CHAPTER 1

INTRODUCTION

1.1 INTRODUCTION OF CFA

In today's electronic systems, various forms of signals are necessary, such as sinusoidal, square, triangular, pulse, waves. Computer and control systems need clock pulses. In communication systems, there are a variety of waveforms needed for modulation waveforms. Triangular waves are used for scanning an electron beam on a CRT screen, in precise time measurements, and in time modulation. The sine wave is the most fundamental waveform. In a mathematical sense, any other waveform can be expressed as the Fourier combination of basic sine waves. Sine waves are used extensively in test, reference, and carrier signals. Despite its simplicity, the generation of a pure sine wave is challenging. Sinusoidal, triangular and rectangular waveforms are generated using current feedback operational amplifiers.

Current-feedback amplifiers (CFA) do not have the traditional differential amplifier input structure, thus they sacrifice the parameter matching inherent to that structure. The CFA circuit configuration prevents them from obtaining the precision of voltage-feedback amplifiers (VFA), but the circuit configuration that sacrifices precision results in increased bandwidth and slew rate. The higher bandwidth is relatively independent of closed-loop gain, so the constant gain-bandwidth restriction applied to VFAs is removed for CFAs. The slew rate of CFAs is much improved from their counterpart VFAs because their structure enables the output stage to supply slewing current until the output reaches its final value. In general, VFAs are used for precision and general purpose applications, while CFAs are restricted to high frequency applications above 100 MHz.

Although CFAs do not have the precision of their VFA counterparts, they are precise enough to be dc-coupled in video applications where dynamic range requirements are not severe. CFAs, unlike previous generation high-frequency amplifiers, have eliminated the ac coupling requirement; they are usually dc-coupled while they operate in the GHz range. CFAs have much faster slew rates than VFAs, so they have faster rise/fall times and less inter-modulation distortion.

1.2 CFA MODEL

The CFA model is shown in Figure 1.1. The non-inverting input of a CFA connects to the input of the input buffer, so it has very high impedance similar to that of a bipolar transistor non-inverting VFA input. The inverting input connects to the input buffer's output, so the inverting input impedance is equivalent to a buffer's output impedance, which is very low. Z_B models the input buffer's output impedance, and it is usually less than 50Ω. The input buffer gain, G_B, is as close to one as IC design methods can achieve, and it is small enough to neglect in the calculations.

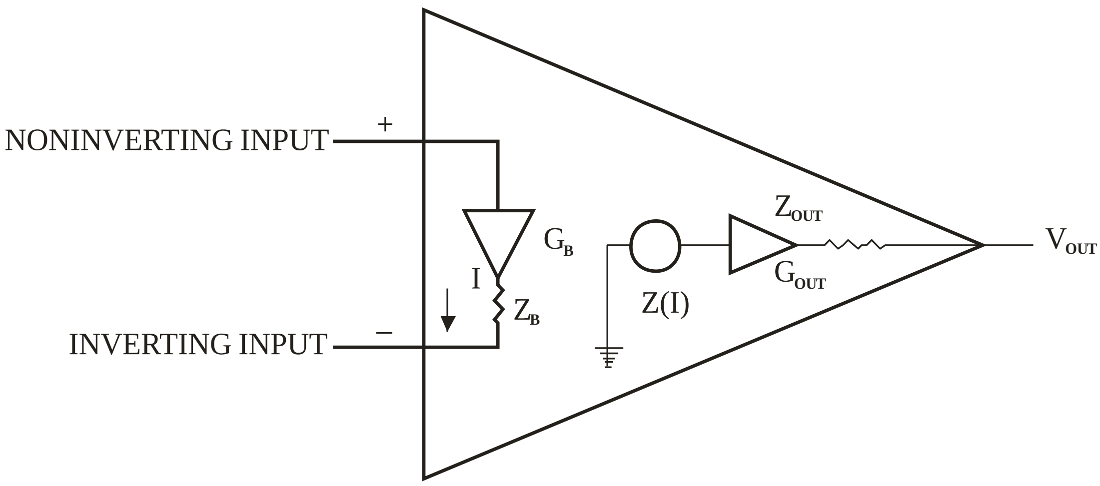


Figure 1.1 CFA Model

The output buffer provides low output impedance for the amplifier. Again, the output buffer gain, G_{OUT}, is very close to one, so it is neglected in the analysis. The output impedance of the output buffer is ignored during the calculations. This parameter may influence the circuit performance when driving very low impedance or capacitive loads, but this is usually not the case. The input buffer's output impedance can't be ignored because affects stability at high frequencies.

The current-controlled current source, Z, is a trans-impedance. The trans-impedance in a CFA serves the same function as gain in a VFA; it is the parameter that makes the performance of the op amp dependent only on the passive parameter values. Usually the trans-impedance is very high, in the MΩ range, so the CFA gains accuracy by closing a feedback loop in the same manner that the VFA does.

1.3 DEVELOPMENT OF THE STABILITY EQUATION

The stability equation is developed with the aid of Figure 1.2. Remember, stability is independent of the input, and stability depends solely on the loop gain, Aβ. Breaking the loop at point X, inserting a test signal, VTI, and calculating the return signal VTO develops the stability equation.

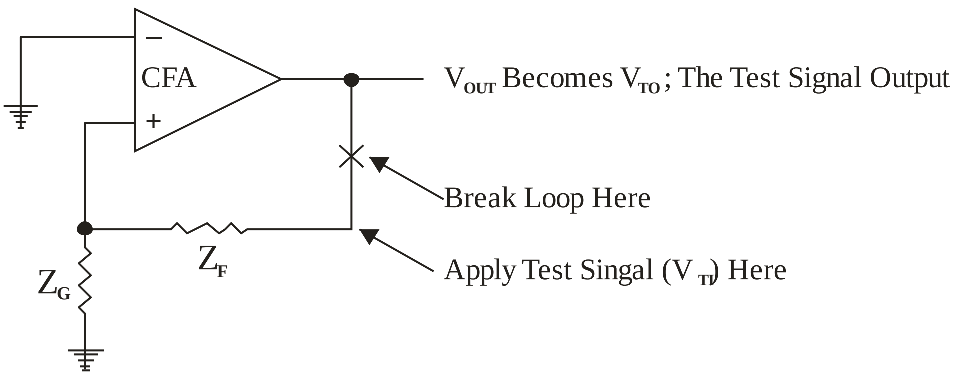


Figure 1.2 Stability Analysis Circuit

The circuit used for stability calculations is shown in Figure 1.3 where the model of Figure 1.1 is substituted for the CFA symbol. The input and output buffer gain, and output buffer output impedance have been deleted from the circuit to simplify calculations. This approximation is valid for almost all applications.

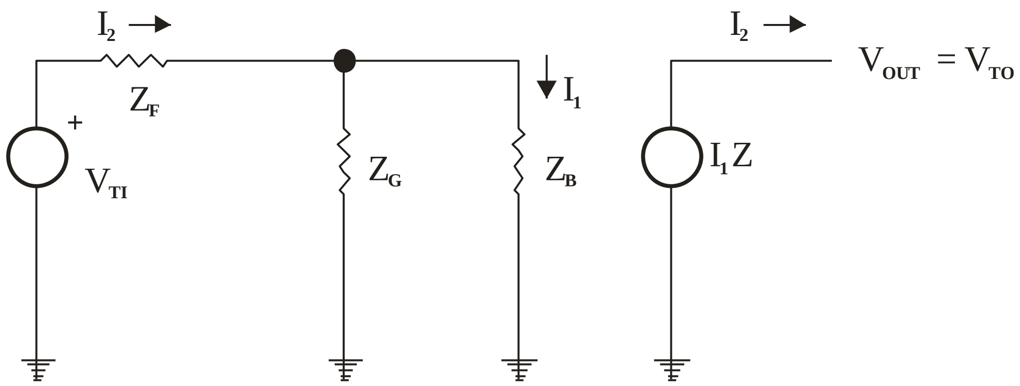


Figure 1.3 Stability Analysis of Circuit

The transfer equation is given in Equation 1.1, and the Kirchhoff's law is used to write Equations 1.2 and 1.3.

$$V_{TO} = I_1 Z \tag{1.1}$$

$$V_{TI} = I_2(Z_F + Z_G \| Z_B) \tag{1.2}$$

$$I_2(Z_G \| Z_B) = I_1 Z_B \tag{1.3}$$

Equations 1.2 and 1.3 are combined to yield Equation 1.4.

$$V_{TI} = I_1 \, (Z_F + Z_G \| Z_B)(1+\frac{Z_B}{Z_G}) = I_1 Z_F (1 + \frac{Z_B}{Z_F \| Z_G}) \tag{1.4}$$

Dividing Equation 1.1 by Equation 1.4 yields Equation 1.5 and this is the open loop transfer equation. This equation is commonly known as the loop gain.

$$A\beta = \frac{V_{TO}}{V_{TI}} = \frac{Z}{(Z_F\left(1+\frac{Z_B}{Z_F \| Z_G}\right))} \tag{1.5}$$

1.4 THE NON-INVERTING CFA

The closed-loop gain equation for the non-inverting CFA is developed with the aid of Figure 1.4, where external gain setting resistors have been added to the circuit. The buffers are shown in Figure 1.4, but because their gains equal one and they are included within the feedback loop, the buffer gain does not enter into the calculations.

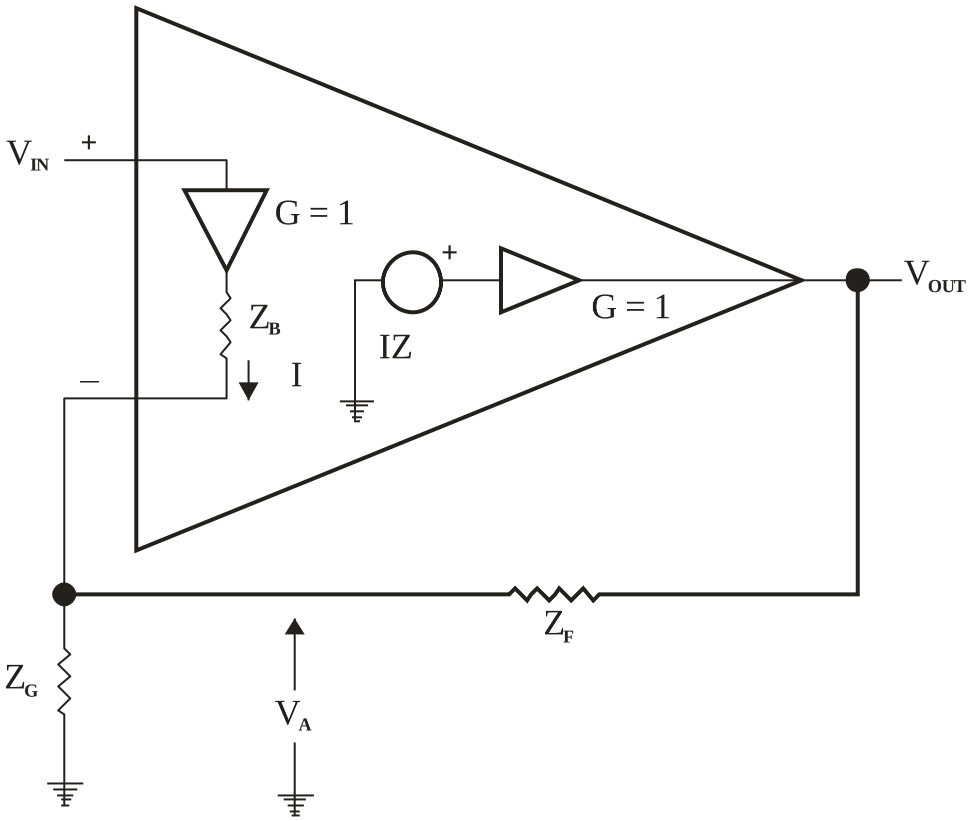


Figure 1.4. Non-inverting CFA

Equation 1.6 is the transfer equation, Equation 1.7 is the current equation at the inverting node, and Equation 1.8 is the input loop equation. These equations are combined to yield the closed-loop gain equation, Equation 1.9.

$$V_{OUT} = IZ \tag{1.6}$$

$$I = \left(\frac{V_A}{Z_G}\right) - \left(\frac{V_{OUT} - V_A}{Z_F}\right) \tag{1.7}$$

$$V_A = V_{IN} - IZ_B \tag{1.8}$$

$$\frac{V_{OUT}}{V_{IN}} = \frac{\frac{Z(1+\frac{Z_F}{Z_G})}{Z_F(1+\frac{Z_B}{Z_F \| Z_G})}}{1+\frac{Z}{Z_F(1+\frac{Z_B}{Z_F \| Z_G})}} \tag{1.9}$$

When the input buffer output impedance, Z_B, approaches zero, Equation 1.9 reduces to Equation 8–10.

$$\frac{V_{OUT}}{V_{IN}} = \frac{\frac{Z(1+\frac{Z_F}{Z_G})}{Z_F}}{1+\frac{Z}{Z_F}} = \frac{1+\frac{Z_F}{Z_G}}{1+\frac{Z_G}{Z}} \tag{1.10}$$

When the transimpedance, Z, is very high, the term Z_F/Z in Equation 1.10 approaches zero, and Equation 1.10 reduces to Equation 1.11; the ideal closed-loop gain equation for the CFA. The ideal closed-loop gain equations for the CFA and VFA are identical, and the degree to which they depart from ideal is dependent on the validity of the assumptions. The VFA has one assumption that the direct gain is very high, while the CFA has two assumptions, that the transimpedance is very high and that the input buffer output impedance is very low. As would be expected, two assumptions are much harder to meet than one, thus the CFA departs from the ideal more than the VFA does.

$$\frac{V_{OUT}}{V_{IN}} = 1 + \frac{Z_F}{Z_G} \tag{1.11}$$

1.5 THE INVERTING CFA

The inverting CFA configuration is seldom used because the inverting input impedance is very low ($Z_B||Z_F + Z_G$). When Z_G is made dominant by selecting it as a high resistance value it overrides the effect of Z_B. Z_F must also be selected as a high value to achieve at least unity gain, and high values for Z_F result in poor bandwidth performance, as we will see in the next section. If Z_G is selected as a low value the frequency sensitive Z_B causes the gain to increase as frequency increases. These limitations restrict inverting applications of the inverting CFA.

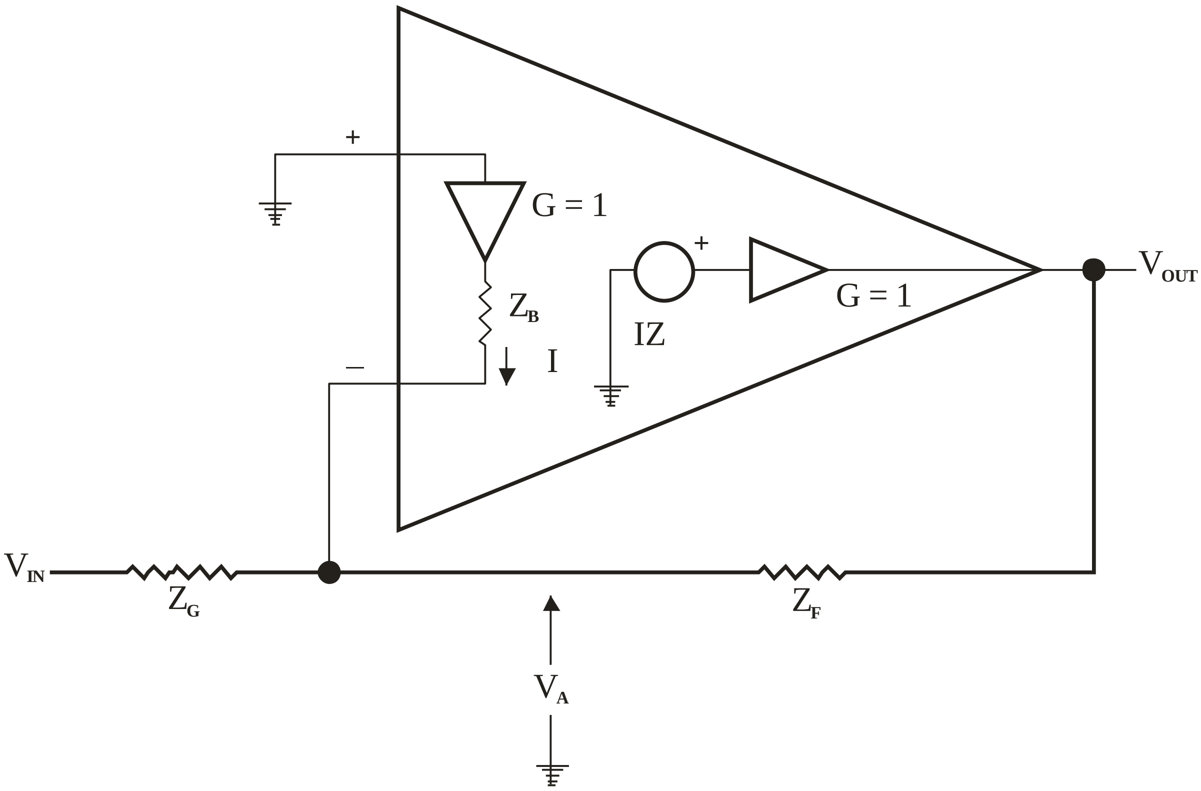


Figure 1.5 Inverting CFA

The current equation for the input node is written as Equation 1.12. Equation 1.13 defines the dummy variable, V_A, and Equation 1.14 is the transfer equation for the CFA. These equations are combined and simplified leading to Equation 1.15, which is the closed-loop gain equation for the inverting CFA.

$$I + \frac{V_{IN} - V_A}{Z_G} = \frac{V_A - V_{OUT}}{Z_F} \tag{1.12}$$

$$IZ_B = -V_A \tag{1.13}$$

$$IZ = V_{OUT} \tag{1.14}$$

$$\frac{V_{OUT}}{V_{IN}} = -\frac{\frac{Z}{Z_G\left(1+\frac{Z_B}{Z_F \| Z_G}\right)}}{1+\frac{Z}{Z_F\left(1+\frac{Z_B}{Z_F \| Z_G}\right)}} \tag{1.15}$$

When ZB approaches zero, Equation 1.15 reduces to Equation 1.16.

$$\frac{V_{OUT}}{V_{IN}} = -\frac{\frac{1}{Z_G}}{\frac{1}{Z}+\frac{1}{Z_F}} \tag{1.16}$$

When Z is very large, Equation 1.16 becomes Equation 1.17, which is the ideal closed loop gain equation for the inverting CFA.

$$\frac{V_{OUT}}{V_{IN}} = -\frac{Z_F}{Z_G} \tag{1.17}$$

The ideal closed-loop gain equation for the inverting VFA and CFA op amps are identical. Both configurations have lower input impedance than the non-inverting configuration has, but the VFA has one assumption while the CFA has two assumptions. Again, as was the case with the non-inverting counterparts, the CFA is less ideal than the VFA because of the two assumptions. The zero Z_B assumption always breaks down in bipolar junction transistors as is shown later. The CFA is almost never used in the differential amplifier configuration because of the CFA's gross input impedance mismatch.

1.6 CFOA AD844 IC INTRODUCTION

The AD844 is a high speed monolithic operational amplifier fabricated using Analog Devices' junction isolated complementary bipolar (CB) process. It combines high bandwidth and very fast large signal response with excellent dc performance. Although optimized for use in current-to-voltage applications and as an inverting mode amplifier, it is also suitable for use in many non-inverting applications.

The AD844 can be used in place of traditional op amps, but its current feedback architecture results in much better ac performance, high linearity, and an exceptionally clean pulse response. This type of op amp provides a closed-loop bandwidth that is determined primarily by the feedback resistor and is almost independent of the closed-loop gain. The AD844 is free from the slew rate limitations inherent in traditional op amps and other current-feedback op amps. Peak output rate of change can be over 2000 V/μs for a full 20 V output step. Settling time is typically 100 ns to 0.1%, and essentially independent of gain. The AD844 can drive 50Ω loads to ±2.5 V with low distortion and is short circuit protected to 80 mA. The AD844 is available in four performance grades and three package options. In the 16-lead SOIC (R) package, the AD844J is specified for the commercial temperature range of 0°C to 70°C.

The AD844A and AD844B are specified for the industrial temperature range of –40°C to +85°C and are available in the CERDIP (Q) package. The AD844A is also available in an 8-lead PDIP (N). The AD844S is specified over the military temperature range of –55°C to +125°C. It is available in the 8-lead CERDIP (Q) package. A and S grade chips and devices processed to MIL-STD-883B, REV. C are also available.

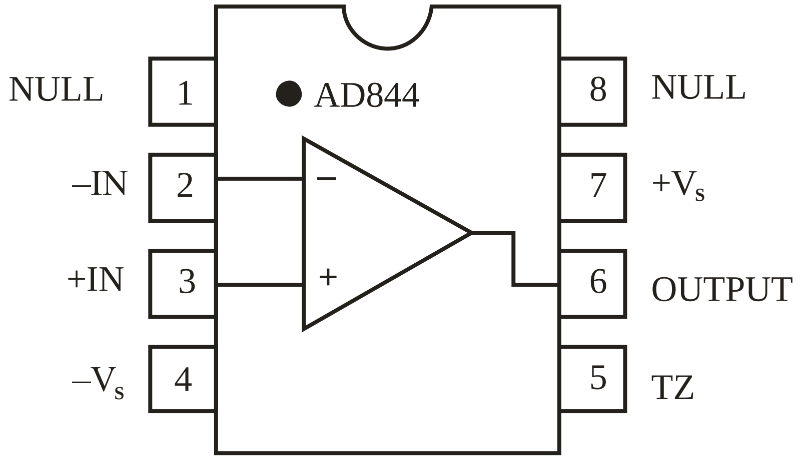


FIGURE 1.6 AD844 IC PIN CONFIGARATION

1.7 PRODUCT HIGHLIGHTS

1. The AD844 is a versatile, low cost component providing an excellent combination of ac and dc performance.

2. It is essentially free from slew rate limitations. Rise and fall times are essentially independent of output level.

3. The AD844 can be operated from $\pm$4.5 V to $\pm$18 V power supplies and is capable of driving loads down to 50Ω, as well as driving very large capacitive loads using an external network.

4. The offset voltage and input bias currents of the AD844 are laser trimmed to minimize dc errors; VOS drift is typically 1μV/°C and bias current drift is typically 9 nA/°C.
5. The AD844 exhibits excellent differential gain and differential phase characteristics, making it suitable for a variety of video applications with bandwidths up to 60 MHz.

6. The AD844 combines low distortion, low noise, and low drift with wide bandwidth, making it outstanding as an input amplifier for flash A/D converters.

1.8 UNDERSTANDING THE AD844

The AD844 can be used in ways similar to a conventional op amp while providing performance advantages in wideband applications. However, there are important differences in the internal structure that need to be understood in order to optimize the performance of the AD844 op amp.

1.8.1 Open-Loop Behaviour

Figure 1.7 shows a current feedback amplifier reduced to essentials. Sources of fixed dc errors, such as the inverting node bias current and the offset voltage, are excluded from this model and are discussed later. The most important parameter limiting the dc gain is the transresistance, R_t, which is ideally infinite. A finite value of R_t is analogous to the finite open-loop voltage gain in a conventional op amp. The current applied to the inverting input node is replicated by the current conveyor so as to flow in resistor R_t. The voltage developed across R_t is buffered by the unity gain voltage follower.

Voltage gain is the ratio R_t/R_{IN}. With typical values of R_t = 3 MΩ and R_{IN} = 50Ω, the voltage gain is about 60,000. The open loop current gain is another measure of gain and is determined by the beta product of the transistors in the voltage follower stage (see Figure 4); it is typically 40,000.

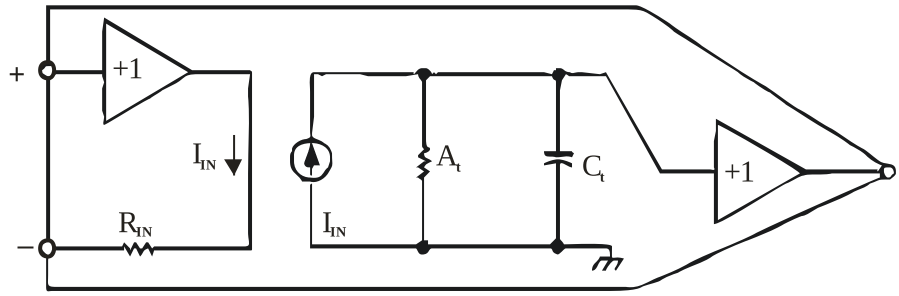


Figure 1.7 Equivalent Schematic

The important parameters defining ac behaviour are the transcapacitance, C_t, and the external feedback resistor (not shown). The time constant formed by these components is analogous to the dominant pole of a conventional op amp and thus cannot be reduced below a critical value if the closed-loop system is to be stable. In practice, C_t is held to as low a value as possible (typically 4.5 pF) so that the feedback resistor can be maximized while maintaining a fast response. The finite R_{IN} also affects the closed-loop response in some applications as will be shown. The open-loop ac gain is also best understood in terms of the transimpedance rather than as an open-loop voltage gain. The open-loop pole is formed by R_t in parallel with C_t. Since C_t is typically 4.5 pF, the open-loop corner frequency occurs at about 12 kHz. However, this parameter is of little value in determining the closed-loop response.

1.8.2 Circuit Description of the AD844

A simplified schematic is shown in Figure 4. The AD844 differs from a conventional op amp in that the signal inputs have radically different impedance. The non-inverting input (Pin 3) presents the usual high impedance. The voltage on this input is transferred to the inverting input (Pin 2) with a low offset voltage, ensured by the close matching of like polarity transistors operating under essentially identical bias conditions. Laser trimming nulls the residual offset voltage, down to a few tens of microvolts. The inverting input is the common emitter node of a complementary pair of grounded base stages and behaves as a current summing node. In an ideal current feedback op amp, the input resistance would be zero.

In the AD844, it is about 50Ω A current applied to the inverting input is transferred to a complementary pair of unity-gain current mirrors that deliver the same current to an internal node (Pin 5) at which the full output voltage is generated. The unity-gain complementary voltage follower then buffers this voltage and provides the load driving power. This buffer is designed to drive low impedance loads, such as terminated cables, and can deliver ±50 mA into a 50Ω load while maintaining low distortion, even when operating at supply voltages of only ±6 V. Current limiting (not shown) ensures safe operation under short circuited conditions.

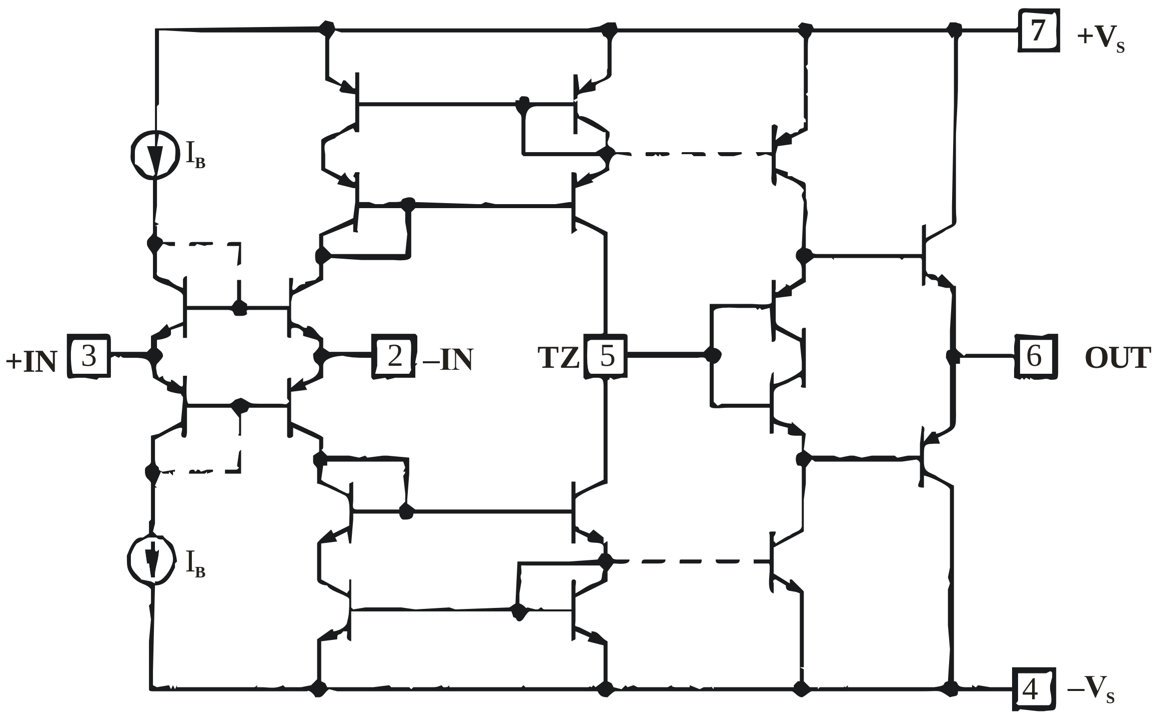


Figure 1.8 Simplified Schematic

It is important to understand that the low input impedance at the inverting input is locally generated and does not depend on feedback. This is very different from the “virtual ground” of a conventional operational amplifier used in the current summing mode which is essentially an open circuit until the loop settles. In the AD844, transient current at the input does not cause voltage spikes at the summing node while the amplifier is settling.

Furthermore, all of the transient current is delivered to the slewing (TZ) node (Pin 5) via a short signal path (the grounded base stages and the wideband current mirrors). The current available to charge the capacitance (about 4.5 pF) at the TZ node is always proportional to the input error current, and the slew rate limitations associated with the large signal response of the op amps do not occur. For this reason, the rise and fall times are almost independent of signal level. In practice, the input current will eventually cause the mirrors to saturate. When using ± 15 V supplies, this occurs at about 10 mA (or ± 2200 V/μs). Since signal currents are rarely this large, classical "slew rate" limitations are absent.

This inherent advantage would be lost if the voltage follower used to buffer the output were to have slew rate limitations. The AD844 has been designed to avoid this problem, and as a result, the output buffer exhibits a clean large signal transient response, free from anomalous effects arising from internal saturation.

1.9 THEORY OF OSCILLATOR

1.9.1 Introduction

There are many types of oscillators, and many different circuit configurations that produce oscillations. Some oscillators produce sinusoidal signals, others produce non-sinusoidal signals. Non-sinusoidal oscillators, such as pulse and ramp (or saw-tooth) oscillators, find use in timing and control applications. Pulse oscillators are commonly found in digital-systems clocks, and ramp oscillators are found in the horizontal sweep circuit of oscilloscopes and television sets. Sinusoidal oscillators are used in many applications, for example, in consumer electronic equipment (such as radios, TVs, and VCRs), in test equipment (such as network analyzers and signal generators), and in wireless systems.

In this chapter the feedback approach to oscillator design is discussed. The oscillator examples selected in this chapter, as well as the mix of theory and design information presented, help to clearly illustrate the feedback approach. The basic components in a feedback oscillator are the amplifier, amplitude limiting component, a frequency-determining network, and a (positive) feedback network. Usually the amplifier also acts as the amplitude-limiting component, and the frequency-determining network usually performs the feedback function. The feedback circuit is required to return some of the output signal back to the input. Positive feedback occurs when the feedback signal is in phase with the input signal and, under the proper conditions, oscillation is possible.

1.9.2 Oscillation Conditions

A basic feedback oscillator is shown in Figure 1.9. The amplifier's voltage gain is Av (jω), and the voltage feedback network is described by the transfer function β (jω). The amplifier gain Av (jω) is also called the open-loop gain since it is the gain between V_o and V_i when V_f = 0.

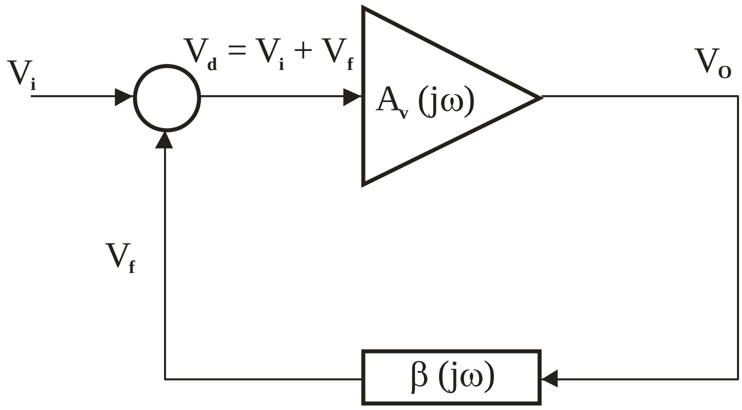


Figure 1.9 basic feedback circuit

(i.e., when the path through β (jω) is properly disconnected). The amplifier gain is, in general, a complex quantity. However, in many oscillators, at the frequency of oscillation, the amplifier is operating in its mid-band region where A_v (jω) is a real constant. When A_v (jω) is constant, it is denoted by A_{vo}.

Negative feedback occurs when the feedback signal subtracts from the input signal. On the other hand, if V_f adds to V_i , the feedback is positive. The summing network in Figure 1.9 shows the feedback signal added to V_i to suggest that the feedback is positive. Of course, the phase of V_f determines if V_f adds or subtracts to V_i . The phase of V_f is determined by the closed-loop circuit in Figure 1.9. If A_v (jω) 0^o A_{vo} and A_{vo}

is a positive number, the phase shift through the amplifier is 0^o, and for positive feedback the phase through β (jω) should be 0^o (or a multiple of 360^o). If A_{vo} is a negative number, the phase shift through the amplifier is ±180° and the phase through β (jω) for positive feedback should be $\pm180° \pm n360°$. In other words, for positive feedback the total phase shift associated with the closed loop must be 0^o or a multiple n of 360^o. From Figure 1.9 we can write

$$v_o = A_v(j\omega)\, v_d \tag{1.18}$$

$$v_f = \beta(j\omega)\, v_o \tag{1.19}$$

and

$$v_d = v_i + v_f \tag{1.20}$$

Thus, from (1.18) to (1.20), the closed-loop voltage gain A_{vf} (jω) is given by

$$A_{vf}(j\omega) = \frac{v_o}{v_i} = \frac{Av(j\omega)}{1-\beta(j\omega)Av(j\omega)} \tag{1.21}$$

The quantity β (jω) A_v (j ω) is known as the loop gain. For oscillations to occur, an output signal must exist with no input signal applied. With v_i =0 in (1.21) it follows that a finite v_o is possible only when the denominator is zero. That is, when

$$1 - \beta\,(j\omega)A_v(j\omega) = 0$$

Or

$$\beta\,(j\omega)A_v(j\omega) = 1 \tag{1.22}$$

Equation (1.22) expresses the fact that for oscillations to occur the loop gain must be unity. This relation is known as the Barkhausen criterion. With A_v (jω) =A_{vo} and letting

$$\beta(j\omega) = \beta_r(\omega) + j\beta_i(\omega)$$

where $\beta_r(\omega)$ and $\beta_i(\omega)$ are the real and imaginary parts of $\beta(j\omega)$, we can express in the form

$$\beta_r(\omega)A_{vo} + j\beta_i(\omega)A_{vo} = 1$$

Equating the real and imaginary parts on both sides of the equation gives

$$\beta_r(\omega)A_{vo} = 1 \Rightarrow A_{vo} = \frac{1}{\beta_r(\omega)} \quad (1.23)$$

and

$$\beta_i(\omega)A_{vo} = 0 \Rightarrow \beta_i(\omega) = 0 \quad (1.24)$$

since $A_{vo} \neq 0$. The conditions in (1.23) and (1.24) are known as the Barkhausen criteria in rectangular form for $A_v(j\omega) = A_{vo}$. The condition (1.23) is known as the gain condition, and (1.24) as the frequency of oscillation condition. The frequency of oscillation condition predicts the frequency at which the phase shift around the closed loop is 0^o or a multiple of 360^o. The relation (1.22) can also be expressed in polar form as

$$\beta(j\omega)A_v(j\omega) = |\beta(j\omega)A_v(j\omega)| \angle \beta(j\omega)A_v(j\omega) = 1 \quad (1.25)$$

and

$$\angle \beta(j\omega)A_v(j\omega) = \pm n360° \quad (1.26)$$

where n = 0, 1, 2, ... Equation (1.26) expresses the fact that the signal must travel through the closed loop with a phase shift of 0^o or a multiple of 360^o. For $A_v(j\omega) = A_{vo}$, then $\angle \beta(j\omega) A_{vo}$ is the angle of $\beta(j\omega)$, and the condition (1.26) is equivalent to saying that $\beta_i(j\omega) = 0$, in agreement with (1.24). Also, for $A_v(j\omega) = A_{vo}$ and with $\beta_i(j\omega) = 0$, (1.25) reduces to (1.23). The conditions in (1.25) and (1.26) are known as the Barkhausen criteria in polar form.

When the amplifier is a current amplifier, the basic feedback network can be represented as shown in Figure 1.10. In this case, $A_i(j\omega)$ is the current gain of the amplifier, and the current feedback factor $\alpha(j\omega)$ is

$$\alpha(j\omega) = \frac{i_f}{i_o}$$

For this network, the condition for oscillation is given by

$$\alpha(j\omega)A_i(j\omega) = 1 \tag{1.27}$$

which expresses the fact that loop gain in Figure 1.10 must be unity. The loop gain can be evaluated in different ways. One method that can be used in some oscillator configurations is to determine $A_v(j\omega)$ and $\beta(j\omega)$ and to form the loop gain $A_v(j\omega)\beta(j\omega)$. In many cases it is not easy to isolate $A_v(j\omega)$ and $\beta(j\omega)$ since they are interrelated. In such cases a method that can usually be implemented is to represent the oscillator circuit as a continuous and repetitive circuit. Hence, the loop gain is calculated as the gain from one part to the same part in the following circuit. An alternate analysis method is to replace the amplifier and feedback network in Figure 1.9 by their ac models and write the appropriate loop equations. The loop equations form a system of linear equations that can be solved for the closed-loop voltage gain, which can be expressed in the general form

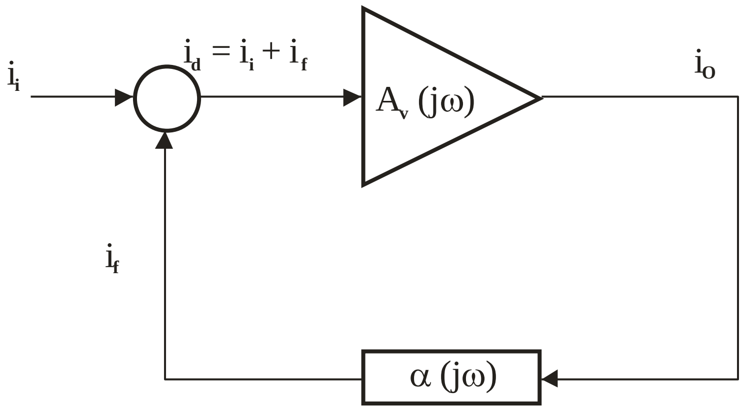


Figure 1.10 The current form of the basic current feedback network.

$$A_{vf}(j\omega) = \frac{v_o}{v_i} = \frac{N(j\omega)}{D(j\omega)} \tag{1.28}$$

Where $N(j\omega)$ represents the numerator polynomial and $D(j\omega)$ is the system determinant of the linear equations. In terms of (1.11) the conditions for oscillations are obtained by setting the system determinant equal to zero (i.e., $D(j\omega)=0$). Setting $D(j\omega)=0$ results in two equations: one for the real part of $D(j\omega)$ (which gives the gain condition), and one for the imaginary part of $D(j\omega)$ (which gives the frequency of oscillation). From circuit theory we know that oscillation occurs when a network has a pair of complex conjugate poles on the imaginary axis. However, in electronic oscillators the poles are not exactly on the imaginary axis because of the nonlinear nature of the loop gain. There are different nonlinear effects that control the pole location in an oscillator. One nonlinear mechanism is due to the saturation characteristics of the amplifier. A saturation-limited sinusoidal oscillator works as follows. To start the oscillation, the closed-loop gain in (1.21) must have a pair of complex-conjugate poles in the right-half plane. Then, due to the noise voltage generated by thermal vibrations in the network (which can be represented by a superposition of input noise signals v_n) or by the transient generated when the dc power supply is turned on, a growing sinusoidal output voltage appears.

The characteristics of the growing sinusoidal signal are determined by the complex conjugate poles in the right-half plane. As the amplitude of the induced oscillation increases, the amplitude-limiting capabilities of the amplifier (i.e., a reduction in gain) produce a change in the location of the poles. The changes are such that the complex-conjugate poles move towards the imaginary axis. However, the amplitude of the oscillation was increasing and this makes the complex poles to continue the movement toward the left-half plane. Once the poles move to the left-half plane the amplitude of the oscillation begins to decrease, moving the poles toward the right-half plane. The process of the poles moving between the left-half plane and the right-half plane repeats, and some steady-state oscillation occurs with a fundamental frequency, as well as harmonics. This is a nonlinear process where the fundamental frequency of oscillation and the harmonics are determined by the location of the poles. Although the poles are not on the imaginary axis, the Barkhausen criterion in (1.22) predicts fairly well the fundamental frequency of oscillation. It can be considered as providing the fundamental frequency of the oscillator based on some sort of average location for the poles.

The movement of the complex conjugate poles between the right-half plane and the left-half plane is easily seen in an oscillator designed with an amplitude limiting circuit that controls the gain of the amplifier and, therefore, the motion of the poles. An example to illustrate this effect is given in Example 1.23. The previous discussion shows that for oscillations to start the circuit must be unstable (i.e., the circuit must have a pair of complex-conjugate poles in the right-half plane). The condition (1.22) does not predict if the circuit is unstable. However, if the circuit begins to oscillate, the Barkhausen criterion in (1.22) can be used to predict the approximate fundamental frequency of oscillation and the gain condition.

1.10 SUMMARY

In the above topic we have seen CFA produce higher oscillation rather than VFA which is requirement of current world. We have seen basic operation of CFA and AD844 8-pin IC. How this 8-IC work. We also do study on oscillator and see what is the condition of oscillations and how we get that, what is feedback and how it generates constant oscillations and we study Barkhausen criteria also.

For most general purpose or high precision low frequency, low noise applications, the VFB op amp is usually the best choice. VFB op amps also work well in single-supply applications because many are available with rail-to-rail inputs and outputs. The VFB op amp has a very flexible feedback network, and is therefore suitable for active filter designs. The CFB op amp offers the ultimate in bandwidth, slew rate, and distortion performance at the expense of dc performance, noise, and the requirement for a fixed value feedback resistor.

CHAPTER 2

LITERATURE REVIEW

2.1 LITERATURE REVIEW ON CFOA OSCILLATOR

Current-feedback amplifiers (CFA) do not have the traditional differential amplifier input structure, thus they sacrifice the parameter matching inherent to that structure. The CFA circuit configuration prevents them from obtaining the precision of voltage-feedback amplifiers (VFA), but the circuit configuration that sacrifices precision results in increased bandwidth and slew rate. The higher bandwidth is relatively independent of closed-loop gain, so the constant gain-bandwidth restriction applied to VFAs is removed for CFAs. The slew rate of CFAs is much improved from their counterpart VFAs because their structure enables the output stage to supply slewing current until the output reaches its final value.

In general, VFAs are used for precision and general purpose applications, while CFAs are restricted to high frequency applications above 100 MHz. Although CFAs do not have the precision of their VFA counterparts, they are precise enough to be dc-coupled in video applications where dynamic range requirements are not severe. CFAs, unlike previous generation high-frequency amplifiers, have eliminated the ac coupling requirement; they are usually dc-coupled while they operate in the GHz range. CFAs have much faster slew rates than VFAs, so they have faster rise/fall times and less intermodulation distortion.

The name, operational amplifier, was given to voltage-feedback amplifiers (VFA) when they were the only op amps in existence. These new (they were new in the late ’40s) amplifiers could be programmed with external components to perform various math operations on a signal; thus, they were nicknamed op amps. Current-feedback amplifiers (CFA) have been around approximately twenty years, but their popularity has only increased in the last several years. Two factors limiting the popularity of CFAs are their application difficulty and lack of precision.

The VFA is familiar component, and there are several variations of internally compensated VFAs that can be used with little applications work. Because of its long history, the VFA comes in many varieties and packages, so there are VFAs applicable to almost any job. VFA bandwidth is limited, so it can't function as well at high signal frequencies as the CFA can. For now, the signal frequency and precision separates the applications of the two op amp configurations. The VFA has some other redeeming virtues, such as excellent precision, that makes it the desirable amplifier in low frequency applications. Many functions other than signal amplification are accomplished at low frequencies, and functions like level-shifting a signal require precision. Fortunately, precision is not required in most high frequency applications where amplification or filtering of a signal is predominant, so CFAs are suitable to high frequency applications. The lack of precision coupled with the application difficulties prevents the CFA from replacing the VFA.

The voltage-feedback operational amplifier (VOA) is used extensively throughout the electronics industry, as it is an easy to use and versatile analogue building block. The architecture of the VOA has several attractive features, such as the differential long-tail pair, high-impedance input stage, which is very good at rejecting common-mode signals. Unfortunately, the architecture of the VOA provides inherent limitations in both the gain-bandwidth trade-off and slew-rate. Typically the gain-bandwidth product (f_T) is a constant and the slew-rate is limited to a maximum value determined by the ratio of the input stage bias current to the dominant-pole compensation capacitor. The current-feedback operational amplifier (CFOA) design has a fundamentally different topology. The most significant advantage of the CFOA over the VOA is in slew rate performance. A very high slew-rate is obtained as a direct result of the use of current as the feedback error signal. Typically values from 500 to 2500 V/ps are quoted by manufacturers of CFOAs, whereas the slew-rates of VOAs are generally much lower, typically in the region of 1 to 100 V/ps. Slew-rate limiting is a major cause of high frequency distortion in high-frequency amplifiers that are handling output signal levels, such as in video signal processing applications. Comparing the input-referred noise performance of the two operational amplifier topologies the CFOA has a lower figure due mainly to the fact that, for the same slew rate performance, the VOA will inevitably use high collector-current bias values with emitter degeneration and hence this increases the relative input-referred noise of the VOA.

The current-feedback operational amplifier (CFOA) is a relatively new arrival in the analogue designer's tool kit. The first monolithic device was produced by Elantec Inc. in 1987 and since then it has intrigued and puzzled many. Is it really better than the conventional and well known voltage-feedback operational amplifier (VOA)? How does the CFOA differ in design? What are the applications where it performs batter, than the VOA and when should the VOA be used in preference to the CFOA? We will attempt to give some insight into the design and performance of the CFOA and to answer these questions[1]. Two second-generation current conveyor (CCII+)- based resistance–capacitance (RC) square/triangular waveform generators, which have been derived from their voltage-mode op-amp-based schemes, with independent control of frequency are presented in this paper. Each configuration consists of two positive second-generation conveyors (CCII(+)-"A" and CCII(+)-"B"), three resistors, and one floating capacitor that is responsible for better linearity. The frequency of the waveform generators can independently be adjusted with any passive device. The circuits were built with commercially available current feedback operational amplifiers (AD844) and passive components used externally and tested for waveform generation and tenability [2]. Square/triangular waveform generators play an important role in instrumentation, communication systems, and signal processing applications.

A three-operational trans-conductance amplifier (OTA)-based design has recently been reported. The design uses OTAs as switching elements and controls the frequency by dc bias current. It has also been seen that the second-generation current conveyor (CCII) is very useful as an analog building block and is receiving regular attention in waveform generators, oscillators, and design of filters. The current conveyor relies upon the ability of the circuit to act as a voltage buffer between its inputs and upon the ability to convey current between two ports at extremely different impedance levels. There are varieties of oscillator circuits that are realized with CCIIs. The present trend is to decrease the number of passive elements in the circuit while keeping the number of active devices to a maximum of three. However, many applications require independent control of amplitude and frequency.

Square/triangular wave generators can easily be realized by using a current source to charge and discharge a grounded timing capacitor followed by a Schmitt trigger. These types of generators feature simple configuration and wide sweep capability. Several circuits can be found in the literature, using either a voltage-mode Schmitt trigger formed from a voltage comparator and a pair of resistors [3], [4] or a current-mode Schmitt trigger, which was first reported in, used as a single current feedback operational amplifier (CFOA) with a pair of resistors and a grounded capacitor [5], [6]. The design of a sinusoidal, triangular, and square wave generator using a high speed current feedback operational amplifier (CFOA). A macro model for the CFOA is developed as well. The frequency of oscillation in all cases are derived and compared with the simulated values.

The achieved frequency of oscillation for sinusoidal wave is 9.4 MHz to 15 MHz and the achieved frequency of oscillation for the triangular/rectangular wave is 6.75 MHz to 59.5 MHz. The advantage of using the current feedback operational amplifier is to get higher frequency of sinusoidal, triangular, and square wave [7]. Use of high speed CFOAs in analog signal processing offers several advantages over voltage feedback operational amplifier (VFOA). These advantages are wide bandwidth (which is relatively independent of the closed loop gain) [8],[9], high slew rate, and ease of realizing various functions with the least possible number of external passive components. They do not require any component-matching requirements [9]. There is also a growing demand/interest in realization of active filter [10], sinusoidal oscillators using CFOA [11-17]. The maximum frequency of operation of any circuit depends on the bandwidth of the amplifier. Signal generator build with a CFOA will operate at higher frequencies than that of a VFOA.

Today's electronic systems require many signal waveform shapes in addition to the sinusoidal waveform. Common waveforms are the square wave, triangular wave, and single pulse wave with fixed duration. Fixed duration pulses are used in communication and control systems. Square waves are used as a clock for digital systems. Triangular waves are used for scanning an electron beam on a CRT screen, in precise time measurements, and in time modulation [19]. This paper presents the generation and analysis of sinusoidal, triangular, and square wave generator using CFOA. Sinusoidal, triangular and square wave generator using CFOA has been presented. The advantages of using CFOA over VFOA in designing oscillator also presented. Theoretical oscillation frequency agrees with the simulated oscillation frequency.

An improved performance current feedback amplifier circuit referred to as CFA feedback current feedback amplifier (CCFA) is described in this paper. This circuit is developed as an augmentation to the operational current feedback amplifier which employs a technique involving the incorporation of the input circuit of the current feedback amplifier in the feedback loop of an operational amplifier to reduce the input impedance at the inverting terminal of the current feedback amplifier. The proposed circuit replaces the operational amplifier with AD844 CFA and the circuit possesses the gain accuracy and bandwidth of the current feedback amplifier. The new realization significantly enhanced the bandwidth accuracy and bandwidth gain-independence. The simulation of the proposed circuit has been performed using PSPICE. Various configurations using AD844 CFA and OPA have been simulated and the satisfactory results were obtained [29]. The current feedback amplifiers (CFAs) are getting popularity as alternative building blocks for analog wave processing, because they offer the following advantages over the conventional op-amps: (i) wide bandwidth which is relatively independent of the close-loop-gain, (ii) very high slew rate and (iii) ease of realizing various functions with the least possible number of external passive components, without any component-matching requirement [30], [31], [32]. The basic circuit topology of CFA results in a high slew rate as well as a constant bandwidth feature (bandwidth gain-independence) which makes the CFA particularly applicable in systems requiring variable closed-loop gains with constant bandwidth such as in automatic gain control [34],[35][36]. Actually, the constant bandwidth property is only the approximation. This arises because the input impedance at the inverting input of the CFA is typically not sufficiently low and as a result the closed-loop bandwidth depends on and other circuit parameters [37]. A new circuit which was referred to as CCFA was proposed and various simulations confirmed the performance in both inverting and non inverting configuration. The proposed circuit is the virtual elimination of the CFA inverting input resistance from the system transfer function. The overall bandwidth of CFA is independent of input resistance R1. The new circuit was found superior to the previously reported CFOA and CFA in terms of gain bandwidth accuracy and several other factors. The CCFA based circuits are suitable for mixed mode signal processing especially for differentiator/integrator blocks at high frequency applications.

Some new active Voltage Controlled RC integrator and differentiator circuit realizations, with both single and differential input capabilities, using a few passive components and a current feedback amplifier (CFA) device, are proposed. A multiplier (ICL - 8013) element has been appropriately utilized in the feed forward / feedback connection to obtain electronic tuning of the time constant (τ_o) by the d.c. control voltage (Vc) of the multiplier. Experimental result on wave processing had been verified in the frequency range of 50 KHz ~ 300 KHz. The active τ o sensitivities in the event of non-ideal CFA are shown to be extremely low. Some new CFA- RC voltage controlled integrator / differentiator (VCI / VCD) circuits having single or dual- input capability are proposed; the feature of electronic tuning of τ o had been obtained through the d.c. control voltage (Vc) of a multiplier element incorporated suitably in the configuration. The reliability equations are derived for both ideal and non-ideal CFA device. The proposed circuits exhibited satisfactory sinusoid response at extended frequency ranges with the expected attenuation of 20 dB/decade is in [35].

A set comprising an active RC integrator and differentiator with time constant multiplication is presented. The proposed integrator and the differentiator can be used for low-frequency signal processing applications. Moreover, no extremely large-valued passive components are needed. In the integrator, the unlimited multiplication of time constants is allowed. Both have been implemented using commercially available current feedback amplifiers. Experimental results demonstrate the theoretical predictions in [37]. A set comprising an active RC integrator and differentiator with time constant multiplication using CFAs has been developed. Since the equivalent resistors have at least a double multiplication effect, it is also suitable to realise low-frequency applications.

Although a number of new building blocks and concepts related to current-mode circuits have been investigated in literature [1-3], the use of current feedback operational amplifiers (CFOAs) as an alternative to the traditional voltage-mode op-amps (VOA), has attracted considerable attention (see [4-6, 16] and the references cited therein) in various instrumentation, signal processing and signal generation applications due to their commercial availability as off-the-shelf ICs as well as due to the well known advantages offered by CFOAs over the VOAs [4-6]. Because of these reasons, use of CFOAs has been extensively investigated in realizing oscillators, for instance, see [6-9, 17] and the references cited therein. Although, a variety of CFOAs are available from various manufacturers, AD844 (from Analog Devices) which contains a CCII+ followed by a voltage buffer is particularly flexible and popular due to the availability of z-terminal of the CCII+ therein as an externally accessible lead which permits AD844 to be used as a CCII+ (one AD844) or as CCII- (realizable with two AD844s) or as a general 4-terminal building block [6]. This paper is concerned with the use of CFOAs in the realization of Voltage-Controlled Oscillators (VCO) which are required in several instrumentation, electronic and communication systems, such as in function generators, in production of electronic music to generate variable tones, in phase locked loops and in frequency synthesizers [10-15].

Although recently, the use of CFOAs in the realization of linear VCOs has been reported in [16] and [17], in this paper, the voltage summing property of the commercially available analog multiplier (AM) AD 534 has been exploited to yield new VCO circuits possessing properties not available in the previously known CFOA-AM VCOs. It may be recalled that, in [16], all existing two CFOA-based single resistance controlled oscillators (SRCO) have been shown to be convertible into VCOs by using FET-based linearized voltage-controlled resistor (VCR), thereby, leading to VCO circuits employing two to three CFOAs, however, in such circuits, the tuning law between oscillation frequency and control voltage is not line ar. More recently, a number of CFOA-based VCOs have been proposed in the literature [17-18] in which the VCO realization has been achieved by devising a CFOA-based RC-active oscillator with two analog multipliers (AM) appropriately embedded to enable independent control of the oscillation frequency through an external control voltage VC applied as a common multiplicative input to both the multipliers such that oscillation frequency becomes a linear function of VC i.e. $f_O \propto V_c$ Whereas VCO circuits of [17] all employ only two CFOAs, by contrast, two VCOs reported in [18] employ only a single CFOA along with three resistors, two grounded capacitors and two AMs. These circuits, however, suffer from the difficulty of employing a variable capacitance for adjusting the circuits to produce oscillations, which is not a very desirable option. Another disadvantage of the circuits of [18] is the requirement of matching of two resistor values which is also not desirable since any mismatch in component values may change the frequency of oscillation or disturb the oscillations altogether. The object of this paper is to present a family of eight new single CFOA-based VCOs which remove the above mentioned difficulties while employing a bare minimum number of active and passive components namely only one CFOA, only two multipliers (essential for obtaining linear control of oscillation frequency), two/three resistors and two capacitors. Experimental results using AD844 CFOAs and AD534 type AMs have been given which confirm the practical viability of the proposed circuits. The advantages of the new CFOA-based VCOs as compared to the previously known CFOA-based VCOs are highlighted. AD844 type CFOA is a 4-port analog building block with x, y being the input terminals and z and w being output terminals which is characterized by the terminal equations $i_y=0$, $v_x= v_y$, $i_z= i_x$, $v_w=v_z$ and its symbolic notation is shown in Fig. 1(a).

A novel current-feedback operational amplifier, the input stage of which is based on the design and use in a repeated pattern of a current transfer cell, exhibits performance characteristics superior to those obtained with an established input architecture: in particular, the common-mode rejection ratio (CMRR) is 105 dB, and the DC offset voltage less than 200 mV. For many years the voltage feedback operational amplifier (VFOA) featured prominently in the design of analogue signal processing systems but, in the 1980s, the requirements of certain sections of the telecommunications industry saw the emergence of the current feedback operational amplifier (CFOA) having a high slew rate and a wide bandwidth that was almost independent of closed-loop voltage gain. However, currently established CFOA designs exhibit a common-mode rejection ratio (CMRR) that is only modest compared with that possible with VFOAs [39].

Despite the age of operational amplifiers as an idea [l], their application and performance continues to thrive and prosper. In particular, the widespread availability of very high speed complementary bipolar processes [2] are sending high speed amplifier designs in a myriad of directions. Designs are being pushed by end applications requiring higher performance. End applications are in turn being pushed by the availability of high speed operational amplifiers which allow the applications to step up and take advantage of the new reality. With this in mind, several aspects of modern high speed, +/- 5 volt operational amplifier design will be discussed. Voltage feedback and current feedback topologies will be addressed with particular emphasis on recent advances and showing the progress of the architectures as time has progressed. Multistage amplifiers, unity gain buffers, and solutions to the low power design problem will be emphasized as well. In addition to wide bandwidth, low distortion is a particularly important design issue and will be addressed separately. Here, the emphasis will be on the communality of the major distortion sources between all of the architectures. Current feedback is not a particularly new idea, dating back at least to vacuum tube amplifiers of the 1940's [3]. In transistor circuits, current feedback dates back to the late sixties where it was applied to the design of instrumentation amplifiers. Although these amplifiers had low gain bandwidth trade-off, they were predominately class-A devices and did not provide the high slew rate that has come to symbolize current feedback today. Perhaps the first amplifier that utilized complementary devices to provide both low gain-bandwidth trade-off and high slew rate appeared in Bob Eckes' Ph.D. dissertation in 1967[40].

A novel CMOS low-voltage current feedback operational amplifier (CFOA) is presented. The proposed CFOA based on a new positive second-generation current conveyor (CCII+). The new circuit allows almost a rail-to-rail input and output operation; also, it reduces the offset voltage and provides high driving current capabilities. The CFOA is operating at supply voltages of ±0.75 V with a total standby current of 338 μA. The circuit exhibits better than 10 MHz bandwidth and 1mA current drive capability. A new CMOS CFOA was presented, analyzed and simulated. The CFOA has improved the input stage open-loop bandwidth and reduced the voltage transfer error. The CFOA block is suitable for low-voltage, low-power applications and characterized by the ability to achieve small voltage, current transfer errors and high output driving current capability [41].

The AD844 is a high speed monolithic operational amplifier fabricated using Analog Devices' junction isolated complementary bipolar (CB) process. It combines high bandwidth and very fast large signal response with excellent dc performance. Although optimized for use in current-to-voltage applications and as an inverting mode amplifier, it is also suitable for use in many non-inverting applications. The AD844 can be used in place of traditional op amps, but its current feedback architecture results in much better ac performance, high linearity, and an exceptionally clean pulse response. This type of op amp provides a closed-loop bandwidth that is determined primarily by the feedback resistor and is almost independent of the closed-loop gain. The AD844 is free from the slew rate limitations inherent in traditional op amps and other current-feedback op amps. Peak output rate of change can be over 2000 V/μs for a full 20 V output step. Settling time is typically 100 ns to 0.1%, and essentially independent of gain. The AD844 can drive 50 Ω loads to 2.5 V with low distortion and is short circuit protected to 80 mA. The AD844 is available in four performance grades and three package options. In the 16-lead SOIC (R) package, the AD844J is specified for the commercial temperature range of 0°C to 70°C. The AD844A and AD844B are specified for the industrial temperature range of –40°C to +85°C. The AD844 can be used in ways similar to a conventional op amp while providing performance advantages in wideband applications. However, there are important differences in the internal structure that need to be understood in order to optimize the performance of the AD844 op amp[33].

2.2 LITERATURE REVIEW ON MCFOA OSCILLATOR

The modified current-feedback operational amplifier (MCFOA) that is used in two voltage mode (VM) frequency filters. Both structures presented employ single active and five passive elements. Circuits enable realizing a low- (LP), band- (BP), and high-pass (HP) filter responses without changing the circuit topology. There is no requirement for additional component or rotation of component to realize band-stop (BS) and all-pass (AP) filter function. Moreover, the independent adjustment of the quality factor Q by varying the value of single passive element without affecting of the natural frequency ω_o is allowed. All the passive and active sensitivities are low. In the PSPICE simulations the BJT simulation model of the MCFOA and its possible realization using the integrated circuit are used to confirm the workability of the proposed circuits. The modified current-feedback operational amplifier (MCFOA) is a combination of the plus-type (CCII+) and minus-type (CCII−) of second-generation current conveyor. Relations between the individual terminals of the MCFOA can be described asfollows:

$$i_Z = i_X, i_Y = -i_W, v_X = v_Y, v_W = v_Z.$$

Using a minimum number of passive components, i.e., new grounded and floating inductance simulators, grounded capacitance multipliers, and frequency-dependent negative resistors (FDNRs) based on one/two modified current-feedback operational amplifiers (MCFOAs), are proposed.

The type of the simulators depends on the passive element selection used in the structure of the circuit without requiring critical active and passive component-matching conditions and/or cancellation constraints. In order to show the flexibility of the proposed MCFOA, a single-input three-output (SITO) voltage-mode (VM) filter, two three-input single-output (TISO) VM filters, and an SITO current-mode (CM) filter employing a single MCFOA are reported. The layout of the proposed MCFOA is also given. We present a modified current-feedback operational amplifier (MCFOA), which is more suitable for realizing oscillation and active filters. A CMOS implementation of the MCFOA and its terminal resistance calculations are given in Section II. As applications, two circuits for realizing the grounded inductor, capacitor multiplier, and frequency-dependent negative resistor (FDNR) using a single MCFOA are proposed in Section III[43]. Oscillators are constructed experimentally using commercially available active devices, AD844s [33], to confirm the workability of the filters. The MCFOA is realized using three AD844s as given in [43].

2.3 SUMMARY

In this chapter we see all the references [1, 43]. Going through every reference we see how much worked have been done in CFOA and MCFOA. There is one paper in which function generator is build sinusoidal, rectangular and triangular. In one reference paper of MCFOA there is only work of simulation of capacitor, inductor and of active filter. Doing complete study of all these references here we are going to simulate function generator using CFOA and MCFOA.

CHAPTER 3
CURRENT FEEDBACK OPERATIONAL AMPLIFIER

3.1 CFOA THEORY

The operational amplifier was originally designed to perform mathematical operations by using voltage as an analogue of another quantity. Op-amps were originally developed in the vacuum tube era. At that time they consumed lots of space and energy. Later operational amplifiers were made smaller by implementing discrete transistors. Now they are implemented as monolithic integrated circuits (ICs). These are highly efficient and cost effective. In the late 1960s, the first integrated op-amp became widely available, which was created by Bob Widlar. It was rapidly superseded by the 741. Op-amp circuits are the basis for the electronic analog computer. They do basic mathematical operations such as addition, subtraction, integration, differentiation, and so on. Today an operational amplifier is a versatile circuit element with many applications. Op-amps usually have parameters within tightly specified limits, with standardized packaging and power supply requirements. With only a few external components, op-amp circuits can perform a wide variety of analog signal processing tasks.

Analog Design has historically been dominated by voltage mode signal processing. Voltage Feedback amplifier architecture has several attractive features. These are the differential long-tail pair, high-impedance input stage, which is very good at rejecting common mode signals. However the voltage feedback amplifier has some limitations. The gain bandwidth product is a constant and slew rate is low compared to the current feedback operational amplifier. Slew rate is limited to a maximum value which is determined by the ratio of the input stage bias current to the dominant pole compensation capacitor.

Recently, the current feedback operational amplifier has become more popular and is widely being used in the electronics and telecommunications industries. It offers some very good features over the voltage feedback operation amplifier. The two great features in the CFOA are slew rate and bandwidth. Although the CFOA is better than the VFOA in slew rate and bandwidth, the CFOA has some disadvantages. These are the input offset voltage (V_{OS}), input offset current (I_{OS}), common mode input range (CMIR), common mode rejection ratio (CMRR), power supply rejection ratio (PSRR), and open loop gain. VFOAs are generally used for precision and general purpose applications. CFOAs are used in high frequency applications especially over 0 -100MHz

3.2 APPLICATION FOR CURRENT FEEDBACK OPERATIONAL AMPLIFIER

There are lots of applications for the CFOA circuit. Included in them are arbitrary waveform driver, high-resolution and high-sampling rate analog to digital converter (ADC) drivers, high-resolution and high-sampling rate digital to analog converter (DAC) output buffers, IF amplification for wireless communications applications, broadcast video and HDTV line drivers, PC and workstation video boards, facsimile and imaging systems, DSL line driver, medical imaging, high-resolution video, high speed signal processing, pulse amplifiers, ADC/DAC gain amplifiers, monitor preamplifiers, low cost precision IF amplifiers, and active filters.

3.3 ADVANTAGES AND DISADVANTAGES OF THE CURRENT FEEDBACK OPERATIONAL AMPLIFIER

The advantages of the current feedback operational amplifier over the voltage feedback operational amplifier are:

1. Slew rate (SR) – slew rate for CFOA is high compared to VFOA.

2. Bandwidth – bandwidth for CFOA is high compared to VFOA.

The disadvantages of current feedback operational amplifiers are:

1. Input offset voltage (VOS) - VOS for a CFOA is high compared to a VFOA.

2. Input offset current (IOS) - IOS for a CFOA is high compared to a VFOA.

3. Common mode input range (CMIR) - CMIR for a CFOA is low compared to a VFOA.

4. Input offset current (IOS) - IOS for a CFOA is high compared to a VFOA.

5. Common mode input range (CMIR) - CMIR for a CFOA is low compared to a VFOA.

6. Common mode rejection ratio (CMRR) - CMRR for a CFOA is low compared to a VFOA.

7. Power supply rejection ratio (PSRR) - PSRR for a CFOA is low compared to a VFOA.
8. Open loop gain - open loop gain for a CFOA is low compared to a VFOA.

3.4 BASICS OF CFOA

The current feedback operational amplifier otherwise known as CFOA or CFA is a type of electronic amplifier whose inverting input is sensitive to current, rather than to voltage as in a conventional voltage-feedback operational amplifier (VFA). The CFOA was invented by David Nelson at Comlinear Corporation.

The circuit symbol of the current feedback operational amplifier (CFOA) is shown in Figure 3.1. It is a four terminal device where Y and X are input, W and Z are output terminals.

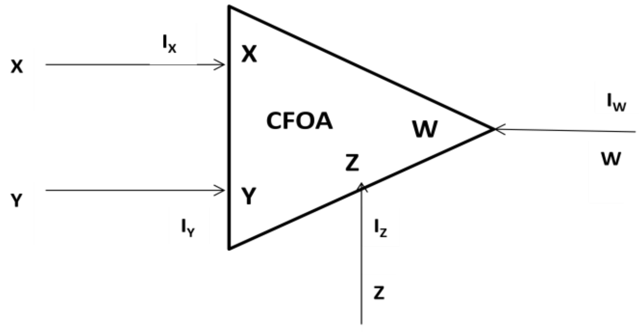


Figure 3.1: Symbol of CFOA

It is characterized with the following matrix equation

$$\begin{bmatrix} I_y \\ Iz \\ V_X \\ V_W \end{bmatrix} = \begin{bmatrix} 0 & 0 & 0 \\ 1 & 0 & 0 \\ 0 & 1 & 0 \\ 0 & 0 & 1 \end{bmatrix} \begin{bmatrix} I_X \\ V_Y \\ V_Z \end{bmatrix} \tag{3.1}$$

3.5 OPERATION OF CFOA

Referring to the schematic shown below, the first section forms the input stage and error amplifier. The inverting input (node where emitters of Q1 & Q2 are connected) is low impedance and hence, sensitive to changes in current. Resistors R1–R4 set up the quiescent bias conditions and is chosen such that the collector currents of Q1 & Q2 are the same. In most designs, active biasing circuitry is used instead of passive resistive biasing, and the non-inverting input may also be modified to become low impedance like the inverting input in order to minimise offsets.

With no signal applied, due to the current mirrors Q3/Q4 & Q5/Q6, the collector currents of Q4 and Q6 will be equal in magnitude if the collector 3ecurrents of Q1 and Q2 are also equal in magnitude. Thus, no current will flow into the buffer's input (or equivalently no voltage will be present at the buffer's input). In practice, due to device mismatches the collector currents are unequal and this results in the difference flowing into the buffer's input resulting in an offset at its output. This is corrected by adjusting the input bias or adding offset nulling circuitry. The second section (Q3–Q6) forms an I-to-V converter. Any change in the collector currents of Q1 and Q2 (as a result of a signal at the non-inverting input) appears as an equivalent change in the voltage at the junction of the collectors of Q4 and Q6. C_s is a stability capacitor to ensure that the circuit remains stable for all operating conditions. Due to the wide open-loop bandwidth of a CFA, there is a high risk of the circuit breaking into oscillations. C_s ensures that frequencies where oscillations might start are attenuated, especially when running with a low closed-loop gain.

The output stage is a buffer which provides current gain. It has a voltage gain of unity (+1 in the schematic).

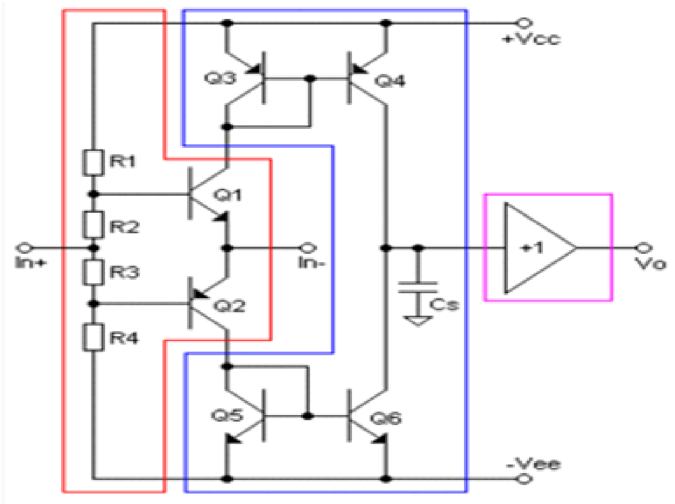


Fig. 3.2: Representative schematic of a current feedback operational amplifier

In simple configurations, such as linear amplifiers, a CFOA can be used in place of a VFA with no circuit modifications, but in other cases, such as integrators, a different circuit design is required. The classic four-resistor differential amplifier configuration also works with a CFOA, but the common mode rejection ratio is poorer than that from a VFA. Internally compensated, VFA bandwidth is dominated by an internal dominant pole compensation capacitor, resulting in a constant gain/bandwidth limitation. CFOAs, in contrast, have no dominant pole capacitor and therefore can operate much more closely to their maximum frequency at higher gain. Stated another way, the gain/bandwidth dependence of VFA has been broken.

In VFAs, dynamic performance is limited by the gain-bandwidth product and the slew rate. CFOA use a circuit topology that emphasizes current-mode operation, which is inherently much faster than voltage-mode operation because it is less prone to the effect of stray node capacitances. When fabricated using high-speed complementary bipolar processes, CFOAs can be orders of magnitude faster than VFAs. With CFOAs, the amplifier gain may be controlled independently of bandwidth. This constitutes the major advantages of CFOAs over conventional VFA topologies.

Disadvantages of CFOAs include poorer input offset voltage and input bias current characteristics. Additionally, the DC loop gains are generally smaller by about three decimal orders of magnitude. Given their substantially greater bandwidths, they also tend to be noisier.

3.6 SUMMARY

In this chapter we study CFOA basic theory, its operation, its terminal equation and its advantages and disadvantages. The application advantages of current feedback and voltage feedback differ. In many applications, the differences between CFB and VFB are not readily apparent. Today’s CFB and VFB amplifiers have comparable performance, but there are certain unique advantages associated with each topology. Voltage feedback allows freedom of choice of the feedback resistor (or network) at the expense of sacrificing bandwidth for gain. Current feedback maintains high bandwidth over a wide range of gains at the cost of limiting the choices in the feedback impedance.

CHAPTER 4
CFOA BASED OSCILLATOR

4.1 SINUSOIDAL OSCILLATOR USING CFOA

The sinusoidal oscillator circuit using CFOA is shown in Figure 4.1. A number of sinusoidal oscillators using CFOA have been proposed. There are two ways to find the frequency of oscillation of this circuit. The first way is to find the loop gain. To find the loop gain, the loop is broken at the CFOA positive input terminal and an input voltage V1 is applied. Then the return voltage that appears at node 1 is found. The loop gain is equated to unity. The second approach is to analyze the circuit and eliminate all current and voltage variables. One equation is obtained that represents circuit operation. Oscillation will start if this equation is satisfied. From the resulting equation the frequency of oscillation and condition of oscillation can be found. The frequency of oscillation for the sinusoidal is

$$f = \frac{1}{\sqrt{2}\,\pi\,\sqrt{(R2\;R3\;C1\;C5)}} \tag{4.1}$$

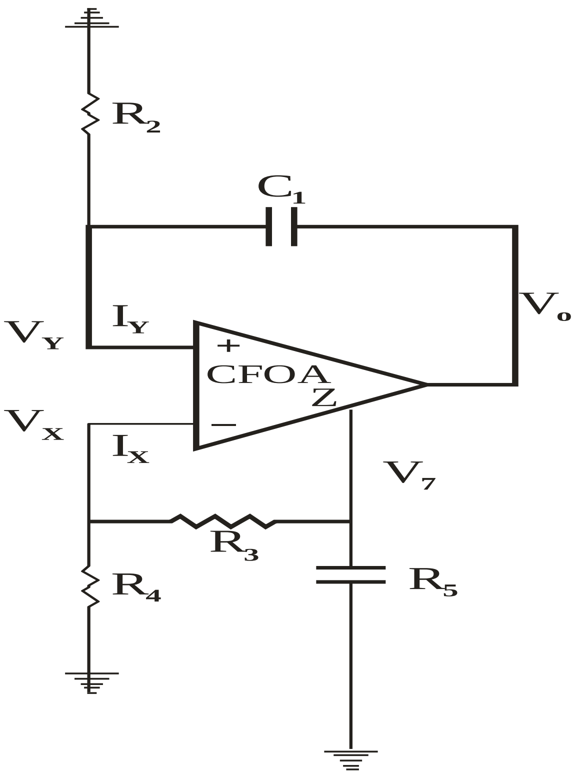


Figure 4.1 sinusoidal oscillator using CFOA.

And,

Condition of oscillation is

$$\frac{C_5}{C_1} = \frac{R_2}{R_4} \tag{4.2}$$

Equations of CFOA are

$$I_x = I_z \tag{4.3}$$

$$V_z = V_o \tag{4.4}$$

$$V_x = V_y \tag{4.5}$$

$$I_z = \frac{V_x - V_o}{R_3} + \frac{0 - V_o}{\frac{1}{sC_5}} \tag{4.6}$$

$$I_z = \frac{V_x}{R_3} - V_o\left(\frac{1}{R_3} + sC_5\right) \tag{4.7}$$

$$I_x = \frac{0 - V_x}{R_4} + \frac{V_o - V_x}{R_3} \tag{4.8}$$

From equation 4.3

$I_x = I_z$

$$\frac{0 - V_x}{R_4} + \frac{V_o - V_x}{R_3} = \frac{V_x}{R_3} - V_o\left(\frac{1}{R_3} + sC_5\right) \tag{4.9}$$

$$V_x\left(\frac{2}{R_3} + \frac{1}{R_4}\right) = V_o\left(\frac{2}{R_3} + sC_5\right) \tag{4.10}$$

From equation 4.5

$V_x = V_y$

$$V_x = V_o\left(\frac{sR_2C_1}{1 + sC_1R_2}\right. \tag{4.11}$$

From equation 4.10 and 4.11

$$V_o\left(\frac{sR_2C_1}{1 + sC_1R_2}\right)\left(\frac{2}{R_3} + \frac{1}{R_4}\right) = V_o\left(\frac{2}{R_3} + sC_5\right) \tag{4.12}$$

$$V_0\left\{\frac{sR_2C_1}{1 + sC_1R_2}\left(\frac{2R_4 + R_3}{R_3R_4}\right) - \left(\frac{2 + R_3C_5}{R_3}\right)\right\} = 0 \tag{4.13}$$

$$V_o\left\{\frac{2sR_2R_4C_1 + sR_2R_3C_1 - (R_4 + s\,C_1R_2R_4)(2 + sR_3C_5)}{R_3R_4(1 + sC_1R_2)}\right\} = 0 \tag{4.14}$$

After solving further

$$s^2 C_1 C_5 R_2 R_3 R_4 + (R_4 R_3 C_5 - R_2 R_3 C_1)s + 2R_4 = 0 \quad (4.15)$$

which is an equation of oscillator

$$as^2 + (\alpha - \beta)s + \gamma = 0 \quad (4.16)$$

Putting s = jω in equation 4.15

$$(j\omega)^2 C_1 C_5 R_2 R_3 R_4 + (R_4 R_3 C_5 - R_2 R_3 C_1)(j\omega) + 2R_4 = 0 \quad (4.17)$$

$$-\omega^2 C_1 C_5 R_2 R_3 R_4 + j\omega(R_4 R_3 C_5 - R_2 R_3 C_1) + 2R_4 = 0 \quad (4.18)$$

Equating real part and imaginary part

Real part = 0

$$2R_4 - \omega^2 C_1 C_5 R_2 R_3 R_4 = 0 \quad (4.19)$$

$$\omega^2 = \frac{2}{C_1 C_5 R_2 R_3}$$

$$f = \frac{1}{\sqrt{2}\,\pi\sqrt{(R2\ R3\ C1\ C5)}} \quad (4.20)$$

Imaginary part =0

$$R_4 R_3 C_5 = R_2 R_3 C_1 \quad (4.21)$$

$$\frac{C_5}{C_1} = \frac{R_2}{R_4} \quad (4.22)$$

This is a condition of oscillation.

4.2 SIMULATION RESULT OF OSCILLATION

When we select the component value of resistor and capacitor which satisfy condition of oscillation, on tuning of resistor and capacitor oscillation comes in orcad PSPICE simulator in the form of sinusoidal wave. Practically $\frac{R_2}{R_4}$ is smaller than 1 and $\frac{C_5}{C_1}$ is also smaller than 1 which is a difference between theoretical and practical oscillation.

Values of components are: C_1=10pf, C_5=1pf, R_2=5kΩ, R_3=2kΩ, R_4=6kΩ

So by using the frequency formula,

$$f = \frac{1}{\sqrt{2}\,\pi\sqrt{(R2\ R3\ C1\ C5)}}$$

On putting the value

$$f = \frac{1}{\sqrt{2}\,\pi\sqrt{(5 \times 10^3 \times 2 \times 10^3 \times 1 \times 10^{-12} \times 10 \times 10^{-12})}}$$

$$f = \frac{1}{\sqrt{2}\,\pi \times 10^{-8}}$$

f = 22.51 MHz

Practically frequency is coming 10 MHz.

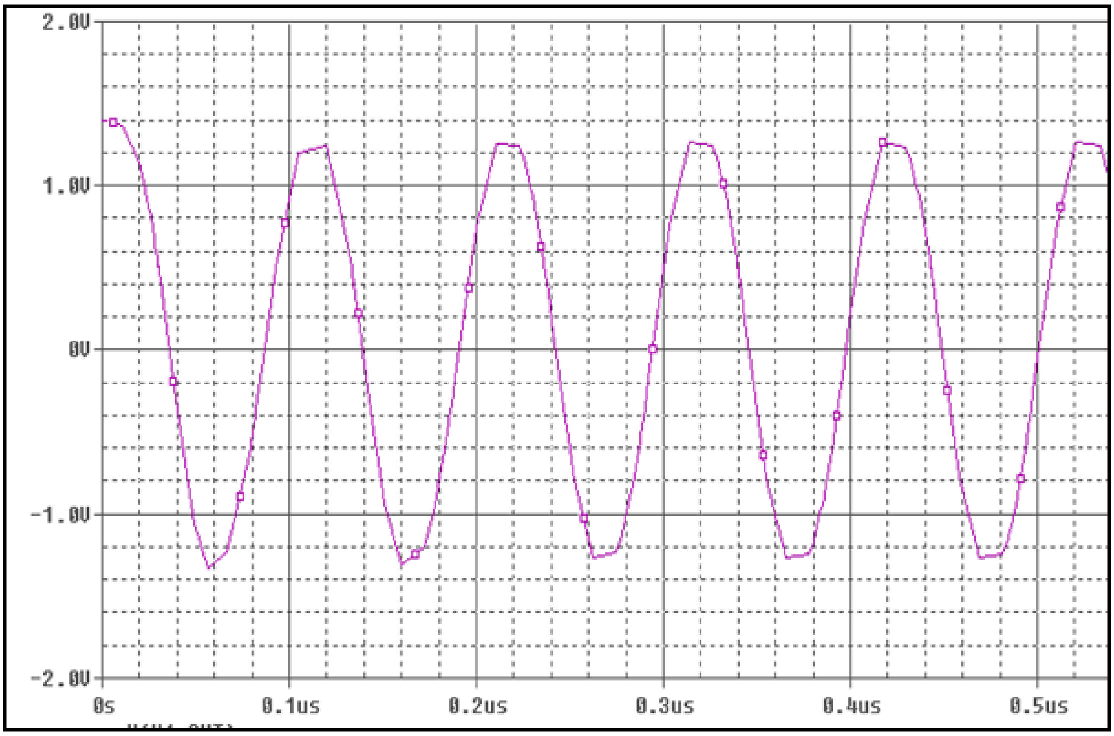


Figure 4.2 sine wave of CFOA oscillator.

4.3 INTEGRATOR USING CFOA

An integrator using a CFOA and its bode plot are shown in Fig. 4.3 and 4.4 respectively. The transfer function for the integrator can be found as

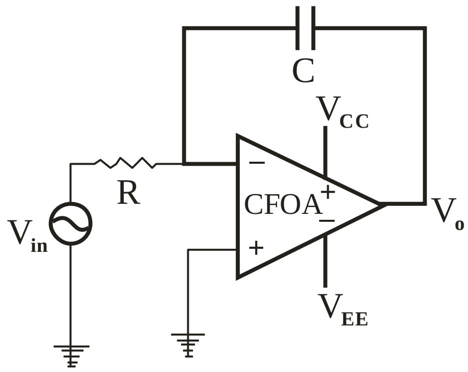


Figure 4.3: CFOA integrator.

$$\frac{V_{in} - V_-}{R} = \frac{V_- - V_o}{R} \quad (4.23)$$

$$V_- = V_+ \quad \text{and} \quad V_+ = 0 \quad (4.24)$$

$$\frac{V_o}{V_{in}} = \frac{1}{sRC} \quad (4.25)$$

$$\frac{V_o}{V_{in}} = \frac{1}{s\tau} \quad (4.26)$$

Where τ = RC, time constant

Division by s in the frequency domain corresponds to integration in the time domain. Due to the negative sign in the transfer function, this integrator is an inverting integrator.

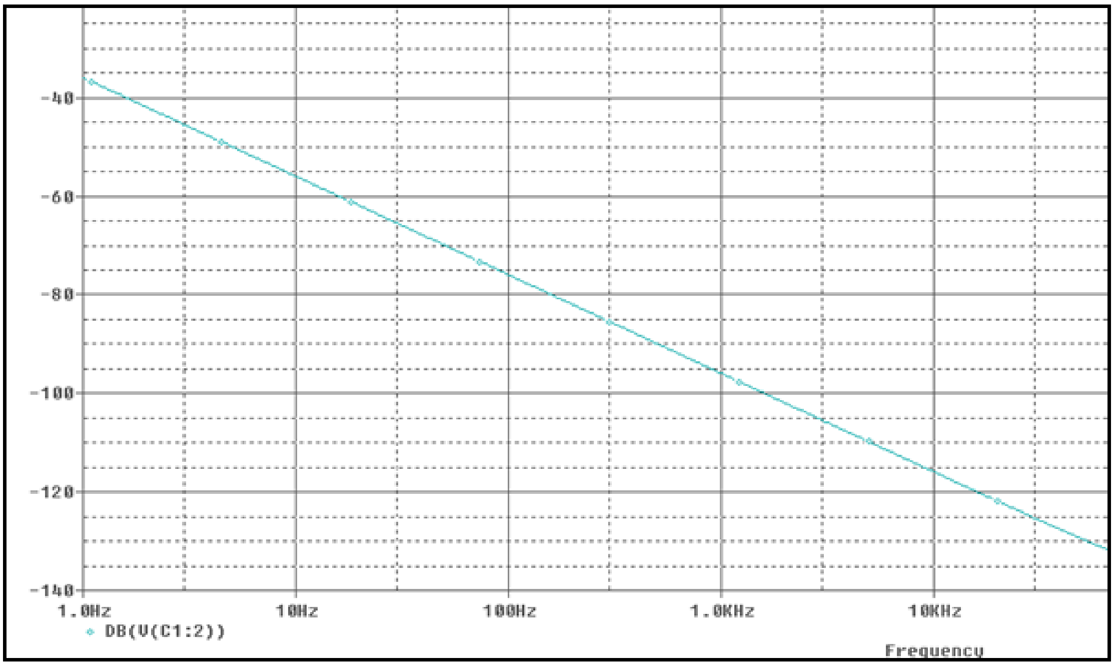


Figure 4.4 Bode plot of CFOA integrator.

4.4 TRIANGULAR AND RECTANGULAR WAVE GENERATOR USING CFOA

4.4.1 Multivibrators

Multivibrators are often used to generate triangular, square, or pulse waves. There are three types of Multivibrators: bistable, monostable and astable. A bistable multivibrator has two stable states. Circuits can stay in either stable state indefinitely. External commands are necessary to change the circuit from one state to the other state. An astable multivibrator toggles periodically from one state to another state without any external command. The timing is set by a suitable circuit usually a capacitor or quartz crystal. A monostable multivibrator has only one stable state. A trigger signal is necessary to force output to the other unstable state. After the circuit stays in the unstable state for a fixed time, the circuit returns to the stable state. A delay is set by a suitable circuit.

4.4.2 Bistable Multivibrators

One of the simplest bistable multivibrators is the comparator. A bistable multivibrator can be made by connecting an amplifier in the positive feedback configuration. In this case the loop gain is greater than unity. The circuit with the positive feedback configuration is shown in Figure 4.5

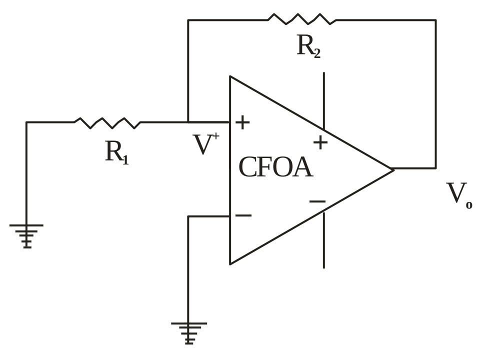


Figure 4.5 A positive feedback loop capable of bistable operation.

The operation of this circuit is explained below. This circuit has no external excitation. Assume a small electrical noise causes the positive input terminal to increase V^+ voltage. This signal is amplified by the amplifier. Let the open loop gain of the amplifier be A, which is a very large number. Thus the output voltage V_o, would be much greater than V^+. The voltage divider will feed V_o R1/(R1+R2) to the positive input terminal of the amplifier. This voltage is amplified by the amplifier again. This regenerative process continues until the op amp saturates and the output voltage reaches the positive rail. Let the positive rail voltage be V_{sat}^{+}. At this time the voltage at the positive input terminal is V_{sat}^{+} R1/(R1+R2). This is positive and keeps the CFOA at positive saturation. This is one of the stable states for this circuit. A negative rail output voltage can be explained by similar ways. Let the negative rail voltage be V_{sat}^{-}. This is the other stable state. This circuit has two stable states. This circuit can exist at one of the two states indefinitely, but cannot stay at the $V_o = 0$ state. Any disturbance, such as noise can cause the circuit to switch to any of the two stable states. The $V_o = 0$ state is called the metastable state.

4.4.3 Transfer Characteristics of a Bistable Circuit

For this circuit, there are two terminals connected to the ground. Either of the two terminals can serve as an input terminal. Figure 4.6 shows a bistable circuit where the input voltage, V_{in} is connected to the negative input terminal. To derive the V_o to V_{in} transfer characteristic, let V_0 be at one of the two possible levels, say V_{sat}^{+} .

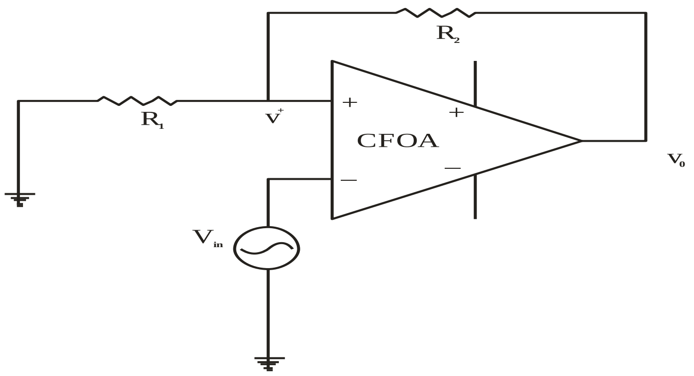


Figure 4.6 A bistable circuit

A bistable circuit where input voltage V_{in} is connected in negative input terminal. Thus the voltage at the positive input terminal of the CFOA is

$$V^{+} = \frac{R_1}{R_1 + R_2} V_{Sat}^{+} \tag{4.27}$$

Let, V_{in} increase from zero voltage. V_o will stay constant at V_{Sat}^{+} until $V_{in} = v^{+}$. As V_{in} begins to increase v^{+}, a net negative value develops between the two input terminals of the op amp. This voltage is amplified by the amplifier. Thus V_o goes negative. The voltage divider will feed V_oR1/(R1+R2) voltage to the positive input terminal of the Op- Amp. Here V_o is negative. So, v^{+} goes to negative. This increases the net negative voltage at the input terminal of the CFOA. The net negative input amplified by the CFOA again and this process will continue until the V_o reaches the negative rail. If this voltage is V_{Sat}^{-}, then any further increase of V_{in} has no effect on V_o. Figure 4.7 (a) shows the input-output characteristics for increasing Vin.

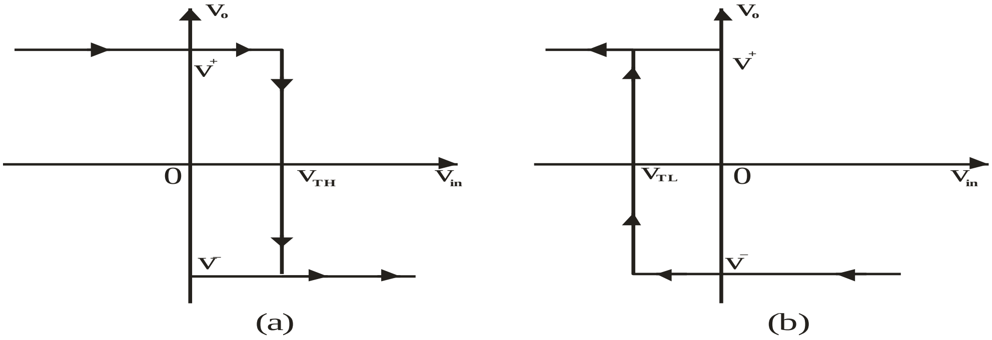


Figure 4.7 (a) Figure 4.7 (b)

(a) Vin – V0 characteristics for increasing Vin, (b) Vin – V0 characteristics for decreasing Vin

$$V^{+} = \frac{R_1}{R_1 + R_2} V_{Sat}^{-} \tag{4.28}$$

The output for the op amp will stay at negative saturation until Vin is equal to V_{Sat}^{-}. R1/(R1+R2). If V_{in} goes below this value, a net positive voltage appears between the op amp input terminals. This voltage is amplified by the amplifier. Thus V_o goes positive. The voltage divider will feed back V_oR1/(R1+R2) to the positive input terminal of the Op- Amp. Here V_o is positive. So, v^{+} goes to a positive value. This increases the net positive voltage at the input terminal of the op amp. The input terminal voltage will be amplified by the CFOA again, and this process will continue until V0 reaches the positive rail. Any further increase of V_{in} has no effect onV_o. Figure 4.7 (b) shows the input output characteristics for decreasing V_{in}. The complete $V_{in} - V_o$ characteristic for the circuit is shown in Figure 4.8.

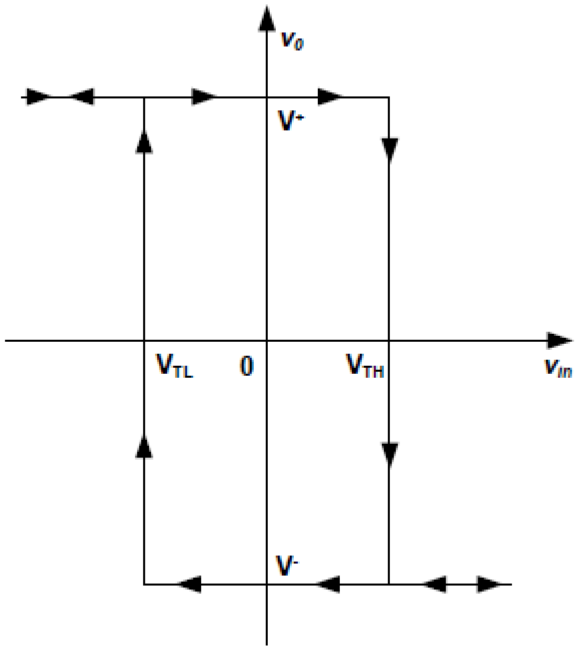
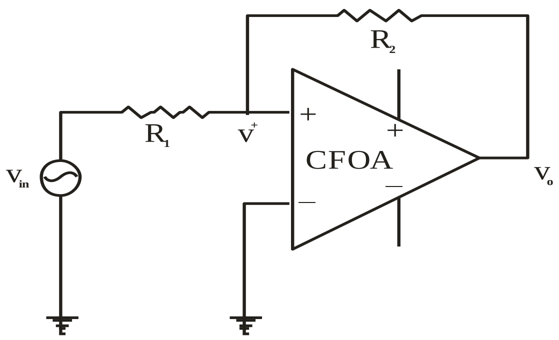


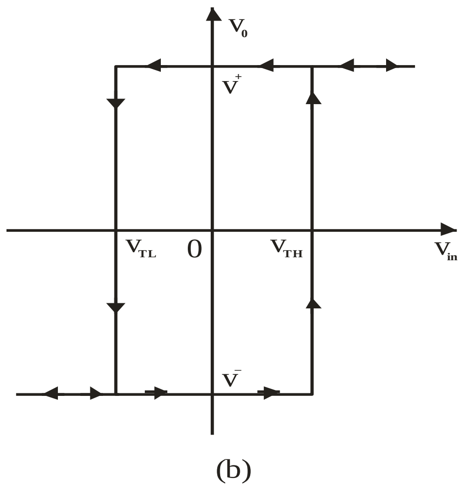
Figure 4.8 Complete $V_{in} – V_0$ characteristic

The circuit changes states at different values of V_{in} for increasing or decreasing V_{in}. It is said that the circuit has hysteresis. The width of the hysteresis is the difference between the high threshold (V_{TH}) and the low threshold (V_{LH}). The bistable circuit changes from the positive state to the negative state for increasing V_{in}. The circuit is an inverting bistable circuit. This bistable circuit is also known as a Schmitt trigger.

A non-inverting bistable circuit and its transfer characteristics are shown in Figure 4.9. The circuit operation can be explained in similar ways as above.



(a)



(b)

Figure 4.9 (a) Non-inverting bistable circuit

Figure 4.9 (b) $V_{in} – V_0$ characteristic for Non-inverting bistable circuit.

Some applications for bistable circuits are comparator, digital memory system, and square wave generator.

4.4.4 Triangular and Rectangular Wave Generator Using CFOA

The triangular and rectangular wave generator circuit is shown in Fig. 4.10. The triangular wave can be generated by alternately charging and discharging a capacitor with a constant current. For this circuit, current drive for the capacitor C_1 is provided by the $CFOA_1$. In the second circuit, $CFOA_2$, is configured as a Schmitt trigger. Here, the diode clamp is used to stabilize the Schmitt trigger output level at $\pm\ nV_{D(ON)}$, where n is the number of diodes and $V_{D(ON)}$ is the forward voltage drop for each diode. Without the diode clamp the Schmitt trigger output would be at the ± of the maximum output voltage of the CFOA.

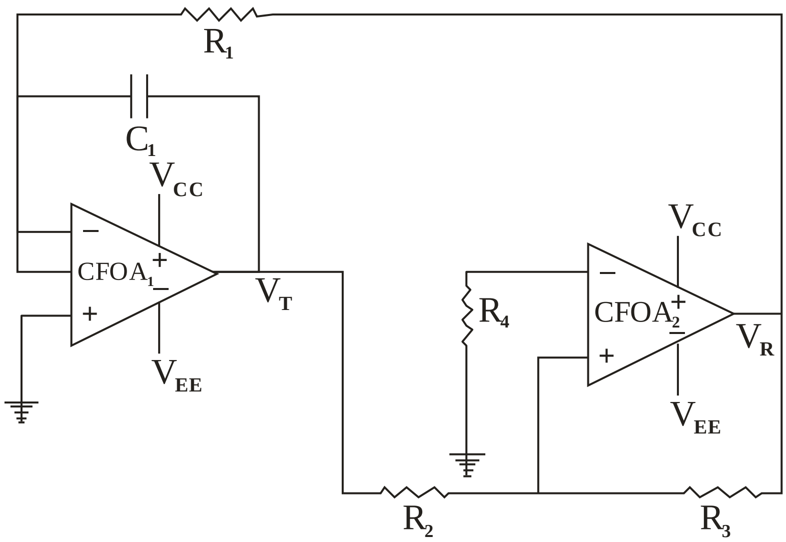


Figure 4.10 Triangular and rectangular wave generator.

The frequency of oscillation of this circuit can be found by the following process. The integrator with capacitor in the feedback path and resistor in the inverting input terminal with branch currents are shown in the Figure 4.10. The upward ramping voltage for the integrator is shown in Figure 4.11. The time required to charge the capacitor by a voltage ΔV is Δt.

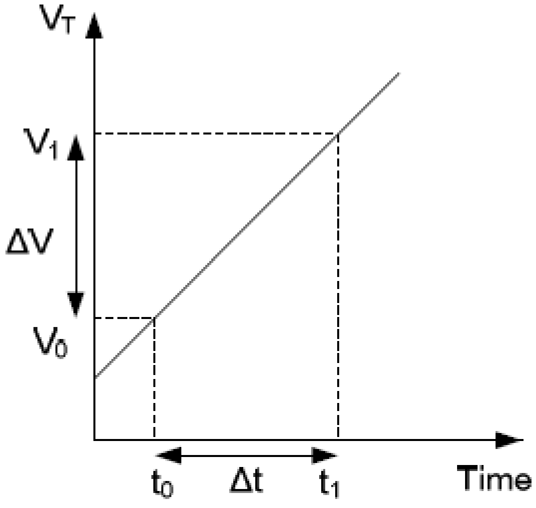


Figure 4.11 Upward ramping voltage for the integrator.

So

$$C_1 \Delta V = I \Delta t \tag{4.29}$$

Rearranging equation (4.29) for Δt,

$$\Delta t = \frac{C_1}{I} \Delta V \tag{4.30}$$

The currents at node 1 of Figure 4.11 is,

$$I = I_1 + I_n \tag{4.31}$$

Using the macro-model for the CFOA in the integrator,

$$I_n = - \frac{V_T}{Z_T} \tag{4.32}$$

Here Z_T is the open loop transimpedance for the CFOA.

$$I = \frac{V_R}{R_1} - \frac{V_T}{Z_T} \tag{4.33}$$

The Schmitt Trigger circuit is shown if Figure 4.12.

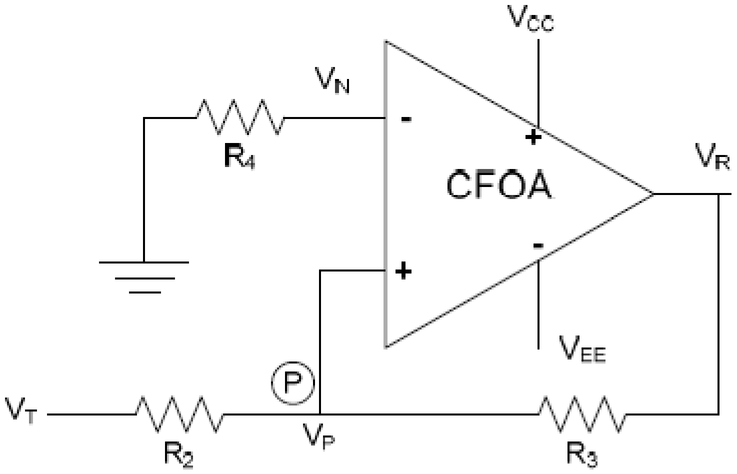


Figure 4.12: Schmitt trigger using CFOA

Using the node equation at node P,

$$\frac{V_T - V_P}{R_2} = \frac{V_P - V_R}{R_3} \tag{4.34}$$

Rearranging equation (4.34) for V_T

$$V_T = V_P\left(1 + \frac{R_2}{R_3}\right) - V_R\frac{R_2}{R_3} \tag{4.35}$$

A sample triangular waveform is shown in Figure 4.13

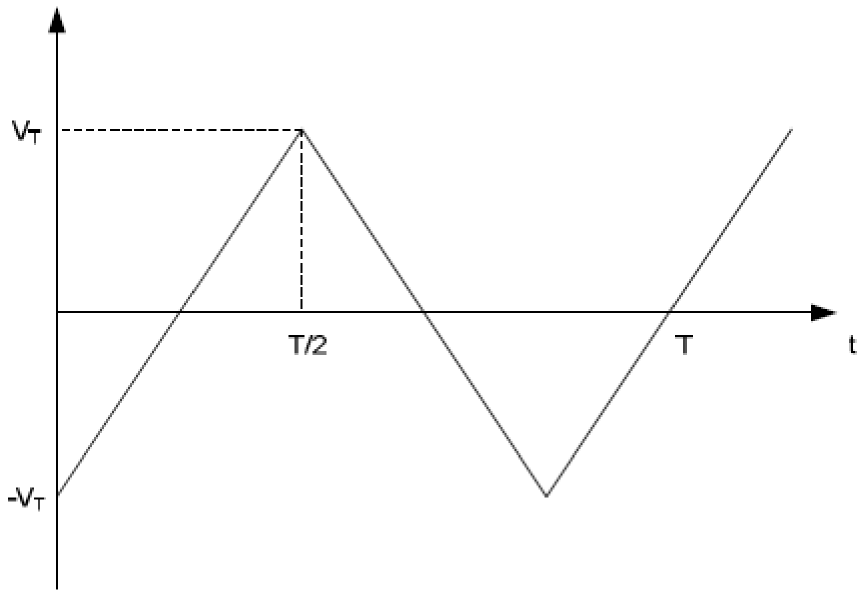


Figure 4.13 Triangular waveform

The time taken to ramp from $-V_T$ to V_T is $T/2$. So

$$\Delta t = \frac{T}{2} \tag{4.36}$$

Here ΔV is the voltage swing for the triangular waveform. Thus

$$\Delta V = 2V_T \tag{4.37}$$

From equation (4.30) and (4.36),

$$\frac{T}{2} = \frac{C_1}{I}\Delta V \tag{4.38}$$

$$\frac{T}{2} = \frac{C_1}{I}2V_T \tag{4.39}$$

$$T = 4\frac{C_1}{I}V_T \tag{4.40}$$

$$T = \frac{4C_1\left[V_P\left(1+\frac{R_2}{R_3}\right) - V_R\frac{R_2}{R_3}\right]}{\frac{V_R}{R_1} - \frac{V_T}{Z_T}} \tag{4.41}$$

The period of the triangular, V_T or rectangular, V_R waveform is.

$$T = \frac{4C_1\left[V_P\left(1+\frac{R_2}{R_3}\right) - V_R\frac{R_2}{R_3}\right]}{\frac{V_R}{R_1} - \frac{V_T}{Z_T}} \tag{4.42}$$

The frequency of oscillation for the triangular and rectangular waveform is

$$f_o = \frac{\frac{V_R}{R_1} - \frac{V_T}{Z_T}}{4C_1\left[V_P\left(1+\frac{R_2}{R_3}\right) - V_R\frac{R_2}{R_3}\right]} \tag{4.43}$$

For the Schmitt trigger,

$$V_P = V_N \tag{4.44}$$

Thus the frequency of oscillation for the triangular and rectangular waveform becomes

$$f_o = \frac{\frac{V_R}{R_1} - \frac{V_T}{Z_T}}{4C_1\left[V_N\left(1+\frac{R_2}{R_3}\right) - V_R\frac{R_2}{R_3}\right]} \tag{4.45}$$

4.4.5 Simulation Results

The frequency of oscillation for the triangular and rectangular waveform is given in equation (4.45).

Here

V_R = Peak voltage of the square waveform

V_N = Peak voltage at the negative input terminal of the $CFOA_2$

Z_T = Open loop trans-impedance of the $CFOA_1$

But here in simulation we are using macro-model of analog device in which V_T/Z_T is very much less than 1, So the open loop trans-impedance of the $CFOA_1$ we are neglecting here. To verify equation (4.45), is used for calculation.

Figure 4.14 shows the calculated and simulated rectangular and triangular wave of circuit shown in figure 4.10, R1 is 950Ω . Here R2 is 2 kΩ, R3 is 20kΩ, R4 is 1kΩ and C1 is 0.8 nF. The values of R2, R3, and C1 are chosen so that the output voltage is stable. The frequency of oscillation is 1 MHz.

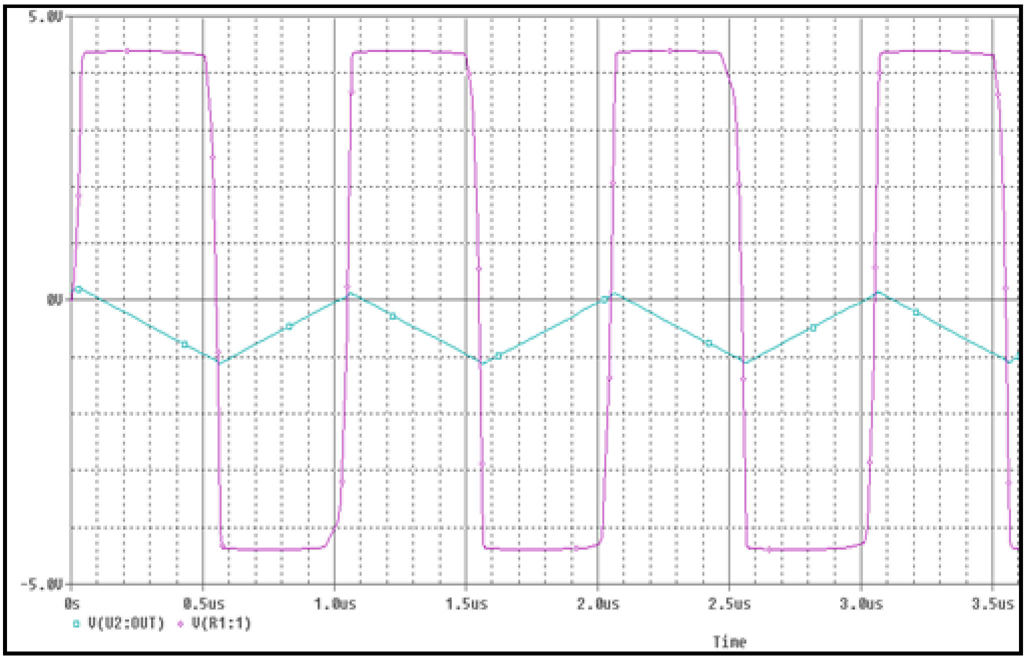


Figure 4.14 Rectangular and triangular wave using CFOA.

4.5 SUMMARY

Analysis of CFOA is done by using AD844 8- pin IC of analog device. Here we build sine wave of 7MHz, integrator, rectangular wave, and triangular wave of 1MHz. simulation is done on orcad pspice 16.5 version. We can can achieve higher oscillation also By using modified CFOA. This can be useful for faster device. So CFA are much better than VFA. This is proven in this chapter and above simulation.

CHAPTER- 5
MODIFIDED CURRENT FEEDBACK OP-AMP

5.1 INTRODUCTION

The circuit symbol of the current feedback operational amplifier (CFOA) is shown in Figure 5.1. It is a four terminal device where Y and X are input, W and Z are output terminals.

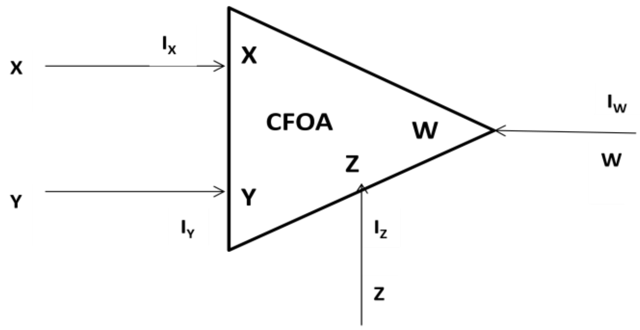


Fig. 5.1: Symbol of CFOA

It is characterized with the following matrix equation

$$\begin{bmatrix} I_Z \\ I_Y \\ V_X \\ V_W \end{bmatrix} = \begin{bmatrix} 0 & 0 & 0 \\ 1 & 0 & 0 \\ 0 & 1 & 0 \\ 0 & 0 & 1 \end{bmatrix} \begin{bmatrix} I_X \\ V_Y \\ V_Z \end{bmatrix} \tag{5.1}$$

Using three CFOA we made a modified current feedback operation amplifier (MCFOA) which is providing higher oscillation than CFOA. We use here figure. 5.3 MCFOA for our simulation result.

Modified CFOA is a four terminal device characterized by the matrix equation:

$$\begin{bmatrix} I_Z \\ I_Y \\ V_X \\ V_W \end{bmatrix} = \begin{bmatrix} \alpha_1 & 0 & 0 & 0 \\ 0 & -\alpha_2 & 0 & 0 \\ 0 & 0 & \beta_1 & 0 \\ 0 & 0 & 0 & \beta_2 \end{bmatrix} \begin{bmatrix} I_X \\ I_W \\ V_Y \\ V_Z \end{bmatrix} \tag{5.2}$$

As it can be seen from above equation, the MCFOA is different from the conventional current-feedback operational amplifier (CFOA) because the W terminal current of the MCFOA is copied to the Y terminal in the opposite direction. However, it is well known that the Y-terminal current of the conventional CFOA is equal to zero. Fig. 5.2 and 5.3 show the symbol and construction using commercially active devices of modified CFOA (MCFOA) respectively.

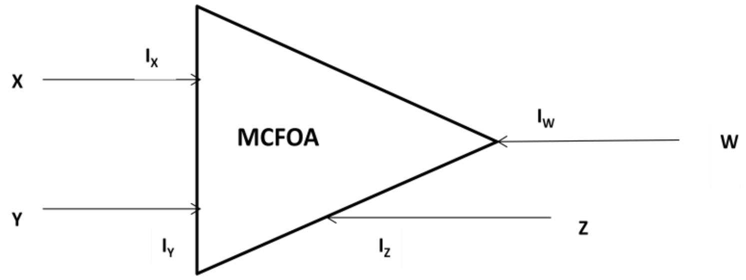


Fig. 5.2: Symbolic representation of the MCFOA.

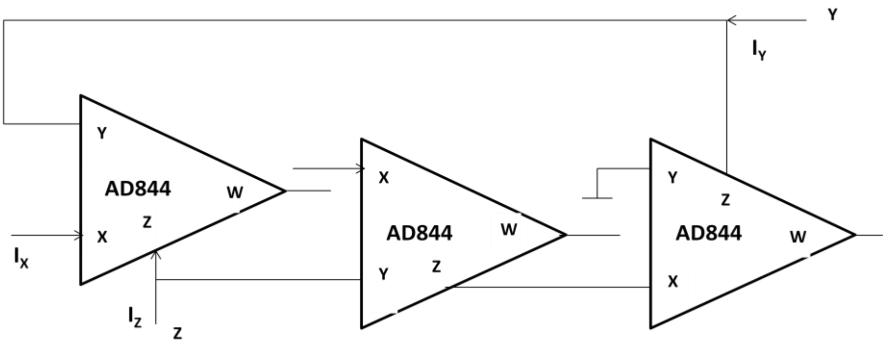


Fig. 5.3: MCFOA construction using commercially available active devices.

5.2 **CMOS REALIZATION OF MODIFIED CFOA**

CMOS realization of modified CFOA proposed is shown in Fig. 5.4. The output resistances of the transistors M9, M10, M19, and M20 in the MCFOA of Fig. 5.4 are assumed to be equal to r0 and, similarly, the output resistances of the transistors M11, M12, M19, and M20 of the MCFOA are equal to r'0. Thus, the resistances seen at terminals Y, Z, X, and W are respectively calculated as

$$R_Y = \frac{r_{03} r_{024}}{r_{08} + r_{024}} \tag{5.3}$$

$$R_Z = \frac{r_{03} r_{018}}{r_{03} + r_{018}} \tag{5.4}$$

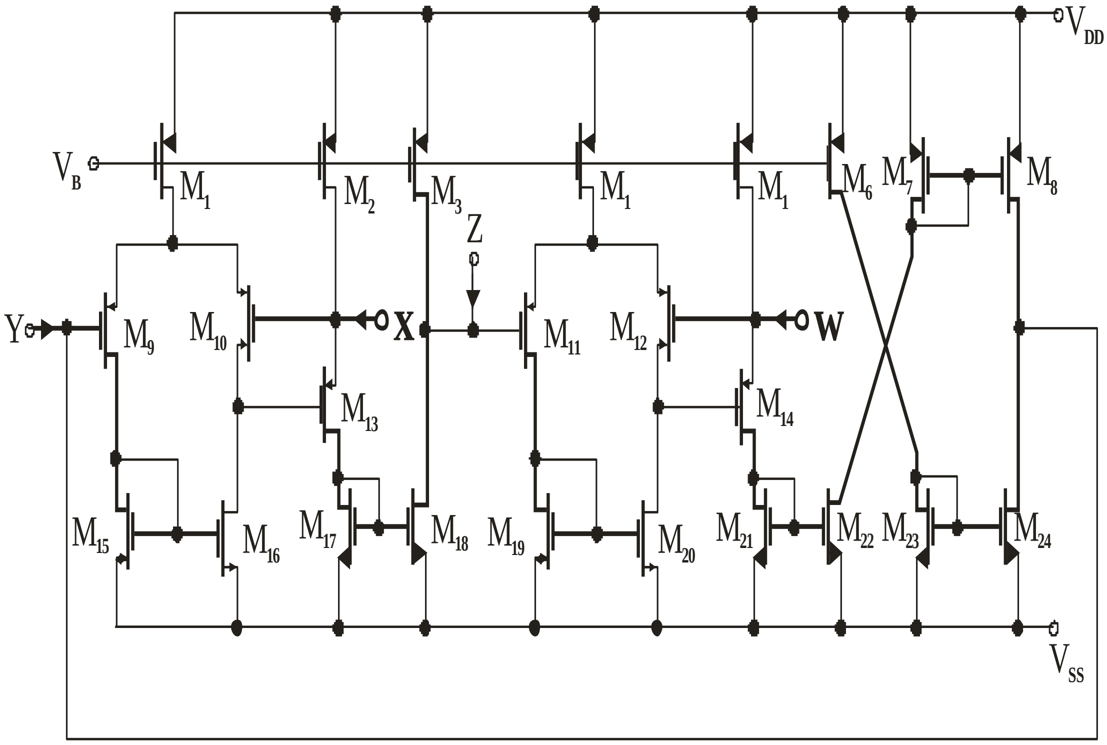


Fig. 5.4: Developed CMOS structure for the MCFOA

$$R_X = \left(\frac{r_{013} + \frac{r_{017}}{1 + g_{m17} r_{017}}}{1 + r_{013} g_{m13}\left(1 + \frac{r_0}{2} g_{m10}\right)}\right) || r_{02} \tag{5.5}$$

$$R_X \cong \frac{2}{g_{m13} g_{m10} r_0}$$

$$R_W = \left(\frac{r_{014} + \frac{r_{021}}{1 + g_{m21} r_{021}}}{1 + r_{014} g_{m14}\left(1 + \frac{r'_0}{2} g_{m12}\right)}\right) || r_{05} \tag{5.6}$$

$$R_W \cong \frac{2}{g_{m12} g_{m14} r'_0}$$

Here, r_{0i} and g_{mi} are the output resistance and transconductance of the i^{th} CMOS transistor, respectively. In (5.3) and (5.6), it is assumed that

$g_{m17}r_{017} >> 1$ and $g_{m21}r_{021} >> 1$, respectively

From (5.3)–(5.6), it can be seen that, while the terminals Y and Z have very high resistances, the terminals X and W exhibit low resistances due to the feedback loops composed of the transistors (M10 , M13) and (M12 , M14), respectively.

The proposed MCFOA can also be compared with a recently reported gain-variable third-generation current conveyor (GVCCIII) and will exhibit its superiority as follows. The GVCCIII is defined by the following matrix equation:

$$\begin{bmatrix} V_X \\ I_Y \\ I_Z \end{bmatrix} = \begin{bmatrix} \beta & 0 & 0 \\ 0 & -\Upsilon & 0 \\ 0 & \pm\alpha & 0 \end{bmatrix} \begin{bmatrix} V_Y \\ I_X \\ V_Z \end{bmatrix} \tag{5.7}$$

The + and - signs of above equation are used for the plus-type GVCCIII (GVCCIII+) and minus-type GVCCIII (GVCCIII-), respectively. It is observed from equation that the GVCCIII is a three terminal active device while the MCFOA is a four-terminal device. In other words, both the GVCCIII and MCFOA have terminals X, Y, and Z, but the MCFOA also has the terminal W. Also, the current gains and in a GVCCIII are constant and ideally equal to unity, but the voltage gain can be controlled by selecting appropriate external resistors. Contrary to variable voltage gain of the GVCCIII, both of the voltage gains and of the MCFOA are constant and ideally equal to unity. On the other hand, using a single GVCCIII or MCFOA as well as one capacitor and two resistors, a grounded inductor realization is possible. Nevertheless, inductor simulator with GVCCIII employs floating passive components and needs a resistive element matching condition.

5.3 CMOS IMPLEMENTATION OF MODIFIED CFOA

For simulation CMOS implementation of modified CFOA proposed is used. The SPICE simulation is performed using 0.25µm, Level 3, TSMC CMOS process parameters provided by MOSIS (AGILENT) and supply voltages taken are $V_{DD} = -V_{SS} = 1.25V$ and Biasing voltage $V_B = 0.8V$. Transistors aspect ratios are reported in Table 3.1.

Table 3.1: Aspect ratio of the CMOS transistor used in the MCFOA of Figure 5.4

PMOS Transistors	**$W_{(\mu m)}/L_{(\mu m)}$**
M_1, M_4 and M_9, M_{10}, M_{11} and M_{12}	1.0/0.25
M_2, M_3, M_5, M_6, M_7 and M_8	2.0/0.25
M_{13} and M_{14}	4.0/0.25
NMOS Transistors	**$W_{(\mu m)}/L_{(\mu m)}$**
M_{15}, M_{16}, M_{17}, M_{18}, M_{19}, M_{20}, M_{21}, M_{22}, M_{23} and M_{24}	0.5/0.25

The commercial current feedback amplifiers AD844 macro model with ± 5 V voltage supply is used to realize the CFOA in figure 3.1.

5.4 SUMMARY

In this chapter we study MCFOA basic theory, its operation, its terminal equation and its advantages and it brings higher oscillation than CFOA. Today's MCFOA and CFOA have comparable performance, but there are certain unique advantages associated with oscillation. Here two kinds of MCFOA one which is builds by three CFOA (AD 844 IC) and another is made by CMOS. The MCFOA builds by CMOS for that terminal resistance and terminal equation, internal circuit and ratio of width/ length is also given.

CHAPTER- 6
MCFOA BASED OSCILLATOR

6.1 SINE WAVE OSCILLATOR USING MCFOA

MCFOA oscillator is build when condition of oscillation is satisfied and can provide higher oscillation which can be used for high frequency oscillation after doing mathematical calculation, we get the desired condition of oscillation and frequency is

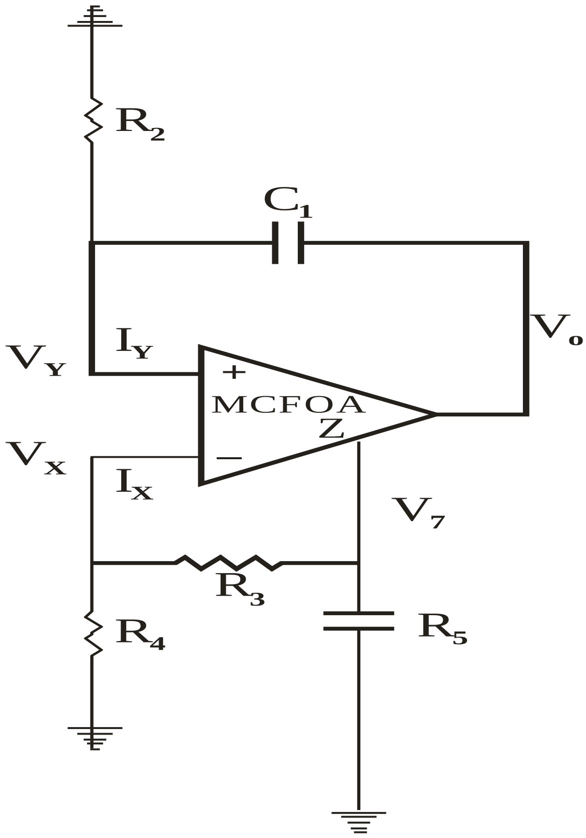


Figure 6.1 Sine wave oscillator using MCFOA.

$$f = \frac{1}{\sqrt{2}\ \pi \sqrt{(R2\ R3\ C1\ C5)}} \tag{6.1}$$

And,

Condition of oscillation is

$$\frac{C_5}{C_1} = \frac{R_2}{R_4} \tag{6.2}$$

6.2 CALCULATION FOR CONDITION OF OSCILLATION AND FREQUENCY

Equations 6.1 and 6.2 of MCFOA are practical and for mathematical calculation we are using ideal equation which is given below.

$$I_x = I_z \tag{6.3}$$

$$I_Y = - I_W \tag{6.4}$$

$$V_Z = V_W \tag{6.5}$$

$$V_x = V_y \tag{6.6}$$

$$I_z = \frac{V_x - V_o}{R_3} + \frac{0 - V_o}{\frac{1}{sC_5}} \tag{6.7}$$

$$I_z = \frac{V_x}{R_3} - V_o\left(\frac{1}{R_3} + sC_5\right) \tag{6.8}$$

$$I_x = \frac{0 - V_x}{R_4} + \frac{V_o - V_x}{R_3} \tag{6.9}$$

From equation:

$$I_x = I_z \tag{6.10}$$

$$\frac{0 - V_x}{R_4} + \frac{V_o - V_x}{R_3} = \frac{V_x}{R_3} - V_o\left(\frac{1}{R_3} + sC_5\right) \tag{6.11}$$

$$V_x\left(\frac{2}{R_3} + \frac{1}{R_4}\right) = V_x\left(\frac{2}{R_3} + sC_5\right) \tag{6.12}$$

From equation 6.6

$$V_x = V_y \tag{6.13}$$

$$V_x = V_o\left(\frac{sR_2C_1}{1 + sC_1R_2}\right) \tag{6.14}$$

From equation 6.13 and 6.14

$$V_o\left(\frac{sR_2C_1}{1 + sC_1R_2}\right)\left(\frac{2}{R_3} + \frac{1}{R_4}\right) = V_o\left(\frac{2}{R_3} + sC_5\right) \tag{6.15}$$

$$V_0\{\frac{sR_2C_1}{1+sC_1R_2}\left(\frac{2R_4+R_3}{R_3R_4}\right)-(\frac{2+R_3C_5}{R_3})\} = 0 \quad (6.16)$$

$$V_o\{\frac{2sR_2R_4C_1+sR_2R_3C_1-(R_4+s\,C_1R_2R_4)(2+sR_3C_5)}{R_3R_4(1+sC_1R_2)}\} = 0 \quad (6.17)$$

After solving further

$$s^2C_1C_5R_2R_3R_4 + (R_4R_3C_5 - R_2R_3C_1)s + 2R_4 = 0 \quad (6.18)$$

Which is a equation of oscillator

$$as^2 + (\alpha - \beta)s + \gamma = 0 \quad (6.19)$$

Putting s = jω in equation 6.18

$$(j\omega)^2\,C_1C_5R_2R_3R_4 + (R_4R_3C_5 - R_2R_3C_1)(j\omega) + 2\,R_4 = 0 \quad (6.20)$$

$$-\omega^2\,C_1C_5R_2R_3R_4 + j\omega(R_4R_3C_5 - R_2R_3C_1) + 2\,R_4 = 0 \quad (6.21)$$

Equating real part and imaginary part

Real part = 0

$$2\,R_4 - \omega^2\,C_1C_5R_2R_3R_4 = 0$$

$$\omega^2 = \frac{2}{C_1C_5R_2R_3}$$

$$f = \frac{1}{\sqrt{2}\,\pi\sqrt{(R2\ R3\ C1\ C5)}} \quad (6.22)$$

Imaginary part =0

$$R_4R_3C_5 = R_2R_3C_1$$

$$\frac{C_5}{C_1} = \frac{R_2}{R_4} \quad (6.23)$$

This is the condition of oscillation.

6.3 SIMULATION RESULT OF OSCILLATION

When we select the component, value of resistor and capacitor as shown in figure 6.1. which satisfy condition of oscillation, on tuning of resistor and capacitor oscillation comes in orcad PSPICE simulator in the form of sinusoidal wave Practically $\frac{R_2}{R_4}$ is smaller than 1 and $\frac{C_5}{C_1}$ is also smaller than 1 which is a difference between theoretical and practical oscillation.

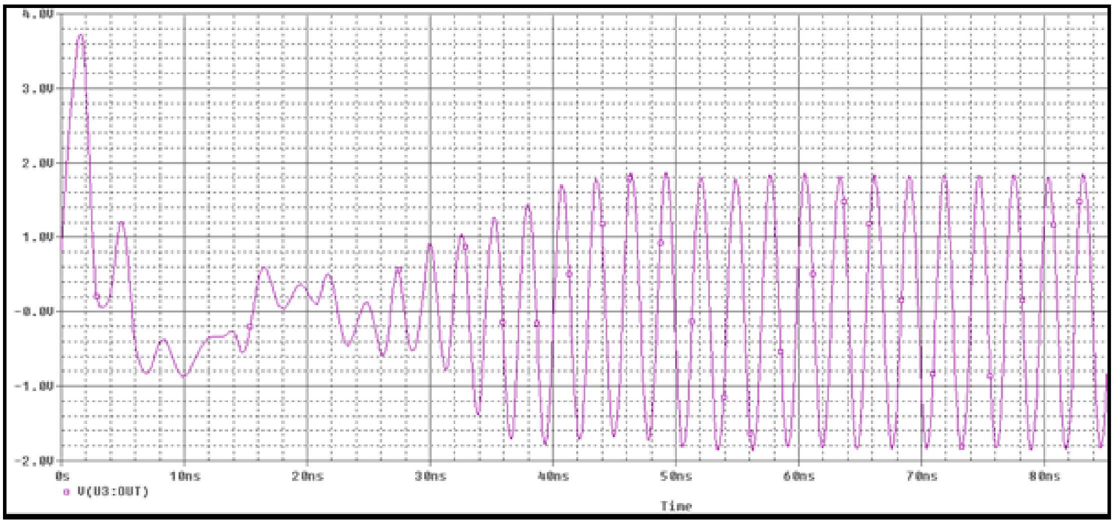


Figure 6.2 (a) Building of sine wave oscillation in MCFOA circuit.

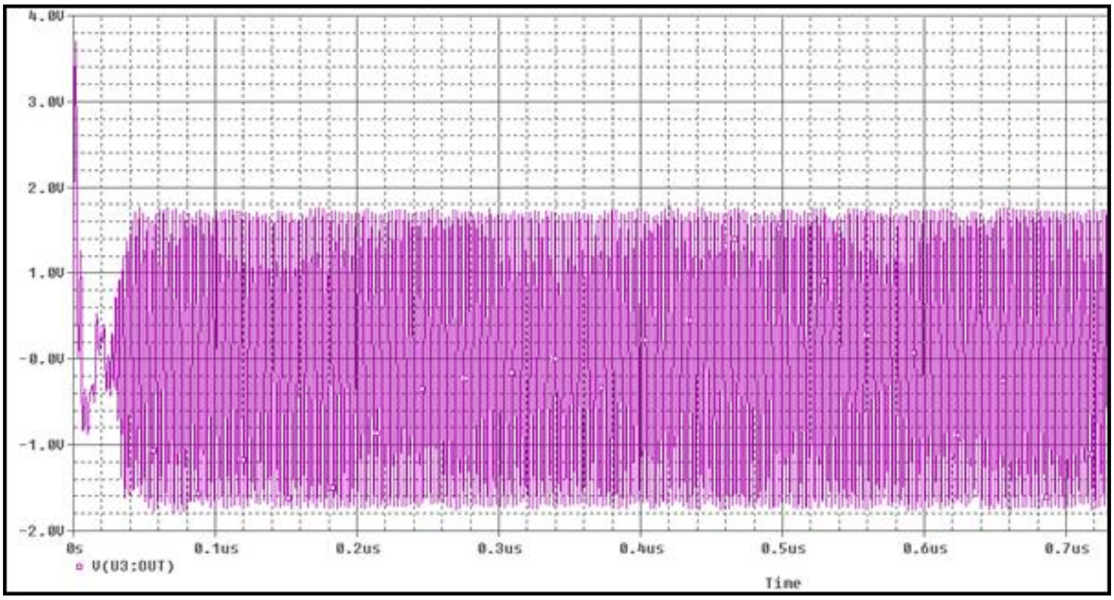


Figure 6.2 (b) Building of sine wave oscillation in MCFOA circuit.

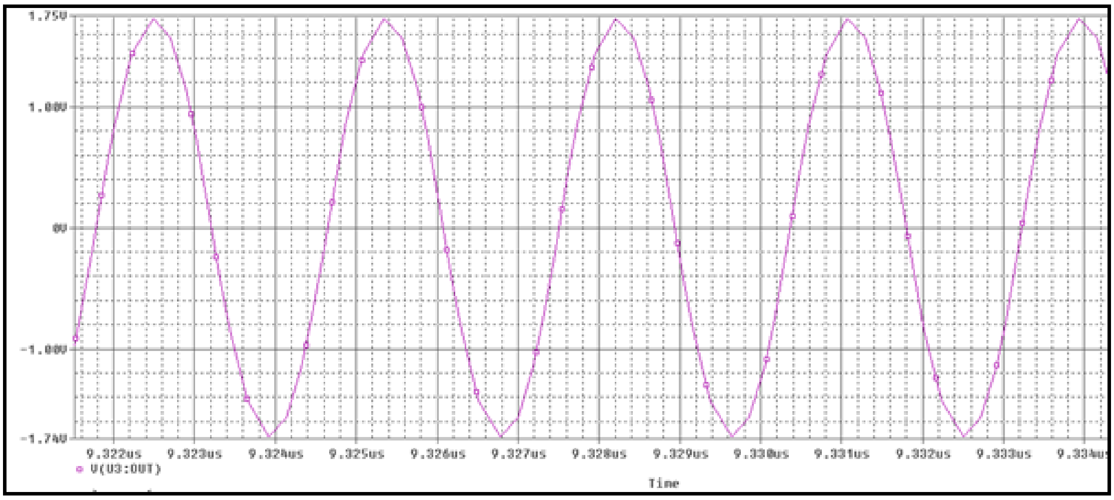


Figure 6.3 Sine wave oscillations by MCFOA circuit.

When we take the capacitor C_1=1.1pf, C_5=1pf, R_2=1kΩ, R_3=1kΩ, R_4=1.2kΩ. After varying the capacitor C_1 we get the practically frequency of 333.33 MHz.

6.4 INTEGRATOR CIRCUIT WITH MCFOA

The integrator realization with the MCFOA is shown in Figure 6.4. The transfer function for this circuit is to be found. Here the positive terminal for the MCFOA is grounded. Thus An integrator using a MCFOA and its bode plot are shown in Fig. 6.4 and 6.5 respectively. The transfer function for the integrator can be found as

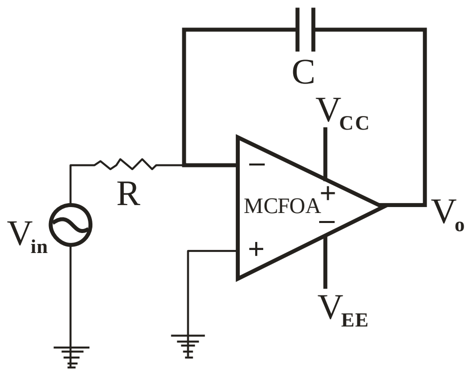


Figure 6.4 MCFOA integrator.

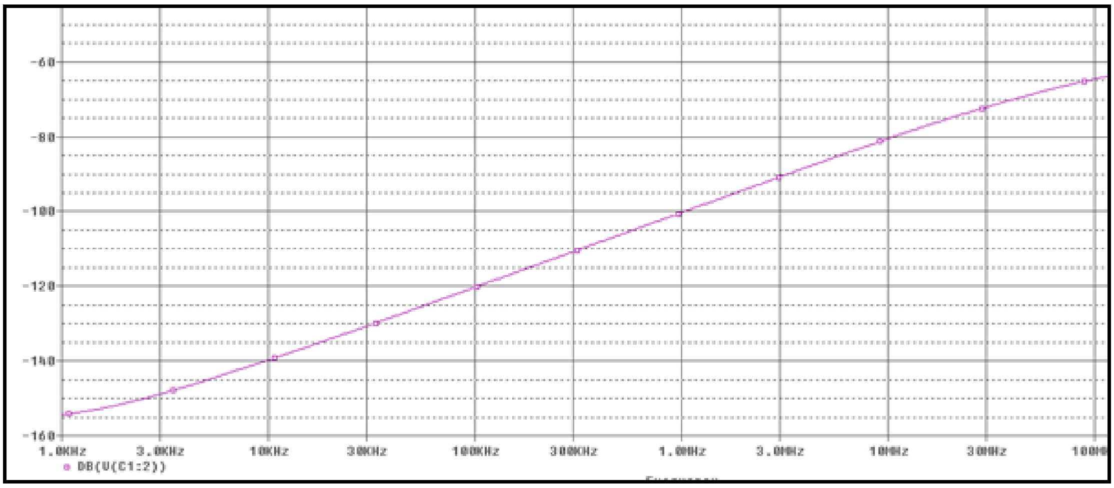


Figure 6.5 bode plot of MCFOA integrator

$$\frac{V_{in} - V_-}{R} = \frac{V_- - V_o}{R} \tag{6.24}$$

$$V_- = V_+ \quad \text{and } V_+ = 0$$

$$\frac{V_o}{V_{in}} = \frac{1}{sRC} \tag{6.25}$$

$$\frac{V_o}{V_{in}} = \frac{1}{s\tau} \tag{6.26}$$

Where τ = RC , time constant

Division by s in the frequency domain corresponds to integration in the time domain. Due to the negative sign in the transfer function, this integrator is an inverting integrator.

6.5 TRIANGULAR AND RECTANGULAR WAVE GENERATOR USING MCFOA

The triangular and rectangular wave generator circuit is shown in Fig 6.6. The triangular wave can be generated by alternately charging and discharging a capacitor with a constant current. For this circuit, current drive for the capacitor C_1 is provided by the $MCFOA_1$. In the second circuit, $MCFOA_2$, is configured as a Schmitt trigger.

Schmitt trigger output would be at the ±5V of the maximum output voltage of the MCFOA.

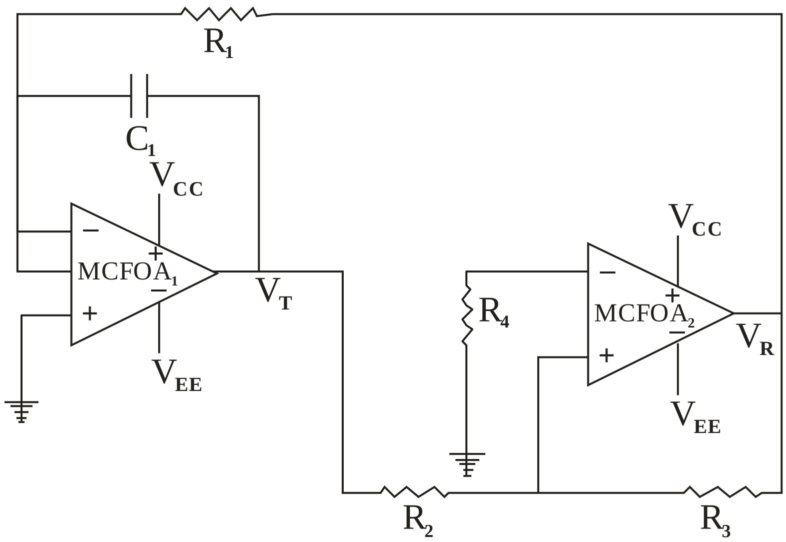
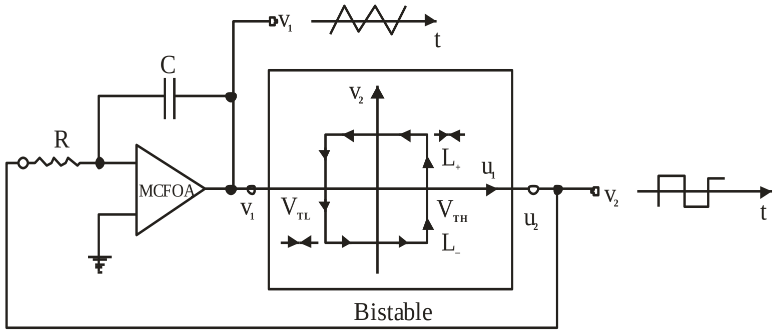


Figure 6.6 Triangular and rectangular wave generator using MCFOA.

The frequency of oscillation of this circuit can be found by the following process. The integrator with capacitor in the feedback path and resistor in the inverting input terminal with branch currents are shown in the Figure 6.6.

The waveforms generated in the astable circuit of Fig. 6.9 can be changed to triangular by replacing the low-pass RC circuit with an integrator. (The integrator is, after all, a low-pass circuit with a corner frequency at dc.) The integrator causes linear charging and discharging of the capacitor, thus providing a triangular waveform. The resulting circuit is shown in Fig. 6.7(a). Observe that because the integrator is inverting, it is necessary to invert the characteristics of the bistable circuit.

We now proceed to show how the feedback loop of Fig. 6.7(a) oscillates and generates a triangular waveform v_1 at the output of the integrator and a square waveform v_2 at the output of the bistable circuit: Let the output of the bistable circuit be at (positive saturation) L_+. A



(a)

Figure 6.7(a) A general scheme for generating triangular and square waveforms.

Current equal to L_+/R will flow into the resistor R and through capacitor C, causing the output of the integrator to linearly decrease with a slope $-L_+/CR$, as shown in Fig. 6.7(c). This will continue until the integrator output reaches the lower threshold V_{TL} of the bistable circuit, at which point the bistable circuit will switch states, It's output becoming negative and the equal to

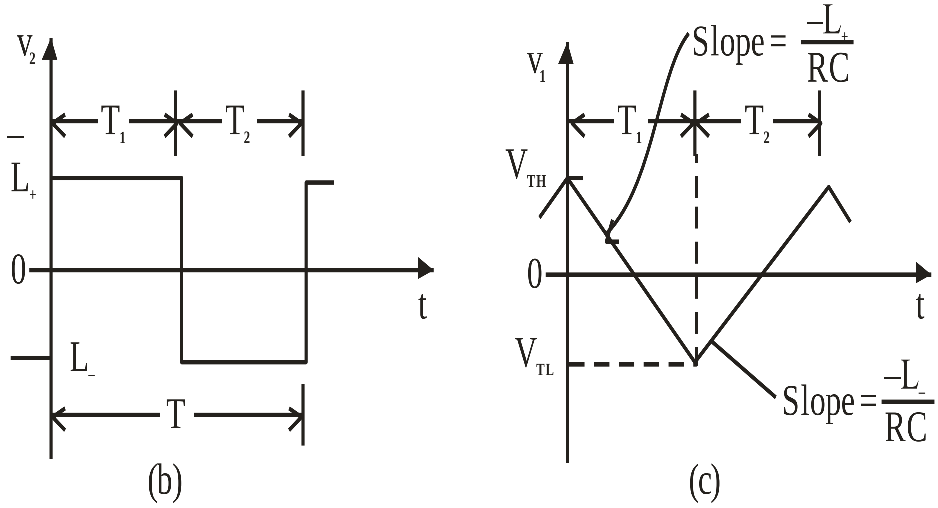


Figure 6.7 (b) and (c) A general scheme for generating triangular and square waveforms.

L_- (negative saturation). At this moment the current through R and C will reverse direction, and its value will become equal to $|L_-|/R$. It follows that that the integrator output will start to increase linearly with a positive slope equal to $|L_-|/CR$. This will continue until the integrator output voltage reaches the positive threshold of bistable circuit, V_{TH}. At this point the bistable circuit switches, its output becomes positive (L_+), the current into the integrator reverse direction, and output of the integrator starts to decrease linearly, beginning a new cycle.

From the discussion above it is relatively easy to derive an expression for the period T of the square and triangular waveforms. During the interval T_1 we have, from Fig. 6.7(c),

$$\frac{V_{TH} - V_{TL}}{T_1} = \frac{L_+}{CR}$$

From which we obtain

$$T_1 = CR\frac{V_{TH}-V_{TL}}{L_+} \qquad (6.27)$$

Similarly, during T_2 we have

$$\frac{V_{TH}-V_{TL}}{T_2} = \frac{-L_-}{CR}$$

From which we obtain

$$T_2 = CR\frac{V_{TH}-V_{TL}}{-L_-} \qquad (6.28)$$

Thus to obtain symmetrical square waves we design the bistable circuit to have $L_+ = -L_-$.

6.6 SIMULATION AND RESULT OF RECTANGULAR AND TRIANGULAR WAVE

For finding the oscillation we used equation 6.27 and 6.28, and value of capacitor is 20pF and resistor is 4kΩ. On tuning the capacitor we get the oscillation. Frequency of rectangular wave is about 30MHz and frequency of triangular wave is also near to 30MHz. the results of simulation are shown in figure 6.9 and 6.10.

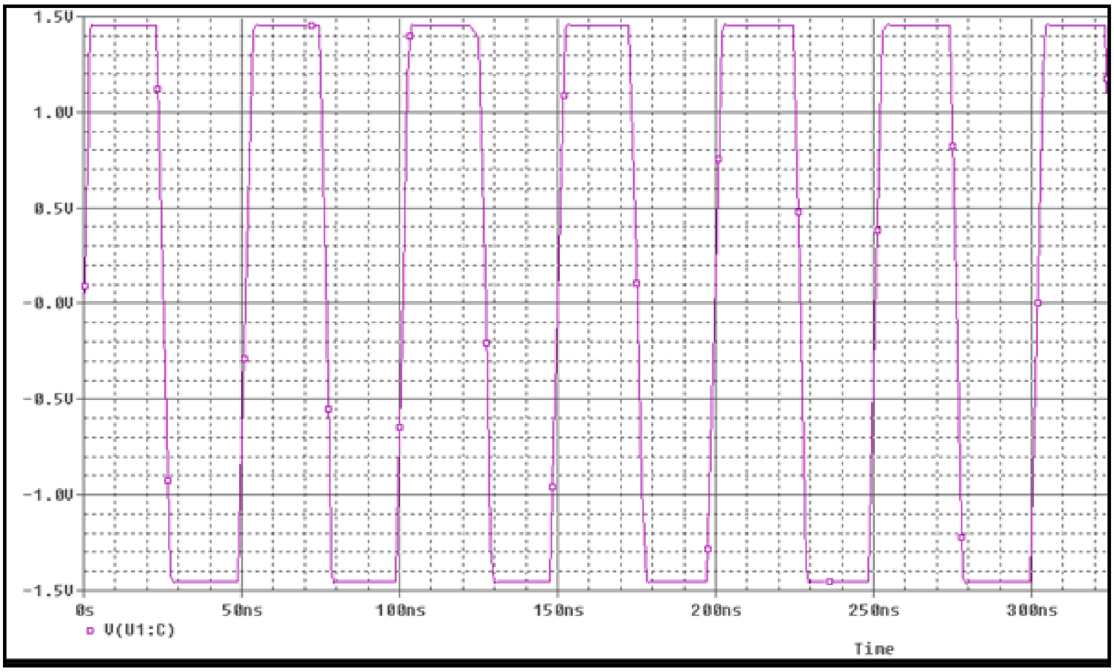


Figure 6.9 Generation of rectangular wave using MCFOA.

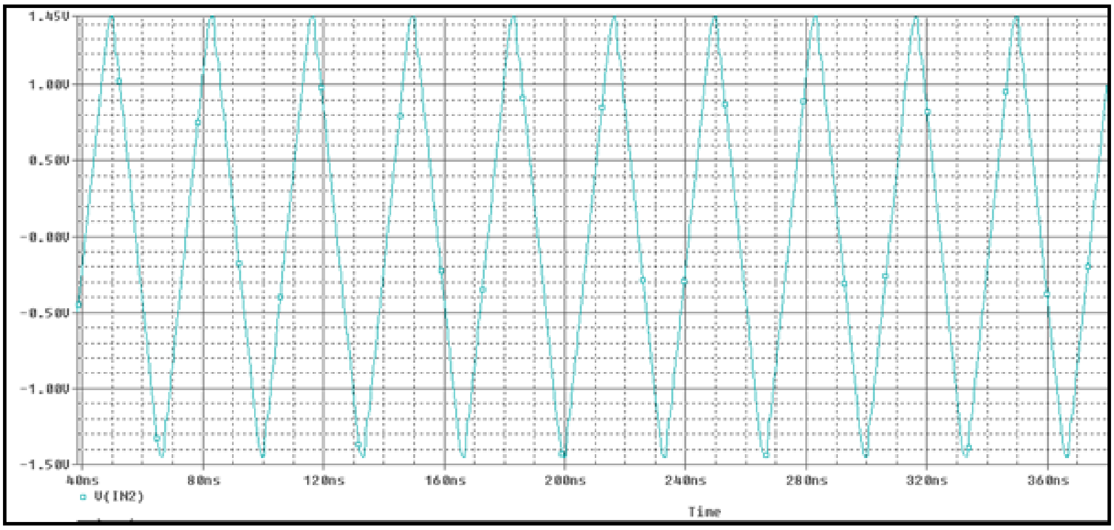


Figure 6.10 Generation of triangular wave using MCFOA.

6.7 SUMMARY

Analysis and simulation of MCFOA is done by using three AD844 8- pin IC of analog device. Here we build sine wave of 333MHz, integrator, rectangular wave, and triangular wave of 20MHz. simulation is done on orcad pspice 16.5 version. Here we achieved higher oscillation than CFOA function generator. This can be useful for faster device. So MCFOA are much better than CFOA. This is proven in this chapter and above simulation.

CHAPTER-7
CONCLUSION AND FURTHER SCOPE

7.1 CONCLUSION

A detailed comparison between a CFA and a VFA is presented in this thesis. The work confirms that the CFOA provides a higher gain-bandwidth product. In this case, the gain-bandwidth product of the CFOA is twice as large as the VOA (and in general, never higher than four times). Sinusoidal, triangular and square wave generator using CFOA has been presented. The advantages of using CFOA over VFOA in designing oscillator also presented. CFOA is giving higher oscillation than voltage feedback amplifier (VFA). Theoretical oscillation frequency agrees with the simulated oscillation frequency. After simulation result of CFOA we worked on MCFOA. Sinusoidal, triangular and square wave generator using MCFOA has been presented. The advantages of using MCFOA over CFOA in designing oscillator also presented and they are giving much higher oscillation than CFOA. All the analytical results which were developed in this work were lastly validated through simulations using ORCAD 16.5. A very good agreement between the expected data and simulations was encountered.

7.2 FUTURE SCOPE OF WORK

We did some theoretical study of MCFOA based on CMOS, which can bring higher oscillation than BJT based MCFOA. Higher frequency oscillation is demand of current technology in the world. So, by using MCFOA based on CMOS can fulfil this requirement.

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