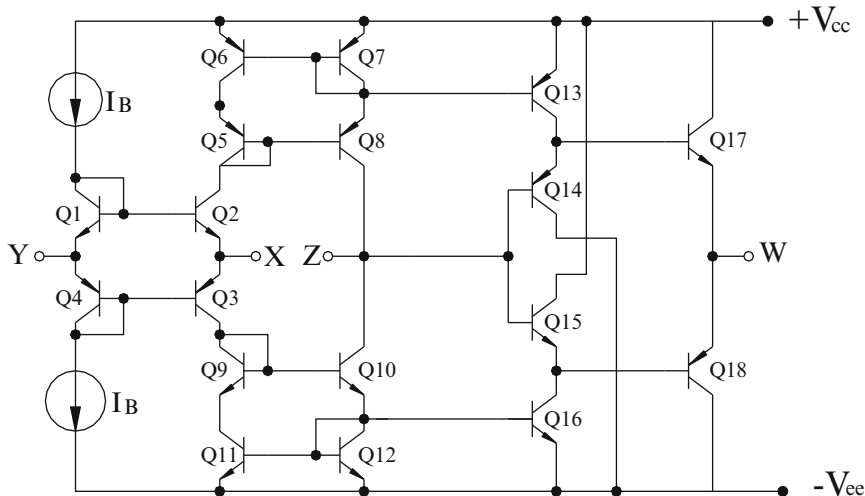


Fig. 2.1 A simplified schematic of the CFOA AD844 (adapted from [1] © 1990 Analog Devices, Inc.)

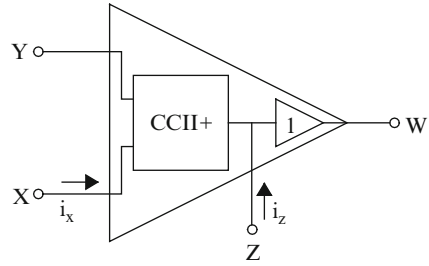
The AD844 from Analog Devices is a high speed monolithic (current feedback) op-amp which has been fabricated using junction- isolated complementary bipolar (CB) process. It has high bandwidth (around 60 MHz at gain of -1 and around 33 MHz at gain of -10) and provides very fast large signal response with excellent DC performance. It has very high slew rate, typically, $2,000 \text{ V}/\mu\text{s}$. Although it is optimized for use in current to voltage conversion applications and as inverting amplifier, it is also suitable for use in many non-inverting and other applications. Typical applications recommended by the manufacturers include Flash ADC input amplifiers, High speed current DAC interfaces, Video buffers and cable drivers and pulse amplifiers.

The AD844 can be used for replacement of traditional VOAs but due to its current feedback architecture results in much better AC performance, high linearity and excellent pulse response. The off-set voltage and input bias currents of the AD844 are laser- trimmed to minimize DC errors such that drift in the offset voltage is typically $1 \mu\text{V}/^\circ\text{C}$ and bias current drift is around $9 \text{ nA}/^\circ\text{C}$. AD844 is particularly suitable for video applications and as an input amplifier for flash type analog-to-digital converters (ADC). A simplified schematic of the AD844 CFOA [1] is shown in Fig. 2.1.

It is interesting to point out that due to AD844 being sold, *disguised* as a large bandwidth, high slew-rate op-amp, initially it almost got unnoticed that its internal architecture, is, in fact, a translinear second generation plus type Current Conveyor¹

¹ The *Current Conveyors* were introduced as new circuit building blocks by Sedra and Smith in [2, 3]; the first generation Current Conveyor (CCI) in [2] and the more versatile, the second generation Current Conveyor (CCII \pm) in [3].

Fig. 2.2 A block diagram of the internal architecture of CFOA AD844



(CCII+) followed by a (translinear) voltage buffer. Its simplified symbolic diagram showing this identification is shown in Fig. 2.2.

Since the internal architecture of AD844 consists of a CCII + followed by a voltage buffer, this flexibility was later found to be useful in allowing the AD844 to be used as a CCII + and CCII– (using two CCII+), as pin-by-pin replacement of a VOA (with Z-pin left open) and lastly, as a 4-terminal building block in its own right.

In view of its front end being a CCII + and the back end being a voltage follower, the terminal equations of the CFOA can be written as

$$i_y = 0, \quad v_x = v_y, \quad i_z = i_x \quad \text{and} \quad v_w = v_z \quad (2.1)$$

In the internal architecture of the CFOA, transistors Q_1 – Q_4 are configured as a mixed translinear cell (MTC) while the collector currents of transistors Q_2 and Q_3 are sensed by two modified p-n-p and n-p-n Wilson Current Mirrors consisting of transistors Q_5 – Q_8 and Q_9 – Q_{12} respectively to create a replica of current i_x at the terminal- Z thereby yielding $i_z = i_x$. The two constant current sources, each equal to I_B , force equal emitter currents in transistors Q_1 and Q_4 thereby forcing input current $i_y = 0$ when a voltage V_y is applied at the input terminal Y. It can be easily proved that with $i_x = 0$, $V_x = V_y$ and the Z-port current i_z will be zero. However, for the case of $i_x \neq 0$, an exact analysis [4] of the circuit using exponential relations between collector currents and base-emitter voltages for the transistors Q_1 – Q_4 yields

$$I_z = I_x = -2I_B \sinh\left(\frac{V_y - V_x}{V_T}\right) \quad (2.2)$$

from which an approximate relation between V_x , V_y and r_x (for $I_x \ll 2I_B$) can be expressed as follows

$$V_x \cong V_y + r_x i_x \quad \text{where} \quad r_x = \frac{V_T}{2I_B} \quad (2.3)$$

If terminal-Z is terminated into an external impedance/load Z_L , a voltage V_z is created which passes through the voltage follower made from another MTC

composed of transistors Q_{13} – Q_{18} for which transistors Q_{13} and Q_{16} provide the DC bias currents. The last stage is characterized by an equation similar to (2.3) which provides $V_w \cong V_z$.

2.3 The Merits and the Advantageous Features of the CFOAs

Two major merits and advantageous features of the CFOAs are (1) its very high (theoretically infinite) slew rate and (2) its capability of realizing amplifiers exhibiting gain-bandwidth decoupling. In the following, we elaborate these two characteristics of the CFOAs.

2.3.1 The Reason and the Origin of the High Slew Rate

In this sub-section we explain the origin and the reason for a very high slew rate of CFOAs as compared to conventional op-amps [5]. Figure 2.3a shows a simplified schematic of an internally compensated type IC op-amp exhibiting the differential transconductance stage consisting of transistors Q_1 – Q_2 – Q_3 – Q_4 , the intermediate gain stage (normally made from a cascade of CC-CE stages) having an inverting gain $-A_{v2}$ and the output stage which is a class AB type push-pull amplifier having both complementary transistors in emitter follower mode providing a voltage gain A_{v3} close to unity.

A straight forward analysis of the first stage reveals that the output current I_{out} is given by

$$I_{out} = I_B \tanh \left(\frac{V_{id}}{2V_T} \right) \quad (2.4)$$

A graphical representation of the above equation is shown in Fig. 2.3b. From this characteristic, it is seen that the output current i_o saturates to $+I_B$ when V_{id} is large and positive while i_o saturates to $-I_B$ when V_{id} is large and negative. Thus, the maximum current available to charge the compensating capacitor C_c , is $\pm I_B$.

If such an op-amp is configured as a voltage follower by a feedback connection from V_{out} to the inverting input terminal of the op-amp and a large step signal is applied to the non-inverting input terminal at $t = 0$. This forces the transistor Q_1 into saturation and Q_2 into cut off due to which $I_{out} = I_B$ and thus, the capacitor C_c is charged *linearly* through constant current I_B .

In view of the high gain of the intermediate stage, for simplicity, its input node can be treated to be at *virtual ground* potential in which case one can write

$$I_{out} = C_c \frac{dV_{out}}{dt} \quad (2.5)$$

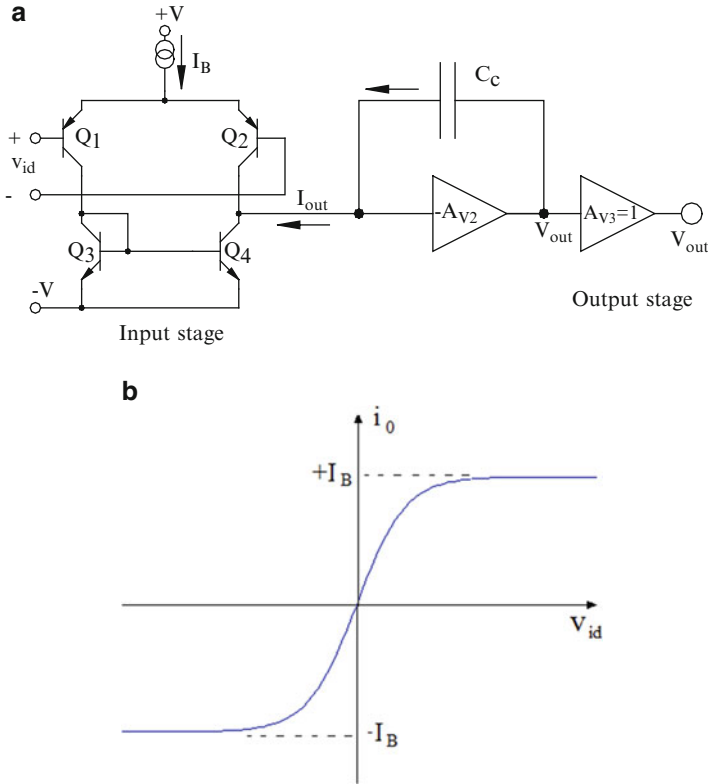


Fig. 2.3 (a) Simplified model of an internally compensated IC op-amp. (b) The tanh-characteristics of the input differential transconductance stage

Hence, the slew-rate (SR) is given by

$$SR = \left. \frac{dV_{out}}{dt} \right|_{\max} = \frac{I_{out}}{C_c} = \frac{I_B}{C_c} \quad (2.6)$$

With $C_c = 30$ pF and $I_B = 19$ μ A (as applicable to a μ A741 type op-amp biased with ± 15 V DC power supplies), the above figure turns out to be around 0.63 V/ μ s which is close to the data sheet value of 0.5 V/ μ s. For a sinusoidal output $V_0 = V_m \sin \omega t$, it can be shown that the maximum frequency ω_{\max} , for which the limitation imposed by the finite slew rate will not come into play, is given by

$$\omega_{\max} = \frac{SR}{V_m} \quad (2.7)$$

Consider now a simplified schematic of the CFOA shown in Fig. 2.4a. An analysis of the input stage of the CFOA, which is made from MTC consisting of

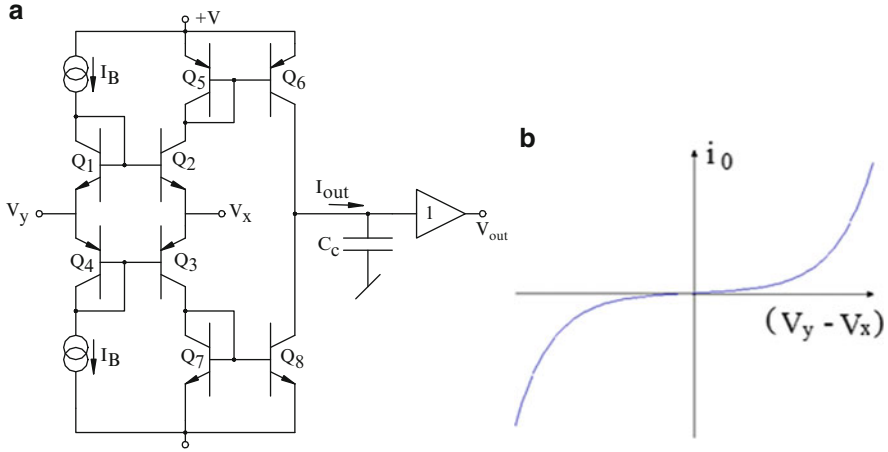


Fig. 2.4 (a) Simplified model of the CFOA. (b) The transfer characteristic between i_o and $(V_y - V_x)$

transistors $Q_1 - Q_4$ shows that the current output coming out of Z- terminal (which charges the compensating capacitor) is given by

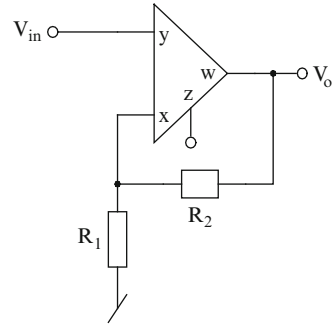
$$I_{out} = 2I_B \sinh\left(\frac{V_y - V_x}{V_T}\right) \quad (2.8)$$

A plot of the resulting transfer characteristic is shown in Fig. 2.4b. Thus, in this case, it is found that for a large differential input voltage $(V_y - V_x)$, the output current which is the charging current of the compensating capacitor would be theoretically infinite. Thus, in contrast to VOAs, CFOAs have ideally infinite slew rate. In practice, slew rates from several hundred $V/\mu s$ to as high as $9,000 V/\mu s$ are attainable. Consequently, a CFOA implementation of a circuit will not have the same kind of limitations on the maximum operational frequency range as prevalent in the corresponding VOA-based circuit. In other words, a CFOA-based circuit would operate satisfactorily over a frequency range much larger than possible for a VOA circuit realizing the same function.

2.3.2 De-coupling of Gain and Bandwidth: Realisability of Variable-Gain, Constant-Bandwidth Amplifiers

It has been explained in the previous chapter that all VOA-based controlled sources suffer from the drawback of gain- bandwidth-conflict. An important advantage of employing CFOAs is that this gain bandwidth conflict can be overcome due to the current feedback prevalent in the same configurations realized with CFOAs

Fig. 2.5 The non-inverting amplifier using a CFOA



(Interestingly, we will see that even two alternative ways of realizing VCVS from CFOAs are also free from the gain-bandwidth-conflict).

Consider now the CFOA-based non-inverting amplifier of Fig. 2.5.

From an analysis of this circuit, taking CFOA characterization as $i_y = 0$, $v_x = v_y$, $i_z = i_x$ and $v_w = v_z = -i_z Z_p$ where Z_p is the parasitic impedance looking into the Z-terminal and consists of a resistance R_p (typically, around 3 M Ω) in parallel with a capacitance C_p (typically in the range 4.5 – 5.5 pF), the maximum gain of the circuit is found to be

$$\frac{V_0}{V_{in}} = \frac{\left(1 + \frac{R_2}{R_1}\right)}{\left(1 + \frac{R_2}{R_p}\right)} \quad (2.9)$$

whereas the –3 dB bandwidth is given by

$$\omega_{3-dB} = \frac{1}{C_p R_2} \left(1 + \frac{R_2}{R_p}\right) \cong \frac{1}{C_p R_2}; \quad \text{for } R_2 \ll R_p \quad (2.10)$$

It is, thus, seen that the bandwidth of the circuit can be fixed by setting the feedback resistor R_2 while the gain can be still varied through the variable resistor R_1 and therefore, the gain and bandwidth have become de-coupled and it has become possible to realize a constant-bandwidth, variable gain amplifier.

2.4 The Demerits and Limitations of CFOAs

2.4.1 Demerits

Despite its significant advantages over traditional VOAs, as explained in previous section, CFOAs generally have the following demerits:

- Relatively inferior DC precision.
- Relatively poor DC offset voltage due to the use of both PNP and NPN transistors.

- Lower CMRR and PSRR than VOAs due to the unsymmetrical complimentary-pair input stage and unequal and un-correlated input bias currents.

A detailed analysis of the input DC current, input offset voltage and maximum input voltage range for the input stage of a CFOA is given in [6] while a comprehensive analysis of output stage has been dealt in [7].

2.4.2 Difficulties with Capacitive Feedback

It should be kept in mind in devising CFOA-based circuits that a capacitive feedback between X and W is not recommended as it often leads to instability. Therefore, an inverting Miller integrator cannot be realized with a CFOA in the same way as the conventional op-amp-based Miller integrator.

2.4.3 Effect of Stray Capacitances and Layout Issues

Another important practical consideration to be taken care of is to take care of the stray capacitances on the inverting input node (X-input) and across the feedback resistor which invariably lead to peaking or ringing in the output response and sometimes even to oscillations. In view of this, appropriate care has to be taken in making an appropriate PCB layout and eliminate any stray capacitances. The performance of a CFOA-based circuit can be improved considerably with a good layout, good decoupling capacitors and low inductance wiring of the components.

2.5 Basic Circuits Using CFOAs

We now show how a number of basic analog circuits such as the four controlled sources, the voltage and current followers, the instrumentation amplifier and the integrators and differentiators can be realized in a number of advantageous ways using CFOAs sans the disadvantages associated with VOA-based realizations of the same functions.

2.5.1 VCVS Configurations

Consider now the various other VCVS realizations depicted in Fig. [2.6a–c](#).

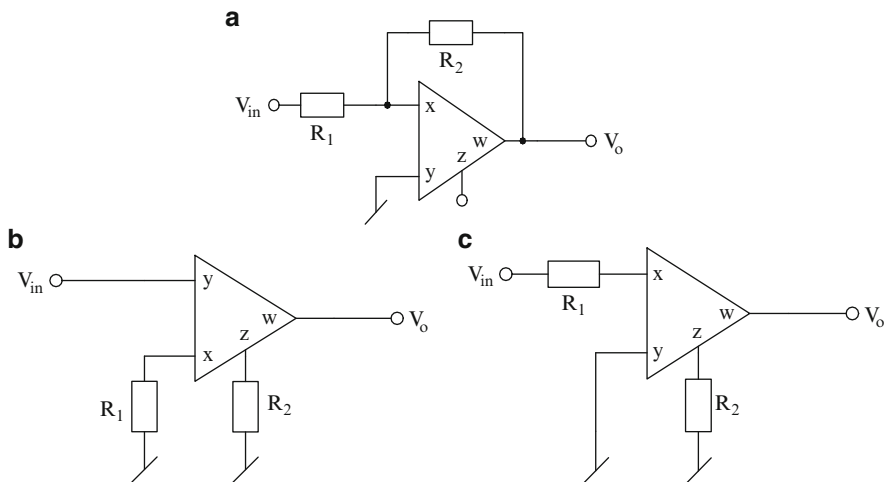


Fig. 2.6 Realization of various other VCVS circuits using a CFOA (a) inverting VCVS, (b) alternative non-inverting VCVS, (c) alternative inverting VCVS

A non-ideal analysis of all the three circuits reveals their non-ideal gains as:

$$\frac{V_0}{V_{in}} = -\frac{\frac{R_2}{R_1}}{\left(1 + \frac{R_2}{R_p}\right)} \text{ for the circuit of Fig. 2.6a} \quad (2.11)$$

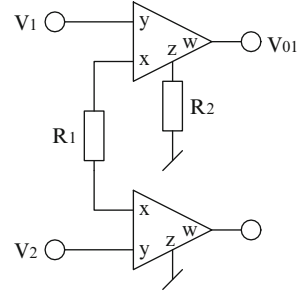
$$\frac{V_0}{V_{in}} = \frac{\frac{R_2}{R_1}}{\left(1 + \frac{R_2}{R_p}\right)} \text{ for the circuit of Fig. 2.6b} \quad (2.12)$$

$$\frac{V_0}{V_{in}} = \frac{-\left(\frac{R_2}{R_1}\right)}{\left(1 + \frac{R_2}{R_p}\right)} \text{ for the circuit of Fig. 2.6c} \quad (2.13)$$

whereas the 3-dB bandwidth in all cases is given by the same value as in (2.10). Thus, in all the cases, the bandwidth can be set by the feedback resistor R_2 after which the gain can still be made variable through a single variable resistance R_1 . Thus, the gain bandwidth conflict is not present in any of the four circuits. It is, therefore, possible to design constant-bandwidth variable-gain amplifiers using CFOAs which unfortunately cannot be done with the same topologies such as those of Figs. 2.5 and 2.6a realized with a traditional VOA.

However, it must be kept in mind that, in practice, constant bandwidth is achievable only for low to medium gains (typically, 1–10). Furthermore, the feedback resistor R_2 also cannot be chosen arbitrarily since this critically affects

Fig. 2.7 An instrumentation amplifier using CFOAs (CFOA-version of Wilson's CCII-based circuit [8])



the stability of the amplifier. In fact, the CFOA parameters r_x (typically, around $50\ \Omega$) and Z-pin parasitics $R_p \parallel \frac{1}{sC_p}$ (where $R_p = 3\text{ M}\Omega$; $C_p = 4.5\text{ pF}$) with the feedback resistance R_2 decide the stability of the non-inverting and inverting amplifiers using CFOAs (if realized with CFOAs configured exactly similar to their VOA-counterparts). The manufacturer determines the optimum value of the feedback resistor R_2 during the characterization of the IC. Normally, lowering R_2 decreases stability whereas increasing R_2 decreases the bandwidth.

2.5.2 Instrumentation Amplifier Using CFOAs

We now show that, contrary to the traditional instrumentation amplifier which requires three VOAs and as many as seven resistors out of which four are required to be completely matched, the use of CFOAs makes it possible to realize a variable gain instrumentation amplifier with no more than two CFOAs along with a minimum number of only two resistors. Such a circuit is readily evolved from a known CCII-based circuit proposed by Wilson [8] and is shown here in Fig. 2.7.

Considering the finite input resistance looking into terminal-X of the CFOA as r_x and taking parasitic output impedance looking into terminal-Z as a resistance R_p in parallel with capacitance C_p , the maximum gain of this circuit is found to be:

$$\frac{V_0}{V_1 - V_2} = \frac{R_2}{(R_1 + 2r_x) \left(1 + \frac{R_2}{R_p}\right)} \quad (2.14)$$

whereas its 3-dB bandwidth is given by the some expression as in (2.10). Thus, it is seen that the bandwidth of the amplifier can be fixed at a constant value by fixing R_2 while the gain can be made variable by changing R_1 . Thus, CFOA-based instrumentation amplifier also does not have the gain-bandwidth-conflict while employing a minimum possible number of passive components for realizing a variable gain.

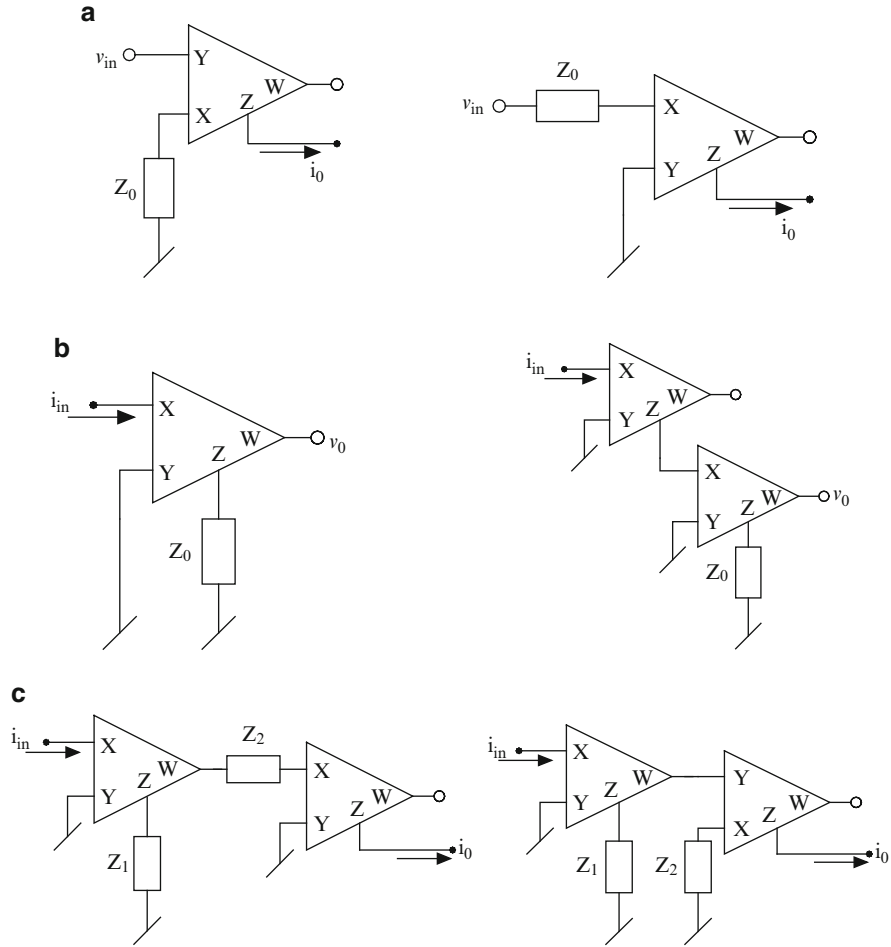


Fig. 2.8 Various controlled sources (a) Voltage controlled current sources. (b) Current controlled voltage sources. (c) Current controlled current sources

2.5.3 VCCS, CCVS and CCCS Configurations

In Fig. 2.8 we show the CFOA-based realization for non-inverting and inverting VCCS, CCVS and CCCS circuits. It may be noted that contrary to VOA-based circuits for VCCS and CCCS requiring as many as four identical resistors the corresponding realizations using CFOAs as in Fig. 2.8a–c employ a minimum possible number of passive components namely only one in case of Fig. 2.8a, b and two in case of Fig. 2.8c respectively thus, no component matching whatsoever is needed. Furthermore, it is straight forward to verify that all these circuits possess the most notable property of CFOA-based circuits i.e. no gain-bandwidth-conflict in the realization of any controlled sources.

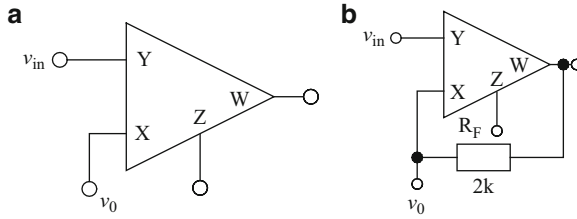


Fig. 2.9 Unity gain voltage followers using CFOA

2.5.4 Unity Gain Voltage and Current Followers

Figure 2.9 shows two different ways of realizing a unity gain voltage follower using CFOAs. In the first case since between terminals Y and X there is already a voltage follower inside the chip, the same voltage buffer can be used as a voltage follower. In the second case, a slightly modified version from [9] is presented which contains a feedback resistor R_F for the self-compensation of the voltage follower.

A non-ideal analysis of the voltage follower of Fig. 2.9b considering the X-port input resistance r_x and Z-port parasitic impedance consisting of a resistance R_p in parallel with a capacitance C_p , reveals the following non-ideal gain function for this circuit

$$\frac{V_0}{V_{in}} = \frac{\left(1 + \frac{R_F}{R_p}\right)}{\left(1 + \frac{r_x + R_F}{R_p}\right)} \left\{ \frac{1 + sC_p(R_p // R_F)}{1 + sC_p(R_p // (r_x + R_F))} \right\} \quad (2.15)$$

If $R_F \gg r_x$, it is seen that a pole-zero cancellation would take place and the resulting voltage gain will be close to unity and will be perfectly compensated for. It is found that for a voltage follower made from AD844-type CFOA, the circuit works quite well with $R_F = 2 \text{ k}\Omega$ [9].

The two possible realizations for unity gain current follower are shown in Fig. 2.10. As expected, none of the two circuits requires any resistors and both the circuits offer ideally zero input resistance and ideally infinite output resistance.

2.5.5 Integrators and Differentiators

In this subsection we first explain some integrators and differentiators [10] realizable similar to their VOAs counter parts. Due to the reason spelt out earlier an inverting integrator with a CFOA is not feasible. Since a capacitive feedback from

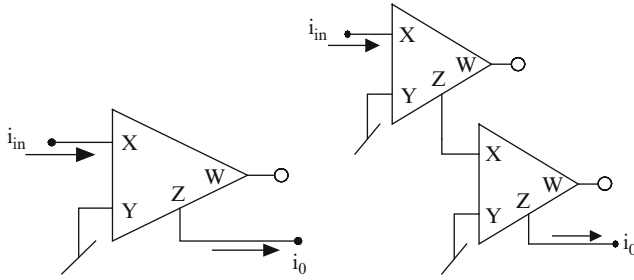


Fig. 2.10 Unity gain current followers using CFOA

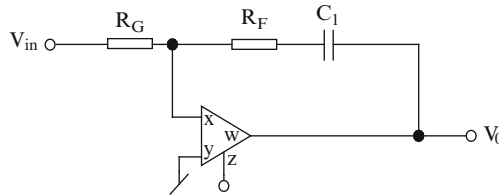


Fig. 2.11 An inverting integrator using a CFOA [10]

W to X leads to instability. However, a slightly modified version with an additional resistance incorporated in the feedback path is still possible as shown in Fig. 2.11.

Addition of resistor R_F is acceptable since at high frequency the resistor is dominant and hence feedback impedance would never drop below the resistor value. The transfer function of this circuit is given by

$$\frac{V_0}{V_{in}} = -\left(\frac{R_F}{R_G}\right) \left[\frac{s + \frac{1}{R_F C_1}}{s} \right] \quad (2.16)$$

$$\approx \frac{-1}{s C_1 R_G}; \quad \text{for } \omega \ll \frac{1}{R_F C_1} \quad (2.17)$$

On the other hand, to realize a non-inverting integrator, one can make Deboo's integrator [11] almost in the same manner as is done with a VOA (see Fig. 2.12) however; this circuit suffers from the drawback of requiring four identical resistors and also has to fulfill a condition to ensure stable operation.

This circuit is characterized by the following transfer function.

$$\frac{V_0}{V_1} \cong \left(\frac{1 + \frac{R_F}{R_G}}{s R_1 C_1} \right) \quad (2.18)$$

Fig. 2.12 CFOA-version of non-inverting Deboo's integrator [10, 11]

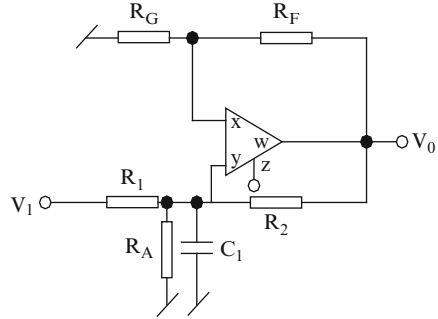
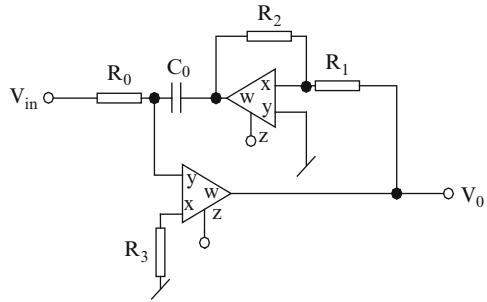


Fig. 2.13 Active-compensated non-inverting CFOA integrator



whereas the condition required for stable operation is

$$\frac{R_2}{R_1 // R_A} \geq \frac{R_F}{R_G} \quad (2.19)$$

To circumvent the above problems, in Fig. 2.13 we show an alternative circuit for creating non-inverting integrator using two CFOAs [12]. This circuit has an in-built compensation for the non-ideal effects of the CFOA parasitic impedances.

The circuit of Fig. 2.13 realizes a non-inverting integrator since its transfer function is given by

$$\frac{V_0}{V_{in}} = \frac{1}{sT} \quad \text{where } T = \frac{C_0 R_0 R_2}{R_1} \quad (2.20)$$

Considering the Z-port parasitic impedance $Z_p = R_p // \frac{1}{sC_p}$ for both the CFOAs, a non-ideal analysis reveals

$$\frac{V_0}{V_{in}} \cong \frac{R_1}{sC_0 R_0 R_2} \varepsilon(s) \quad (2.21)$$

Fig. 2.14 A differential integrator using a CFOA

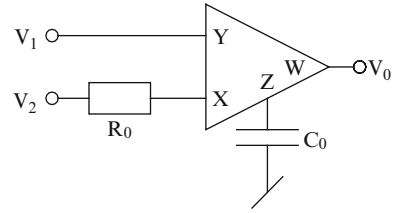
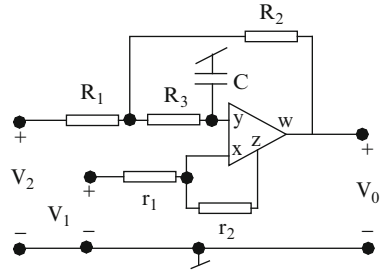


Fig. 2.15 Dual input integrator proposed by Lee and Liu (adapted from [13])
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for $R_i < < R_{pi}$, $i = 0-3$. The error function $\varepsilon(s)$ is given by

$$\varepsilon(s) = \frac{1 + sT_2}{1 + sT_1 + s^2T_1T_2} \quad \text{with } T_1 = C_P R_2; T_2 = C_P R_1 R_3 / R_2 \quad (2.22)$$

From the above, the phase error is given by

$$\phi \cong \omega(T_2 - T_1) - \omega^3 T_1^2 T_2 \quad (2.23)$$

Hence, for negligible phase error, one requires $T_1 = T_2$ which gives the required condition as $R_3 = R_2^2 / R_1$.

From the above, it is seen that with $R_3 = R_2^2 / R_1$, the phase error is minimized and active-compensation is achieved.

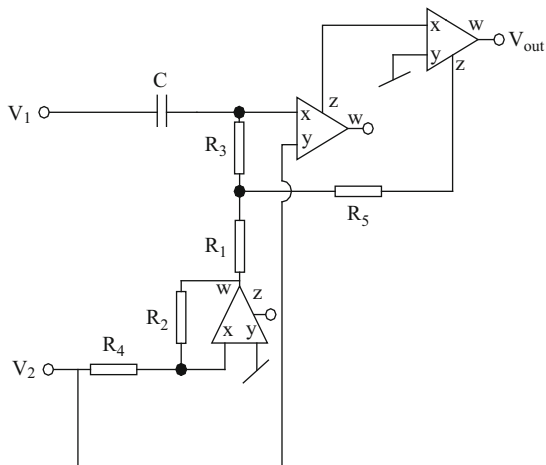
In the above cases, the circuits devised using CFOAs are exactly similar to their VOA counterparts. However, since a CFOA has an in built CCII+, there is an alternative way of realizing inverting/non-inverting integrators. A general circuit to realize an integrator in an alternative manner is shown in Fig. 2.14. An analysis of this circuit shows that the output voltage is given by

$$V_0 = \frac{1}{sC_0 R_0} (V_1 - V_2) \quad (2.24)$$

Thus, both inverting and non-inverting integrators can be realized from this circuit as special cases by grounding V_1 or V_2 respectively. A differentiator is obtainable from the same circuit by interchanging the resistor and the capacitor.

We now show a circuit which can perform the operation of dual input integrator using a single CFOA proposed by Lee and Liu [13] (Fig. 2.15).

Fig. 2.16 Dual input differentiator using CFOAs proposed by Lee and Liu (adapted from [13])
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Analysis of this circuit reveals

$$V_{out} = \frac{V_2 \frac{R_2}{R_1} - V_1 \frac{r_2}{2r_1}}{sC \left(R_3 \left(1 + \frac{R_2}{R_1} \right) + R_2 \right) - \left(\frac{r_2}{2r_1} - \frac{R_2}{R_1} \right)} \quad (2.25)$$

If we choose $R_2/R_1 = r_2/2r_1$ the circuit realizes a dual input integrator with output voltage given by

$$V_{out} = \frac{1}{s\tau} (V_2 - V_1) \quad (2.26)$$

$$\tau = CR_1 \left[1 + R_3 \left(\frac{1}{R_1} + \frac{1}{R_2} \right) \right] \quad (2.27)$$

From equations (2.26) and (2.27) it is seen that the time constant of the integrator can be varied by changing the resistor R_3 . The circuit operates well within the frequency range of 450 Hz to 1 MHz with a phase error of 5°.

A dual-input differentiator [13] is shown in Fig. 2.16. The input of this circuit with $R_4 = (R_2 + R_3)$ and $R_5 = R_2$, is given by

$$V_{out} = V_2 \left(\alpha - \frac{R_2}{R_3} (1 - \alpha) \right) + sC(V_1 - V_2) \left[R_2 + R_1 \left(1 + \frac{R_2}{R_3} \right) \right] \quad (2.28)$$

If $\alpha = R_2/(R_2 + R_3)$, (2.28) reduces to

$$V_{out} = sCR_2(V_1 - V_2) \left(1 + R_1 \left(\frac{1}{R_2} + \frac{1}{R_3} \right) \right) \quad (2.29)$$

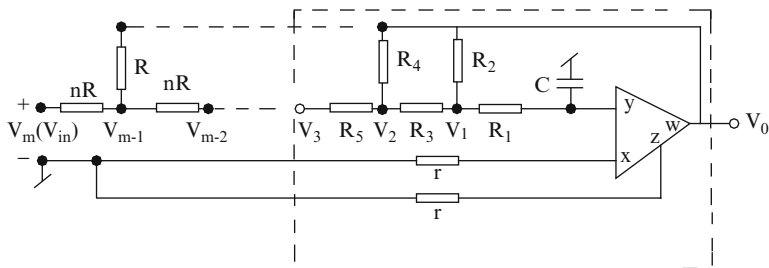


Fig. 2.17 Integrator with time constant multiplication proposed by Lee and Liu (adapted from [14] © 2001 IET)

Hence, the time constant can be varied by changing R_1 . Over an operating frequency range of 1–100 kHz, this circuit works well with a phase error of the order of $\pm 10^\circ$.

In Fig. 2.17 we show another integrator circuit which was proposed by Lee and Liu in [14] and has the facility for time constant multiplication. Analysis of this circuit, as in [14], shows that its transfer function is given by

$$\frac{V_0}{V_{in}} = \frac{1}{snRC \frac{1}{2^m \sqrt{n^2 + 4n}} \left[(n + 2 + \sqrt{n^2 + 4n})^m - (n + 2 - \sqrt{n^2 + 4n})^m \right] + 1} \quad (2.30)$$

By appropriate selection of m and n , a desired multiplication factor can be achieved. For instance, if we take $m = 3$ and $n = 10$ with V_3 as the input, it is possible to achieve a multiplication factor of 143.

The transfer function of the differentiator circuit of Fig. 2.18 is given by

$$V_0 = sC \frac{R_5 \left(1 + \frac{R_2}{R_1}\right) + R_2 \left(1 + \frac{R_5}{R_4}\right) \left(\frac{R_3}{R_1} + \frac{R_3}{R_2} + 1\right)}{1 - \frac{R_2 R_3}{R_1 R_4}} V_{in} \quad (2.31)$$

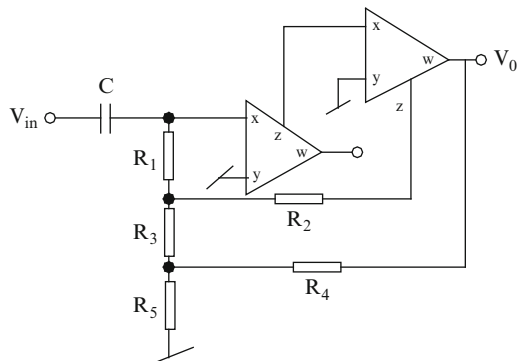
If $R_2 = R$, $R_4 = 2R$, $R_1 = R_3 = R_5 = 2nR$ the above equation can be expressed as

$$V_0 = 4sRC \left[n \left(1 + \frac{1}{2n}\right) + (n+1)(n+1) \right] \quad (2.32)$$

As an example, if we select $R_2 = 1 \text{ k}\Omega$, $R_4 = 2 \text{ k}\Omega$, $R_1 = R_3 = R_5 = 10 \text{ k}\Omega$ then the multiplication factor turns out to be 166.

In reference [14], it has been demonstrated that the circuit of Fig. 2.17 works well with in the frequency range 200Hz to 1 MHz with a phase error of less than 12° whereas for the differentiator of Fig. 2.18, the operating frequency range has been found to be 100 Hz to 10 kHz with a phase error less than 6° .

Fig. 2.18 Differentiator with time constant multiplication proposed by Lee and Liu (adapted from [14] © 2001 IET)



It must be mentioned that yet another differential integrator implemented from two AD 844-type CFOAs and capable of operating up to several MHz without encountering any stability problem was presented by Maundy et al. in [15].

2.6 Commercially Available Varieties of CFOAs

Although a wide varieties of CFOAs are available from various IC manufacturers, optimized with respect to a chosen parameter, it is interesting to note that the key building blocks used are two types of *mixed translinear cells*. In the following, we identify these two basic blocks and then briefly describe the internal architecture and characteristics/parameters of some exemplary IC CFOAs available from leading analog IC manufacturers.

2.6.1 The Mixed-Translinear-Cells (MTC) as Building Blocks of CFOAs

Most of the CFOA architectures have the internal structure of a CCII + followed by a voltage buffer. Since a CCII + itself has a voltage follower between its Y and X terminals, it, therefore, follows that a typical CFOA architecture would have two voltage followers (VF): one between Y and X terminals and the other between Z and W terminals. Furthermore, there has to be a mechanism of sensing the current flowing into the low-input impedance terminal X of the input VF, creating a copy of the same and making it available at the high output impedance Z-terminal where a compensating capacitor can be connected either internally or externally. Two standard configurations for realizing VFs are the two *mixed translinear Cells*

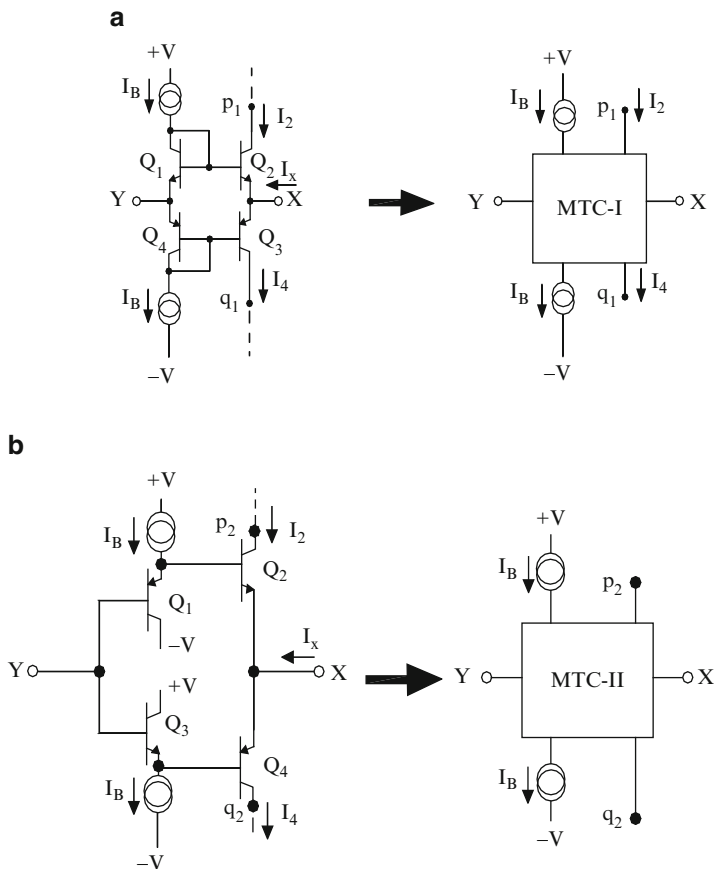


Fig. 2.19 The two types of mixed-translinear cells (MTC) (a) MTC-I, (b) MTC-II (adapted from [4] ©1997 Taylor & Francis)

(MTC) [4, 16] shown in Fig. 2.19a, b. An analysis of the type-I MTC reveals that the current I_x and differential input $(V_y - V_x)$ are related by the following equation:

$$I_x = 2I_B \sinh\left\{\frac{V_y - V_x}{V_T}\right\} \quad (2.33)$$

Incidentally, type-II MTC of Fig. 2.19b, although has a different topology, it is also governed by the same equation [4]. This equation can be re-arranged as:

$$\frac{V_y - V_x}{V_T} = \sinh^{-1}\left(\frac{I_x}{2I_B}\right) \cong \frac{I_x}{2I_B}; \text{ for } I_x \ll 2I_B \quad (2.34)$$

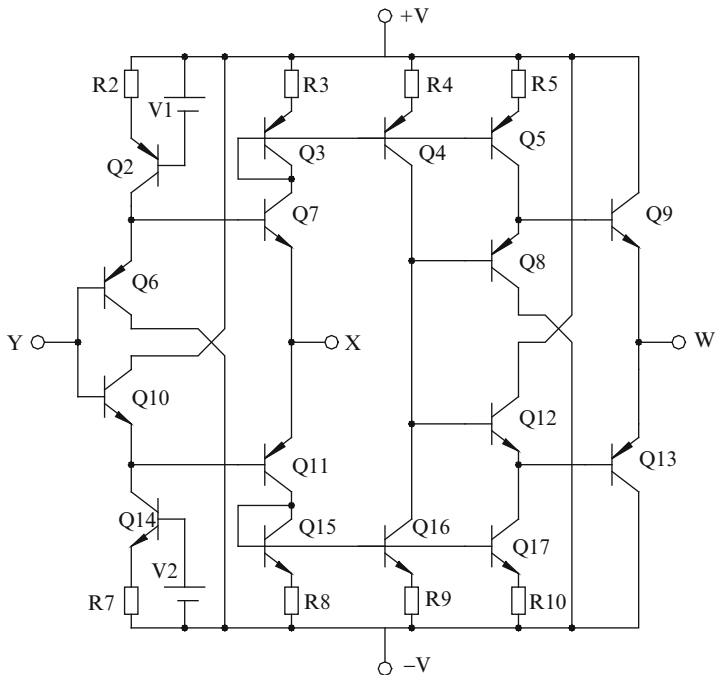


Fig. 2.20 Elantec dual/quad CFOA EL2260/EL2460 (adapted from [17] © 1995 Intersil American Inc.)

$$V_x \cong V_y - I_x r_x \quad \text{where} \quad r_x = \frac{V_T}{2I_B} \quad (2.35)$$

Note that when r_x is zero, one gets $V_x = V_y$ (as it should be, in the ideal case).

2.6.2 Elantec Dual/Quad EL2260/EL2460

Figure 2.20 shows a simplified schematic of Elantec dual/quad 130 MHz CFOA EL 2260/EL 2460 [17]. As can be seen, this architecture has both input and output buffers as type- II MTC and no compensating lead is available externally. This CFOA provides 130 MHz 3-dB band width (for a gain of +2) with a slew rate of 1,500 V/ μ s.

2.6.3 Intersil HFA 1130

Intersil HFA1130 (Fig. 2.21) CFOA is an ideal choice for applications requiring output limiting which allows the designer to set the maximum positive and negative output levels thereby protecting the later stages from damage or input saturation [18].

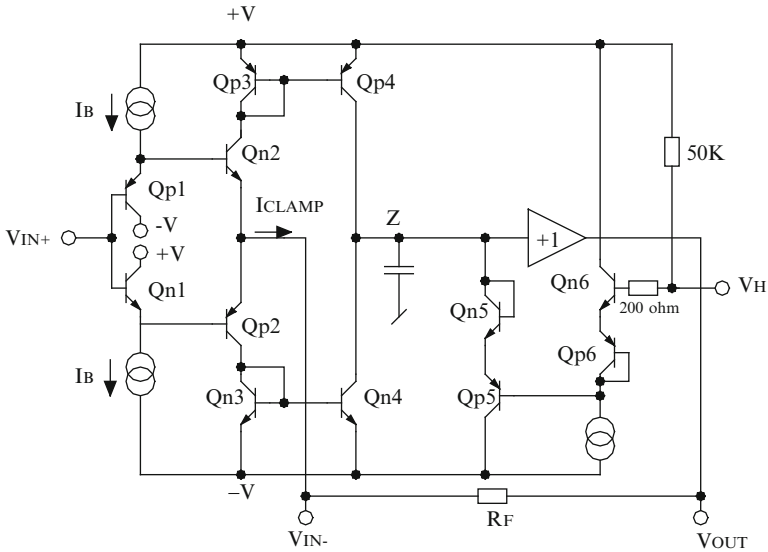


Fig. 2.21 Intersil HFA1130 output-limiting low-distortion CFOA (adapted from [18] © 2005 Intersil American Inc.)

The mechanism of high clamp (V_H circuit) can be explained as follows. The unity gain buffer made from type-II MTC forces V_{IN-} to track V_{IN+} and sets up a slewing current $= (V_{IN-} - V_{OUT})/R_F$. This current through the mirror action of the current mirrors Q_{p3} – Q_{p4} and Q_{n3} – Q_{n4} creates a replica of this current at the high impedance node Z. The base voltage of Q_{p5} is $2V_{BE}$ (Q_{n6} and Q_{p6}) less than V_H which permits the conduction of Q_{p5} whenever the voltage at the Z node reaches a voltage $= Q_{p5}$'s base $+2V_{BE}$ (Q_{p5} and Q_{n5}) in this manner the transistor Q_{p5} clamps node Z whenever Z reaches to a voltage level $= V_H$. The resistance R_1 acts as a pull-up resistance to ensure functionality with the clamp input floating. There is similar circuit (not shown in this diagram) which provides a symmetrical low clamp control by voltage V_L .

HFA1130 has a slew rate of the order of $2,300 \text{ V}/\mu\text{s}$ and -3 dB bandwidth of 850 MHz and is capable of provide a high output current of the order of 60 mA and is recommended for applications in the design of residue amplifier, video switching and routing, pulse and video amplifiers, Flash A/D Driver, RF/IF signal processing and Medical imaging systems.

2.6.4 AD8011 from Analog Devices

Figure 2.22 shows a simplified schematic of the two-stage CFOA AD8011 from Analog Devices [19]. The input stage is a type-I MTC with a complementary second gain stage created from the pair of transistors Q_5 and Q_6 . The circuit

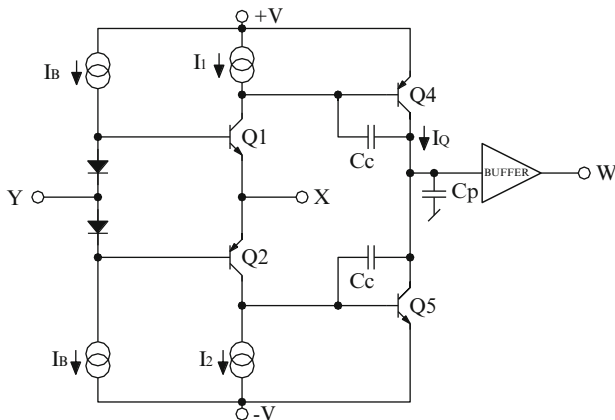


Fig. 2.22 Simplified schematic of the Analog Devices two-stage CFOA AD8011 (adapted from [19] © 1995 Analog Devices Inc.)

provides low distortion; high speed and high current drive while running on low quiescent currents. This CFOA has a -3 dB bandwidth of 57 MHz, slew rate of 3,500 V/ μ s, output current of 30 mA with quiescent power of 12 mW.

2.6.5 THS 3001 from Texas Instruments Inc.

Figure 2.23 shows the CFOA THS3001 from TI has 420 MHz 3-dB bandwidth for gain of +1, and has slew rate of 6,500 V/ μ s with current output drive as high as 100 mA. The simplified schematic of this CFOA is shown in Fig. 2.23.

This CFOA is built by using a 30-V dielectrically isolated, complementary bipolar process with NPN and PNP transistors possessing f_T of several GHz. This configuration implements an exceptionally high performance CFOA having wide bandwidth, high slew rate, fast settling time (40 ns) and low distortion (THD -80 dBc at 10 MHz).

Lastly, it may be pointed out that a wide varieties of CFOAs optimized for enhancement of one or more of the several specific performance features such as higher slew rate, increased output current drive capability, wider bandwidth etc. are available from leading IC manufacturers. For further information, the readers are referred to the datasheets of various IC manufacturers. Lastly, it may be pointed out that CFOAs with slew rate as high as 9,000 V/ μ s (such as THS3202 from Texas Instruments Inc.) are available as off-the-shelf items.

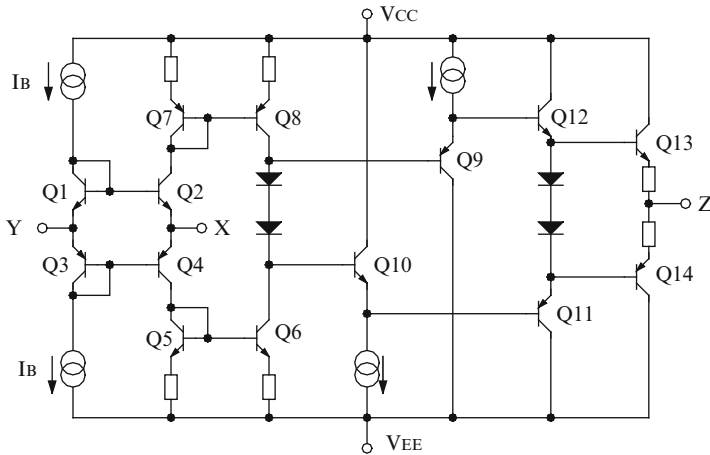


Fig. 2.23 A simplified equivalent of 420-MHz, high-speed CFOA THS 3001 type CFOA (adapted from [10] © 2009 Texas Instruments Inc.)

2.7 Concluding Remarks

In this chapter, we have outlined the distinct merits of CFOAs over VOAs particularly the mechanism leading to a high (theoretically infinite) slew rate and the resolution of the gain-bandwidth conflict resulting in the notable property of the CFOA-based circuits (particularly VCVS structures) of providing constant-bandwidth with variable gains. We have also outlined the various de-merits of the CFOAs [5] namely, their inferior CMRR, unsymmetrical input bias dc currents, high input offset voltage and lower PSRRs etc.

Various basic analog circuit building blocks using CFOAs were outlined and a number of examples of commercially available CFOAs from leading IC manufacturers were highlighted.

In spite of their limitations, CFOAs are quite useful for numerous applications which can be carried out more efficiently with CFOAs than with VOAs, with one or more of the following advantages: employment of smaller number of external passive components, elimination of passive component-matching requirements in several cases and higher operational frequency range. In fact, the nature of many high frequency applications of CFOAs is such that the very high slew rate puts the CFOA in the spotlight [20].

In view of the above, it must be emphasized that the focus of the subsequent chapters of the present book would be primarily on those applications where the CFOAs are found to provide significant advantages and/or resulting in novel circuits—the type of which cannot be realized with conventional VOAs.

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