CHAPTER – 5
SIGNAL GENERATORS USING OTRA

The content and results of the following papers have been reported in this chapter.


5.1 INTRODUCTION

Signal generators are an important class of circuits and find wide application in communication, instrumentation and measurement, and control systems. Generating signal carriers for information transmission, clock pulses for timing and control, test signals for automatic test and measurement and audio signals for music and speech synthesis are some of the most common examples. These systems require signals of various standard shapes such as sinusoidal, square, triangular, sawtooth etc. Thus the function of a signal generator is to produce a waveform of prescribed characteristics such as shape, frequency, amplitude, and duty cycle. Sometimes these characteristics are designed to be programmable via some suitable external control signals. In general, signal generators employ some form of feedback together with devices possessing frequency dependent characteristics to produce signals. Signal generators can broadly be classified as (i) Harmonic/Sinusoidal oscillators and (ii) Non linear/Relaxation oscillators. Sinusoidal oscillators are also termed as linear oscillators. In this chapter design of various sinusoidal oscillators using OTRA is dealt with. In the following section a review of reported work on sinusoidal oscillators is presented.

5.2 SINUSOIDAL OSCILLATOR

The sine wave is one of the most fundamental waveforms since any other waveform can be expressed as a Fourier combination of basic sine waves. Sinusoids can be generated either by appropriate shaping the triangular waveform or through a positive feedback loop consisting of an amplifier and a frequency selective network. All the proposed circuits discussed in this chapter utilize the positive feedback approach which is described briefly in this section.

The functional block diagram of a positive feedback loop consisting an amplifier having transfer function $A(s)$ and a frequency selective network with transfer function $\beta(s)$ is shown in Fig. 5.1, wherein $x_i$ and $x_o$ represent the input and output signals respectively and $x_f$ is the feedback signal. The closed loop gain $A_f(s)$ of this system can be derived as

$$A_f(s) = \frac{A(s)}{1-A(s)\beta(s)}$$  \hspace{1cm} (5.1)
The characteristic equation of this system can thus be written as

\[ 1 - A(s)\beta(s) = 0 \]  

(5.2)

where \( A(s)\beta(s) \) is called the loop gain \( L(s) \). If at a specific frequency the loop gain of this system is equal to unity then \( A_f \) will be infinite implying that at this frequency the circuit will have a finite output for zero input signal. Such a circuit is by definition is an oscillator. Thus the condition for the closed loop system to provide sustained oscillations of frequency \( \omega_0 \) can be represented as

\[ L(j\omega_0) = A(j\omega_0)\beta(j\omega_0) = 1 \]  

(5.3)

This condition is known as the Barkhausen criterion. For the circuit to oscillate at a single frequency the oscillation criterion must be satisfied for a single frequency only; else the resulting waveform will not be a simple sinusoid.

Fig. 5.1 Functional block of a sinusoidal oscillator.

In following subsections sinusoidal oscillators that provide single phase, quadrature phase and multiphase outputs have been proposed. All the proposed circuits utilize the positive feedback approach and Barkhausen criterion is satisfied for a single frequency only.

**5.2.1 Single Phase Oscillator**

Only a few OTRA based single phase oscillator circuits [61], [65] are available in literature
and a detailed study of these structures suggests that these circuits

- Use floating passive element [65]
- Do not have non interactive control on CO and FO [61], [65]
- Do not support equal component usage which is preferred from integrated circuit implementation viewpoint [65]
- Have complex CO and/or FO [61], [65]

Therefore in the following section a single phase oscillator is presented, having all the passive components virtually grounded with non interactive control on FO and CO and also supports equal component design.

### 5.2.1.1 Proposed Circuit

The proposed oscillator circuit is shown in Fig. 5.2. Using routine analysis the characteristic equation of the proposed oscillator can be expressed as

$$s^2 C_1 C_3 + s(C_1 G_3 - C_2 G_1) + G_1 G_2 = 0 \quad \text{where} \quad G_l = \frac{1}{R_l}$$

(5.4)

![Fig. 5.2 The proposed sinusoidal oscillator.](attachment:image_url)

From (5.4) the condition of oscillation (CO) and frequency of oscillation (FO) can
respectively be derived as

\[ CO: \quad c_1 g_3 = c_2 g_1 \]  

\[ FO: \quad f_0 = \frac{1}{2\pi} \sqrt{\frac{g_1 g_2}{c_1 c_3}} \]  

(5.5)  

(5.6)

It is clear that FO can be controlled independent of CO through resistance \( R_2 \); and it is also possible to achieve independent control of CO by varying \( C_2 \).

**5.2.1.2 MOS-C Implementation**

The proposed configuration can be made fully integrated by implementing the resistors using matched transistors operating in linear region as explained in section 2.8. The MOS-C implemented oscillator is shown in Fig. 5.3. The resistance value may be adjusted by appropriate choice of gate voltages thereby making CO and FO electronically tunable. It also exhibits the feature of orthogonal controllability of \( \omega_0 \) and \( Q_0 \) through gate bias voltage.

![Fig. 5.3 MOS-C Implemented Oscillator.](image-url)
5.2.1.3 Sensitivity Analysis

The passive sensitivities of the frequency of oscillation of the proposed oscillators may be computed as

\[ S_{C_1}^{\omega_0} = S_{C_3}^{\omega_0} = S_{R_1}^{\omega_0} = S_{R_2}^{\omega_0} = -\frac{1}{2} \quad (5.7) \]

All passive sensitivities are lower than unity in magnitude.

5.2.1.4 Nonideality Analysis

In analysis so far, ideal characteristics of the OTRA have been considered. However, due to the parasitics of the OTRA the behaviour of the proposed circuit may deviate. Taking this effect into account (5.4) modifies to

\[ s^2(C_1 + C_p)(C_3 + C_p) + s((C_1 + C_p)G_3 - C_2G_1) + G_1G_2 = 0 \quad (5.8) \]

Hence the modified CO and FO can be computed to be

\[ (C_1 + C_p)G_3 = C_2G_1 \quad (5.9) \]

\[ f_0 = \frac{1}{2\pi} \sqrt{\frac{G_1G_2}{(C_1+C_p)(C_3+C_p)}} \quad (5.10) \]

However from (5.8) it may be noted that the effect of \( C_p \) can be absorbed in \( C_1 \) and \( C_3 \) without increasing the order of the circuit and hence achieving self compensation.

5.2.1.5 Simulation and Experimental Results

To verify the theoretical predictions, an oscillator for frequency of 159 KHz is designed and simulated through SPICE using 0.5µm, CMOS process parameters provided by MOSIS (AGILENT). The CMOS schematic of OTRA [38] as shown in Fig. 2.9 is used for simulations. The resistive component values for this design are computed as \( R_1 = 10 \, \text{K}\Omega \), \( R_2 = 50 \, \text{K}\Omega \), \( R_3 = 10 \, \text{K}\Omega \) for chosen values of \( C_1 = 20 \, \text{pF} \), \( C_2 = 30 \, \text{pF} \) and \( C_3 = 100 \, \text{pF} \). A
simulated output of the designed circuit is depicted in Fig. 5.4(a). The simulated output frequency is observed to be 156.82 KHz and is in close agreement with the theoretical value. The frequency spectrum of the output is shown in Fig. 5.4(b). The % THD was observed to be 4.79 %. The working of the proposed oscillator is verified experimentally as well, using

![Simulated result](image1)

![Frequency spectrum](image2)

![Experimental result](image3)

Fig. 5.4 Oscillator output. (a) Simulated result. (b) Frequency spectrum. (c) Experimental result.
commericially available IC AD844 to implement OTRA. The component values are chosen as chosen $R_1 = R_3 = 10 \text{ K}\Omega$, $R_2 = 51 \text{ K}\Omega$, $C_1 = 22 \text{ pF}$, $C_2 = 33 \text{ pF}$ and $C_3 = 100 \text{ pF}$. The experimental output is depicted in Fig. 5.4(c) having experimental FO as 143.2 KHz. Minor variations in experimental results can be attributed to component values difference and the CFOA parasitics.

5.2.2 Quadrature Oscillator

Quadrature oscillators (QO) provide two single frequency outputs which are in quadrature phase. QOs are widely used in the field of communication, instrumentation and power electronics. In communications QO circuits are commonly used in single-sideband generators, and quadrature mixers [89]. They are also utilized for vector generators or selective voltmeters [89]. QO can be designed as either second order or third order systems. Compact realization is one of the key advantages of second order design as it uses fewer numbers of passive and/or active components whereas third order QO gives better accuracy, frequency response, and low harmonic distortion. One can trade off compact realization, for better distortion performance achieved out of third order realization, depending upon the application. Few second order QOs using OTRA [36], [51], [53], [57] are available in open literature however no third order QO design has been reported so far. Tapping this gap and keeping the advantage of third order QO in view, an OTRA based third order QO has been designed and is presented in following subsection.

5.2.2.1 Proposed Circuit

The general scheme of third order QO, as depicted in Fig. 5.5, is discussed in [90] and has been adapted for implementation with OTRA. It consists of a second order LPF having transfer function $A(s)$ and an inverting integrator in feedback loop with transfer $\beta(s)$. This forms a closed loop system having loop gain as $A(s)\beta(s)$ and the criterion for oscillations to occur, is given by

$$1 - A(s)\beta(s) = 0$$  (5.11)
If this criterion is satisfied the closed loop system will produce sustained oscillations. The voltages $v_1$ and $v_2$ will have quadrature phase relationship as $v_2$ is integrated output of $v_1$.

![Functional block diagram of QO](image1)

Fig. 5.5 Functional block diagram of QO [90].

The proposed third order QO circuit is shown in Fig. 5.6. It consists of an OTRA based second order LPF [44] and an inverting integrator connected in feedback. The LPF consists of OTRA 1 and associated circuitry, whereas the OTRA 2 and passive components $R_1$ and $C_1$ form the inverting integrator.

![OTA based QO](image2)

Fig. 5.6 OTRA based QO.
Using routine analysis the characteristic equation of circuit can be expressed as

\[ s^3C^2C_1 + 2s^2C_1GC + sc_1G^2 + G^2G_1 = 0 \]  \hspace{1cm} (5.12)

where \( G = \frac{1}{R} \) and \( G_1 = \frac{1}{R_1} \).

The CO and FO can be derived from this characteristic equation of (5.12) as

\[
\text{FO: } f_0 = \frac{G}{2\pi C} \\
\text{CO: } \frac{G}{C} = \frac{G_1}{2C_1}
\]

By suitable selection of \( G \) and \( C \) values the FO can be adjusted to desired value and proper selection of \( G_1 \) and \( C_1 \) would satisfy the CO.

### 5.2.2.2 MOS-C Implemented QO

The resistors connected to the input terminals of OTRA can be implemented using MOS transistors operating in non-saturation region thereby making oscillator circuit MOS-C realizable and the oscillator frequency electronically tunable. The MOS-C implementation of the proposed circuit is shown in Fig. 5.7.
5.2.2.3 Nonideality Analysis

The output of the QO may deviate due to non-ideality of OTRA in practice. Considering the effect of nonideality of OTRA into account (5.12) modifies to

\[ s^3 C(C + C_p)(C_1 + C_p) + s^2 G(C_1 + C_p)(2C + C_p) + sG^2(C_1 + C_p) + G^2 G_1 = 0 \]  \hspace{1cm} (5.15)

From (5.15) it is found that the FO changes to

\[ f_0 = \frac{G}{2\pi \sqrt{C(C_p+C)}} \]  \hspace{1cm} (5.16)

The modified CO can be expressed as

\[ \frac{G}{C} = \frac{G_1(C+C_p)}{(C_p+C_1)(2C+C_p)} \]  \hspace{1cm} (5.17)

The effect of \( C_p \) can be eliminated by pre-adjusting the value of capacitors \( C \) and \( C_1 \), which are connected in feedback path of the low pass filter and in inverting integrator respectively, thus achieving self compensation.

5.2.2.4 Simulation Results

The functionality of the proposed QO is verified through SPICE simulations using CMOS schematic of OTRA shown in Fig. 2.9 with 0.5 \( \mu \)m AGILENT CMOS process parameters. For all transistors used for resistance realization W/L ratio is taken as 10\( \mu \)m/2\( \mu \)m. The proposed QO is designed to work at 159 KHz by selecting the component values as \( C_1 = C = 100 \) pF, \( R = 10 \) K\( \Omega \) and \( R_1 = 5 \) K\( \Omega \). The gate voltage \( V_g \) and \( V_{g1} \) are set to be 0.86 V and 0.73 V respectively while keeping \( V_a = V_{a1} = 1.0 \) V for designing \( R = 10K\Omega \), \( R_1 = 5K\Omega \). Simulated FO was observed to be 150 KHz. The resistance values designed using MOSFETs might differ slightly from the theoretically calculated values causing slight deviation between theoretical and simulated FO. The output of the QO is shown in Fig. 5.8 (a) and Fig. 5.8 (b) depicts the frequency spectrum. The % THD was observed to be 4.1%.
5.2.3 Multiphase Sinusoidal Oscillator

Multiphase sinusoidal oscillators (MSO) provide n outputs equally spaced in phase and find extensive application in the field of power electronics and communications. In
communications MSO circuits are commonly used in single-sideband generators, and phase modulators. They are also utilized for control of single phase-to-three-phase PWM converters [91], and for a decoupled dynamic control of a six-phase two-motor drive system [92].

An extensive literature review revealed that OTRA based MSO circuits have not been reported earlier. Exploring this void three MSO circuits are proposed and their detailed description is presented in following section. The first circuit utilizes \( n \) OTRAs such that \( n \geq 3 \), to produce \( n \) odd-phase oscillations which are equally spaced in phase and are of equal amplitude. The second circuit utilizes \((n+1)\) OTRAs having \( n \geq 4 \), to produce \( n \) odd or even phase oscillations equally spaced in phase. The third circuit utilizes a single-resistance-controlled (SRCO) sinusoidal oscillator circuit employing a single OTRA [65], whose output is subsequently used to drive a phase shifter network. The phase shifter circuit consists of OTRA based \((n-1)\) noninverting LPFs giving a total of \( n \) phase oscillations. The circuit is tunable and has a low component count. An Automatic Gain Control circuitry (AGC) has also been implemented for the second and third circuit, which helps in the stabilization of the signal amplitude.

**5.2.3.1 Proposed Circuit-I**

The first circuit is based on the scheme discussed in [93] and has been adapted for implementation with OTRA. It consists of \( n \) cascaded stages of first-order inverting LPFs. The output of the \( n \)th stage is fed back to the input of the first stage as shown in Fig. 5.9 (a). The OTRAs based circuit implementation of this scheme is shown in Fig. 5.9 (b). The OTRAs have been connected in the inverting mode such that the gain \( G(s) \) of each block can be expressed as

\[
G(s) = \left( -\frac{K}{1 + sCR} \right)
\]  

(5.18)

where \( K = \frac{R}{R_1} \)
From Fig. 5.9 (b), the open loop gain $L(s)$ can be expressed as

$$L(s) = \left( - \frac{K}{1+sCR} \right)^n$$

(5.19)

For oscillations to occur, the Barkhausen criterion [84] must be satisfied, hence

$$\left( - \frac{K}{1+sCR} \right)^n = 1$$

(5.20)

The above equation yields

$$(1 + sCR)^n + (-1)^{n+1} K^n = 0$$

(5.21)
Equation (5.21) will converge only for odd values of \( n \) such that \( n \geq 3 \). Thus the circuit will give rise to equally spaced oscillations having a phase difference of \((360/n) \degree\). Considering the case for \( n = 3 \) then (5.21) reduces to

\[
(1 + j\omega_0 CR)^3 + K^3 = 0
\]  

(5.22)

Equating real and imaginary parts of (5.22) gives the FO and CO as

**FO:** \( f_0 = \sqrt{3} / (2\pi RC) \)  

(5.23)

**CO:** \( K = 2 \)  

(5.24)

Similarly for \( n = 5 \), (5.21) would reduce to

\[
(1 + j\omega_0 CR)^5 + K^5 = 0
\]  

(5.25)

Hence FO and CO can be obtained as

**FO:** \( f_0 = 0.727 / (2\pi RC) \)  

(5.26)

**CO:** \( K = 1.236 \)  

(5.27)

The proposed circuit is simple to realize and has a low component count. It produces \( n \) odd-phase oscillations of equal amplitudes with a phase difference of \((360/n) \degree\) and an iterative control of CO and FO can be achieved for the oscillator.

### 5.2.3.2 Proposed Circuit-II

In this subsection another scheme is used which is capable of producing \( n \) odd or even phase oscillations. The block diagram of this scheme is shown in Fig. 5.10(a) which employs \( n \) cascaded non inverting first order LPFs and an inverting voltage amplifier circuit in the feedback loop. The OTRA based implementation of this scheme is shown in Fig. 5.10(b). An inverting voltage amplifier with a simple AGC circuit is connected in the feedback loop of the oscillator. For stabilization of the signal amplitude the inverting voltage amplifier can be designed with automatic gain control circuitry (AGC). In OTRA based MSO of Fig. 5.10(b),
the diodes $D_1$ and $D_2$ along with resistors $R_4$, and $R_5$ constitute the AGC circuit. The gain $G(s)$ of each block can be computed as

$$G(s) = \left( \frac{K}{1+sCR} \right)$$

(5.28)

Fig. 5.10(a) Generalized scheme for producing $n$ – odd / even phase oscillations. (b) OTRA based circuit with AGC.
where $K = (R/R_1)$. The loop gain for the system can be expressed as

$$L(s) = (-1)K_X \left( \frac{K}{1+sCR} \right)^n$$  \hspace{1cm} (5.29)$$

$$K_X = \frac{R_3}{R_2}$$  \hspace{1cm} (5.30)$$

$K_X$ is effectively maintained at a value 1 with the help of the AGC circuitry. After applying the Barkhausen criterion the characteristic equation is obtained as

$$(1 + sCR)^n + K^n = 0$$  \hspace{1cm} (5.31)$$

The equation converges for all value of $n \geq 3$, odd or even. As an example for $n = 4$, (5.31) reduces to

$$(1 + j\omega_0 CR)^4 + K^4 = 0$$  \hspace{1cm} (5.32)$$

which gives FO and CO as

$\text{FO: } f_0 = 1/(2\pi RC)$  \hspace{1cm} (5.33)$$

$\text{CO: } K = 1.414$  \hspace{1cm} (5.34)$$

For $n = 3$, the characteristic equation, FO and CO would be same as (5.25), (5.26) and (5.27) respectively. Initially $K_X$ is kept at a value slightly higher than 1 so that the oscillations can begin, once the amplitude crosses a certain threshold the diodes get switched on and bring down the value of resistance $R_3$ thus bringing down the effective value of $K_X$. This is possible because the input terminals of the OTRA are virtually grounded. Thus, a dynamic equilibrium maintains the value of $K_X$ at 1. Noninteractive tuning of this circuit is not possible as CO and FO are not independent. However, the circuit is versatile and can achieve both even and odd phase oscillations. The oscillations achieved are equally spaced in phase having a phase difference of $(180/n)^\circ$. 
5.2.3.3 Proposed Circuit-III

The scheme of third circuit is shown in Fig. 5.11(a). It uses an SRCO followed by (n-1) phase shifter blocks thus producing n phase shifted outputs. The OTRA based configuration is shown in Fig. 5.11(b). It is based on a SRCO oscillator proposed in [65] to which an AGC circuit similar to one used in circuit II has also been added. The open loop gain of this circuit [65] is obtained as

\[
L(s) = \left( \frac{sC_2R_2R_3(1-sC_1R_1)}{R_1+R_3+sR_1R_3(C_1+C_2)} \right)
\]

(5.35)

![Diagram of Proposed Circuit-III](image)

Fig. 5.11 (a) SRCO based MSO. (b) OTRA based implementation.
Accordingly, the FO and CO for SRCO oscillator are obtained as

\[ f_0 = \frac{1}{2\pi} \sqrt{\frac{R_1 + R_3}{C_1 C_2 R_1 R_2 R_3}} \]  
\[ \frac{R_2}{R_1} = \frac{C_1}{C_2} + 1 \]  

It can be observed that by controlling R₃ the frequency can be controlled without affecting the CO, which makes the circuit tunable. AGC has been achieved by adjusting R₂, which is used to control the loop gain, as seen from (5.35). At the output of the SRCO with AGC, (n-1) subsequent OTRA based phase shifter blocks can be connected to produce n oscillation. The phase shift produced by each phase shifter block can be given as

\[ \theta = \tan^{-1}(\omega_0 CR) \]  

To add to the flexibility, if the OTRA in the phase shifter block is connected in inverting mode the phase shift produced will be

\[ \theta = 180° - \tan^{-1}(\omega_0 CR) \]  

Hence, the phase shifter can be adjusted to obtain a phase shift of either 0°-90° or 90°-180° depending on its configuration. This circuit provides the flexibility of achieving the desired phase shift without connecting too many active elements.

5.2.3.4 Nonideality Analysis

The output of the proposed circuits may deviate from the theoretical predictions due to nonidealities associated with CFOAs used for realizing OTRA. In this section the nonideality analysis for Circuit II which provides both even and odd phase outputs, is presented. Considering the nonideal model of CFOA as shown in Fig. 2.8, the non inverting OTRA based LPF block can be redrawn as shown in Fig. 5.12. KCL at nodes 1 and 2 can respectively be written as
Assuming $R_{x1} = R_{x2} = R_x$ and using the relation $R_x \ll R_{z1}$, from (5.40) and (5.41) the transfer function of non inverting LPF can be expressed as

$$\frac{v_o}{v_{in}} = \frac{K (1+sCR_x)}{(1+sCR)} \tag{5.42}$$

Using (5.42) the loop gain given by (5.29) modifies and can be represented as

$$L(s)_{in} = (-1)K_x \left( \frac{K(1+sCR_x)}{1+sCR} \right)^n \tag{5.43}$$

![Diagram](image.png)

Fig. 5.12 Nonideal model of the non inverting LPF.
Assuming $K_x = 1$, the modified characteristic equation can be written as

$$(1 + sCR)^n + K^n (1 + sCR_x)^n = 0$$

(5.44)

For $n = 3$ the CO and FO for the circuit can be determined as

**FO:**

$$f_0 = \frac{\sqrt{3}(R+R_x K^3)}{2\pi \sqrt{(R^3 + K^3 R_x^3) C}}$$

(5.45)

**CO:**

$$(K^3 + 1) = 9 \frac{(R+R_x K^3)(R^2 + K^3 R_x^2)}{(R^3 + K^3 R_x^3)}$$

(5.46)

Since $R_x \ll R$ then (5.45) and (5.46) respectively reduce to (5.23) and (5.24). Thus the deviation from ideal behavior caused due to practical model can be kept small if the external resistors are chosen to be much larger than $R_x$.

The nonideality analysis for Circuit I and Circuit III can be worked out in a similar manner.

5.2.3.5 Simulation and Experimental Results

The proposed circuits have been simulated using SPICE to validate the theoretical predictions, employing OTRA realization of Fig. 2.7. Simulation results of circuit I having $n = 3$ and component values $R_1 = 0.5$ KΩ, $R = 1$ KΩ and $C = 100$ pF are shown in Fig. 5.13(a). The simulated frequency of oscillation was 2.838 MHz against the calculated value of 2.757 MHz having frequency error of 2.93%. The simulated and theoretical frequencies of oscillation as a function of capacitance ($C$) are shown in Fig. 5.13(b). It shows that the simulated values are in close agreement with the ideal values.

Figure 5.14 (a) shows the simulation results of circuit II having $n = 4$ and component values $R_1 = 0.707$ KΩ, $R = 1$ KΩ and $C = 100$ pF. The simulated value achieved was 1.595 MHz against the theoretical value of 1.591 MHz with a frequency error of 0.25%. The simulated and theoretical frequencies of oscillation as a function of capacitance ($C$) are shown in Fig. 5.14(b) and both the curves are in close agreement.
Fig. 5.13 Simulation result for circuit I. (a) Output waveform for $n = 3$. (b) Frequency error curve.
Fig. 5.14 Simulation results for circuit II. (a) Output waveforms for $n = 4$.
(b) Frequency error curve.
The simulation results of circuit III having two phase shifter blocks along with the SRCO are depicted in Fig. 5.15(a). The design is obtained for an FO of 2.361 MHz with a phase shift of 45°. The component values chosen are $R_1 = R_3 = R_4 = R_5 = R_p = 1 \, \text{K} \Omega$, $R_2 = 2 \, \text{K} \Omega$, $R = 1.434 \, \text{K} \Omega$ and $C_1 = C_2 = C = 50 \, \text{pF}$.

Fig. 5.15 Simulation results for circuit III (a) Output waveforms having two phase shifter networks of 45° each. (b) Frequency error curve.
The observed FO was 2.220 MHz against the calculated value of 2.361 MHz. The simulated and theoretical frequencies of oscillation as a function of capacitance are shown in Fig. 5.15 (b).

The functionality of the proposed MSO circuits is verified through hardware also. The commercial IC AD844AN is used to implement an OTRA as discussed in section 2.8. Supply voltages used are ± 5 V. Figure 5.16(a) shows the experimental results for circuit II, having, n = 3 for component values $R_1 = 2.7$ KΩ, $R = 5.4$ KΩ, $R_3 = R_4 = R_5 = 1$ KΩ and $C = 3.3$ nF. Observed FO is around 15.7 KHz and is in close agreement to calculated FO of 15.469 KHz.

The output of circuit III consisting of SRCO along with two phase shifter blocks is depicted in Fig. 5.16 (b) for component values $R_1 = R_P = 2.7$ KΩ, $R_2 = 5.4$ KΩ, $R_4 = R_5 = 1$ KΩ, $R_3 = 250$ Ω, $R = 10$ KΩ and $C = C_1 = C_2 = 3.3$ nF. These component values result in theoretical FO as 43.4 KHz and the observed frequency is around 44.72 KHz. The slight variation in experimental values of FO, from phase to phase, as seen in Fig. 5.16 may be due to tolerance of the component values.

![Fig. 5.16 (a) Experimental result for circuit II for n = 3, (b) Experimental result for circuit III with two phase shifter blocks.](image)

**5.3 CONCLUDING REMARKS**

In this chapter OTRA based VM single, quadrature and multiphase sinusoidal oscillator configurations are presented. The proposed single and quadrature phase structures provide
non interactive control on FO and CO both and are made electronically tunable through MOS implemented resistors. Design description of three MSO circuits follows this sequence. The MSO circuit-I provides only odd phase oscillations whereas the remaining two structures provide both even and odd phase oscillations. These circuits are very accurate in producing a wide range of oscillation frequency with the desired phase shift. Hence, they are capable of replacing voltage mode op-amps based MSO as they are free from the limitations of conventional voltage mode op-amps.

SPICE simulations, are included to demonstrate the workability of all proposed oscillators. Experimental results for MSO circuits have also been included. The effect of non ideal behavior of OTRA in practice has been analyzed for all proposed structures.