CHAPTER IV

System Circuit Design

The system used for this application similar to the analog computer in which it increases every frequency of the input signal by an amount of the desired frequency. In this chapter the theory and numerical calculation of this system will be presented seperately in details. The last section of this chapter will be an illustration of this whole system.

4.1 Band Pass Constant Phase Shift Network

Because the input signals are in the audio frequency band therefore the constant phase shift network must be a band pass phase shift network. There are many methods to fulfill our requirement such as Dome's method [7] and Dickey's phase shift method [10] . In our application the method we chose is widely used in most of the audio system such as method used to shift phase in quadraphonic four-channels application. The circuit consists of two active lead-lag phase shift networks in parallel. The phase difference of each network varies as the logarithmic function of frequency, but when two networks are in parallel their output phase difference will be independent of frequency, see equation (4.13) The network which is described above is shown in Fig. 4.1

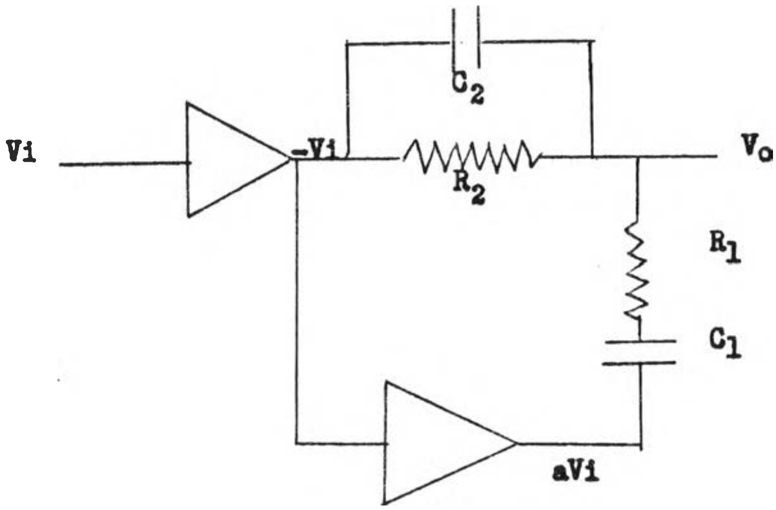

Fig. 4.1 Band Pass Phase Shift Network

Let V_i = Input signal voltage

V_o = Output signal voltage

a = Voltage gain of the operational amplifier

R = Resistance

Z = Impedance of condenser $= \frac{1}{jwC}$

From Fig. 4.1 we can write the equation

$$\frac{-V_i - V_o}{\frac{R_2 Z_2}{R_2 + Z_2}} = \frac{V_o - aV_i}{Z_1 + R_1} \qquad \ldots\ldots\ldots\ldots\ldots(4.1)$$

$$\frac{V_o - aV_i}{-(V_o + V_i)} = \frac{(Z_1 + R_1)\ (R_2 + Z_2)}{R_2 Z_2}$$

$$\frac{aV_i - V_o}{V_o + V_i} = \frac{\left(\frac{1}{jwC_1} + R_1\right)\left(R_2 + \frac{1}{jwC_2}\right)}{R_2 \ \frac{1}{jwC_2}}$$

$$1 + \frac{aV_i - V_o}{V_o + V_i} - 1 = \frac{(1 + R_1 jwC_1)\ (R_2 jwC_2 + 1)}{jwC_1 \quad jwC_2 R_2 \quad \frac{1}{jwC_2}}$$

$$\frac{V_o + V_i + aV_i - V_o}{V_o + V_i} - 1 = \frac{1 - R_1 R_2 w^2 C_1 C_2 + jw\ (R_2 C_2 + C_1 R_1)}{jwC_1 R_2}$$

$$\frac{V_i\ (1 + a)}{V_o + V_i} - 1 = j\ (wC_2 R_1 - \frac{1}{R_2 wC_1}) + \frac{C_2}{C_1} + \frac{R_1}{R_2}$$

$$\frac{V_i\ (1 + a)}{V_o + V_i} = j\ (wC_2 R_1 - \frac{1}{R_2 wC_1}) + \frac{C_2}{C_1} + \frac{R_1}{R_2} + 1 \quad \ldots\text{.}(4.2)$$

if we define

$$\beta = wC_2 R_1 - \frac{1}{R_2 wC_1} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (4.3)$$

$$\alpha = \frac{C_2}{C_1} + \frac{R_1}{R_2} + 1 \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (4.4)$$

then $$\frac{V_i\ (1 + a)}{V_o + V_i} = j\beta + \alpha$$

$$\frac{V_o + V_i}{V_i\ (1 + a)} = \frac{1}{j\beta + \alpha}$$

$$\frac{V_o}{V_i} + 1 = \frac{1 + a}{j\beta + \alpha}$$

$$\frac{V_o}{V_i} = \frac{1 + a}{j\beta + \alpha} - 1$$

$$\frac{V_o}{V_i} = \frac{1 + a - (j\beta + \alpha)}{j\beta + \alpha}$$

$$\frac{V_o}{V_i} = \frac{1 + a - (j\beta + \alpha)}{j\beta + \alpha}$$

if we define $a = 2\alpha - 1$ .... (4.5)

$$\frac{V_o}{V_i} = \frac{1 + 2\alpha - 1 - j\beta - \alpha}{j\beta + \alpha}$$

$$\frac{V_o}{V_i} = \frac{\alpha - j\beta}{\alpha + j\beta}$$

$$\left|\frac{V_o}{V_i}\right| = \sqrt{\frac{\alpha^2 + \beta^2}{\alpha^2 + \beta^2}} = 1$$

From derived equation ,the magnitude has shown to be a constant, so with this conclusion it is an all pass network.

From equation (4.2), the frequency f_o at which the phase shift is equal to zero is given by writing that the imaginary part of the equation is equal to zero.

then $\beta = 0$

$$w_0 C_2 R_1 - \frac{1}{R_2 w_0 C_1} = 0$$

$$w_0 C_2 R_1 = \frac{1}{R_2 w_0 C_1}$$

$$w_0^2 = \frac{1}{R_2 \; C_1 \; R_1 C_2}$$

if $R_1 C_1 = R_2 C_2$

then $$2 \pi f_o = w_o = \frac{1}{R_1 C_1} = \frac{1}{R_2 C_2} \quad \text{.........} \quad (4.6)$$

From equation (4.5), (4.4), (4.6), if $R_1 C_1 = R_2 C_2$

then $$a = 2 \alpha - 1$$

$$= 2 \left[\frac{C_2}{C_1} + \frac{R_1}{R_2} + 1 \right] - 1$$

$$= \frac{2C_2}{C_1} + \frac{2R_1}{R_2} + 2 - 1$$

$$= \frac{2C_2 R_2 + 2R_1 C_1}{R_2 C_1} + 1$$

$$= \frac{4 \; C_1 R_1}{C_1 R_2} + 1$$

$$= \frac{4R_1}{R_2} + 1 \quad \text{...............} \quad (4.7)$$

From equation (4.4) and (4.6), **if** $R_1C_1 = R_2C_2$

then

$$\alpha = \frac{C_2}{C_1} + \frac{R_1}{R_2} + 1$$

$$= \frac{R_1}{R_2} + \frac{R_1}{R_2} + 1$$

$$= \frac{2R_1}{R_2} + 1 \qquad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots (4.8)$$

From equation (4.3) and (4.6)

$$\beta = wC_2R_1 - \frac{1}{R_2wC_1}$$

$$= \frac{R_1w}{R_2w_o} - \frac{R_1w_o}{R_2w}$$

$$= \frac{R_1}{R_2}\left[\frac{w}{w_o} - \frac{w_o}{w}\right] \qquad \ldots\ldots\ldots\ldots (4.9)$$

then

$$\frac{\beta}{\alpha} = \frac{\frac{R_1}{R_2}\left[\frac{w_o}{w_o} - \frac{w_o}{w}\right]}{\frac{2R_1}{R_2} + 1}$$

$$= \frac{1}{\frac{R_2}{R_1} + 2} \cdot \left[\frac{w}{w_o} - \frac{w_o}{w}\right]$$

if we define

$$Q = \left[\frac{1}{\frac{R_2}{R_1} + 2}\right] \qquad \ldots\ldots\ldots\ldots\ldots\ldots (4.10)$$

From the transfer equation in page 17

$$\frac{V_o}{V_i} = \frac{\alpha - j\beta}{\alpha + j\beta}$$

$$= \frac{1 - j\frac{\beta}{\alpha}}{1 + j\frac{\beta}{\alpha}}$$

$$= \frac{1 - jQ\left[\frac{w}{w_o} - \frac{w_o}{w}\right]}{1 + jQ\left[\frac{w}{w_o} - \frac{w_o}{w}\right]}$$

then $$\frac{V_o}{V_i} = \frac{1 + jQ\left[\frac{w_o}{w} - \frac{w}{w_o}\right]}{1 - jQ\left[\frac{w_o}{w} - \frac{w}{w_o}\right]}$$

form Appendix C, the phase angle $\emptyset$ is given by

$$\tan\frac{\emptyset}{2} = Q\left[\frac{w_o}{w} - \frac{w}{w_o}\right]$$

or. $$\emptyset = 2\tan^{-1}\left[Q\left(\frac{fo}{f} - \frac{f}{fo}\right)\right] \quad \text{.........} \quad (4.11)$$

form [8] and Appendix A5-7, we can write Eq. (4.11) in approximation that

if $\frac{f_o}{4} \leq f \leq 4f_o$, $$\emptyset \approx -4\,Q \cdot \ln\frac{f}{f_o} \quad \text{..............} \quad (4.12)$$

For broad-band phase shift, we require two different sets of logarithmic-equation phase shift network in parallel as shown in Fig. 4.2

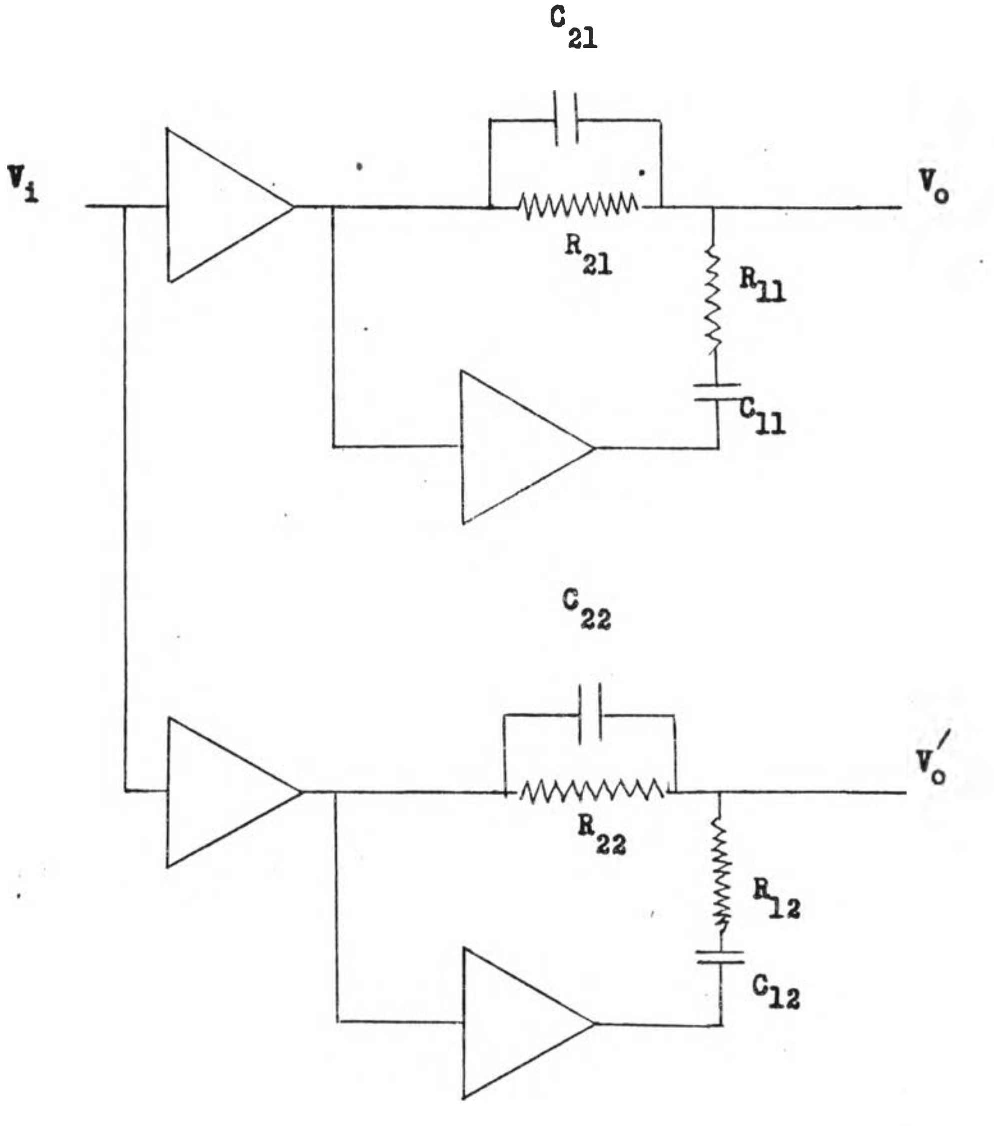

Fig. 4.2 Band Pass Constant Phase Shift Network.

The phase difference ψ between two outputs can be calculated from Eq. (4.12)

$$\psi = \phi_2 - \phi_1$$

$$= - 4Q \ln \frac{f}{f_{o2}} + 4Q \ln \frac{f}{f_{c1}}$$

$$= 4Q \left[\ln \frac{f}{f_{o1}} - \ln \frac{f}{f_{o2}} \right]$$

$$= 4Q \ln \left[\frac{f}{f_{o1}} \cdot \frac{f_{o2}}{f} \right]$$

$$= 4Q \ln \frac{f_{o2}}{f_{o1}} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (4.13)$$

$$= \text{constant}$$

but
$$f_{o1} = \frac{1}{2\pi R_{11} C_{11}} = \frac{1}{2\pi R_{21} C_{21}} \quad \ldots\ldots\ldots (4.14)$$

$$f_{o2} = \frac{1}{2\pi R_{12} C_{12}} = \frac{1}{2\pi R_{22} C_{22}} \quad \ldots\ldots\ldots (4.15)$$

then
$$\psi = + 4Q \ln \frac{R_{21} C_{21}}{R_{22} C_{22}} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (4.16)$$

In our application we have limited ourselves to an audio frequency range of 800 Hz to 4000 Hz which corresponds roughly to the frequency range of speeches.

then $$f_{o1} = \frac{1}{2\pi R_{21}C_{21}} \approx 800 \text{ Hz} \qquad (4.17)$$

$$f_{o2} = \frac{1}{2\pi R_{22}C_{22}} \approx 4000 \text{ Hz} \qquad (4.18)$$

the value for R_{21} from equation (4.17) must be in the range of hundred of kilo ohms in order to match the input inpedance of the operational amplifier, if f_{o1} is in the range of thousand hertz, the velue of C_{21} would be in the range of $0.001\ \mu F$

The same reasoning is also applied with Eq (4. 18) on finding a value for C_{22} and R_{22}

The values for our first trial were found to be

$$C_{21} = C_{22} = 0.001\ \mu F$$

$$R_{21} = 220\ k\Omega$$

$$R_{22} = 43\ k\Omega$$

from equation (4.16) since the required phase difference ψ is $90°$

$$\psi = \frac{\pi}{2} = +\ 4Q \ln \frac{220}{43}$$

or $$\frac{3.1416}{2} = +\ 4Q \times 1.632428$$

then $$Q = +\ 0.2405612 \qquad (4.19)$$

from equation (4.10) and (4.19) we can calculate R_{11} and R_{12},

$$Q = \frac{1}{\frac{R_{21}}{R_{11}} + 2} = \frac{1}{\frac{R_{22}}{R_{12}} + 2} = 0.2405$$

then $$\frac{R_{21}}{R_{11}} = \frac{R_{22}}{R_{12}} = 2.1569463$$

$$\frac{220k\Omega}{R_{11}} = \frac{43k\Omega}{R_{12}} = 2.1569463$$

$$R_{11} = 102 \quad k\Omega$$

$$R_{12} = 20 \quad k\Omega$$

from Eq (4.6) C_{12}, C_{11} can be calculated,

$$C_{12} = C_{11} = \frac{R_{22}}{R_{12}} \times C_{22}$$

$$= \frac{43}{20} \times 0.001 \quad \mu F$$

$$= 0.00215 \quad \mu F$$

from equation (4.7) the gain of the operational amplifier is

$$a = \frac{4R_{11}}{R_{12}} + 1$$

$$= \frac{4 \times 102}{220} + 1$$

$$= 2.85$$

then f_{01} and f_{02} for our first trial are:

$$f_{01} = \frac{1}{2\pi R_{11} C_{11}} = \frac{1}{2\pi R_{21} C_{21}} = 743 \text{ Hz}$$

$$f_{02} = \frac{1}{2\pi R_{22} C_{22}} = \frac{1}{2\pi R_{12} C_{12}} = 3700 \text{ Hz}$$

Because in our design we use the approximation that

$$\psi = 2 \tan^{-1} \left| Q \left(\frac{f_o}{f} - \frac{f}{f_o}\right) \right| \approx -4Q \ln \frac{f}{f_o}$$

The phase difference between the outputs of the two phase shift networks will differ somewhat from $\frac{\pi}{2}$ see Appendix A and Fig. 4.3. By trial and error, if we choose f_{o2} to be 3500 Hz instead of 3700 Hz, then the new parameters are as shown below:

$$f_{02} = \frac{1}{2\pi R_{22} C_{22}} = \frac{1}{2\pi R_{12} C_{12}} = 3500 \text{ Hz}$$

$R_{12} = 20 \text{ k}\Omega$, $C_{22} = 0.001 \ \mu\text{F}$

we can solve that $C_{12} = 0.0022 \ \mu\text{F}$

$R_{22} = 45 \text{ k}\Omega$

then $$Q_2 = \frac{1}{\frac{R_{22}}{R_{12}} + 2} = 0.2352$$

These parameters are adopted. The circuit diagram using these valuesare shown in fig 4.4. The photo of the real circuit is shown in fig 4.6. This circuit give a satisfactory result that is the phase difference is equal to $90° \pm 5°$ in the frequency range of 300 Hz to 9000 Hz (see fig. 4.3).

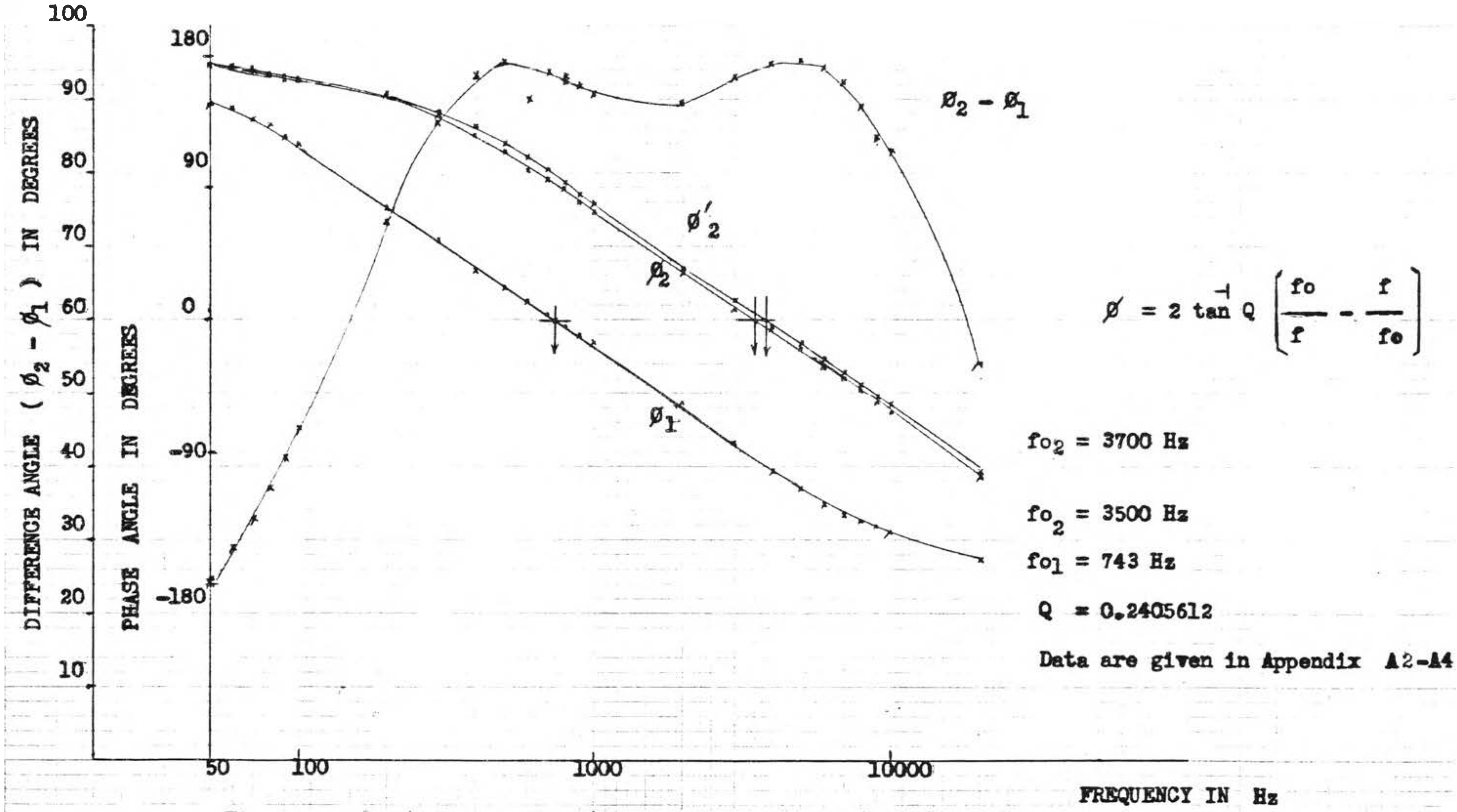

Fig. 4.3 Characteristic of Constant Phase Shift Network .

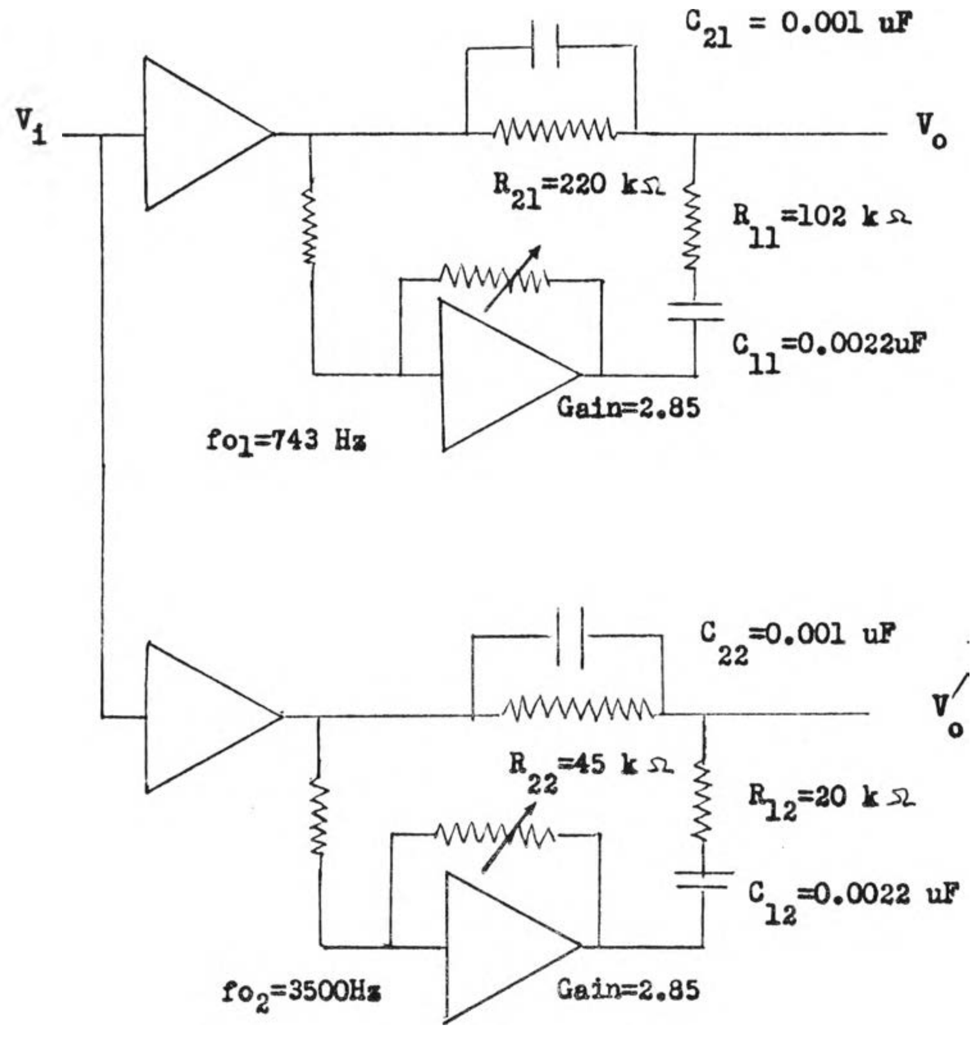

Fig. 4.4 Band Pass Phase Shift Circuit Diagram.

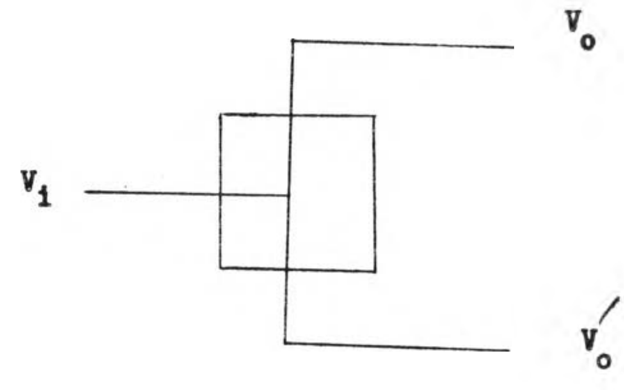

Fig. 4.5 Block Diagram of The Band Pass Phase Shift Circuit.

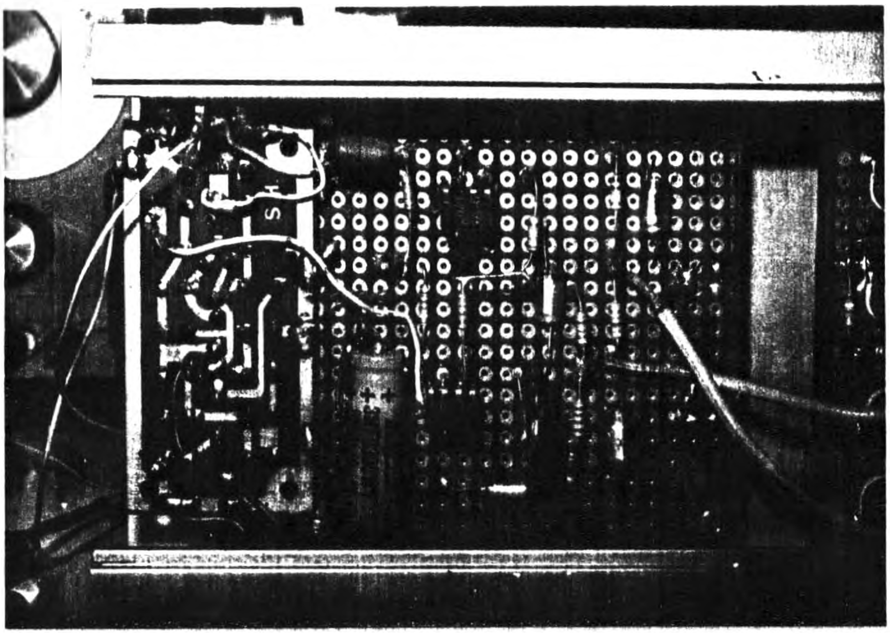

Fig. 4.6 Photo of The Band Pass Phase Shift Circuit.

4.2 Signal Summing Circuit

Two signalscan be added together by a simple operational amplifier as shown in the Fig 4.7

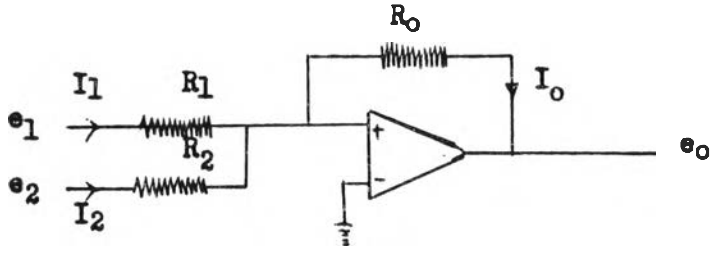

Fig 4.7 Signal Summing Circuit

Let e_1, e_2 = Input signal

e_o = Output signal

from Fig 4.7 we can write the equation that

$$e_o = -R_o I_o$$

$$I_o = I_1 + I_2$$

$$I_1 = \frac{e_1}{R_1}$$

$$I_2 = \frac{e_2}{R_2}$$

$$e_o = -R_o (I_1 + I_2)$$

$$= -R_o \left(\frac{e_1}{R_1} + \frac{e_2}{R_2} \right)$$

Then the output signal e_o is a function of the sum of input signals e_1 and e_2. In our practical circuit we use μA 741 as operational amplifier which has an internal frequency compensation with high input voltage range and excellent temperature stability.

Our practicle circuit is shown in Fig. 4.8

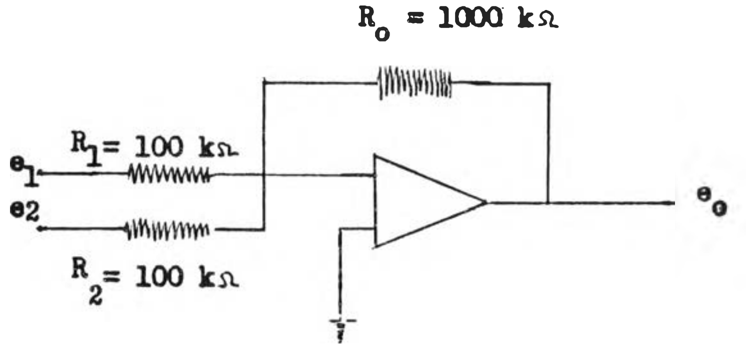

Fig. 4.8 Circuit Diagram of The SignalSumming Circuit

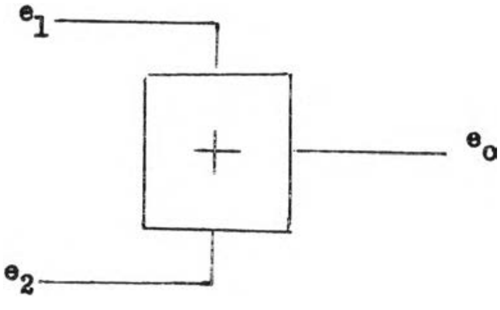

Fig. 4.9 Block Diagram of The Signal Summing Circuit

4.3 Quadrature Signal Generator with Nonlinear Amplitude Limiting

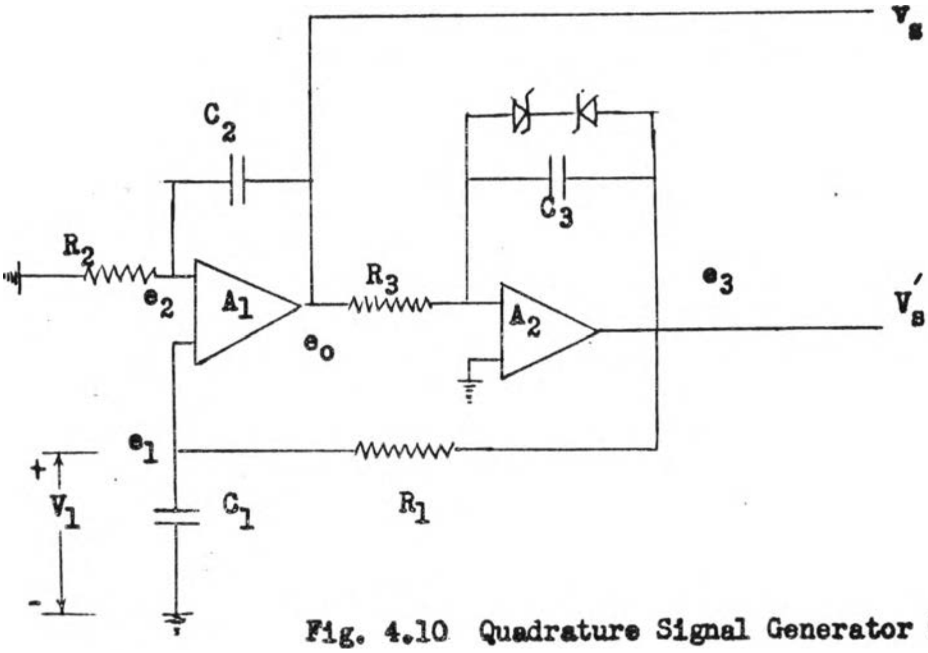

Fig. 4.10 Quadrature Signal Generator Circuit

A quadrature oscilator circuit which uses amplitude limiting is shown in figure 4.10, two operational amplifiers are used as integrators in this circuit, two zener diodes are used as an amplitude limiting devices. The behavior of the circuit is better understood if the nonlinear limiting on A_2 is first ignored, after the linear behavior is described the effect of the nonlinearity can be considered.

In considering the linear behavior, let us assume that there is an initial voltage of V_1 across capacitor C_1 and other initial conditions are zero.

Laplace transforms of voltage e_1, e_2 and e_3 are given by

$$E_1(s) = \frac{1}{R_1C_1s + 1} E_3(s) + \frac{V_1}{s}$$

$$E_2(s) = \frac{R_2C_2s}{R_2C_2s + 1} E_o(s)$$

$$E_3(s) = \frac{1}{R_3C_3\ s} E_o(s)$$

Assuming ideal operational aplifiers $E_1(s)$, $E_2(s)$ will be equal.

$$E_1(s) = E_2(s)$$

Defining $T_1 = R_1C_1$

$$T_2 = R_2C_2$$

$$T_3 = R_3C_3$$

then the output $E_o(s) = \dfrac{(s + \frac{1}{T_1})(s + \frac{1}{T_2}) V_1}{s^3 + (\frac{1}{T_1})s^2 + (\frac{1}{T_1T_3})s + \frac{1}{T_1T_2T_3}}$

if $T_1 = T_2$

$$E_o(s) = \frac{(s + \frac{1}{T_1})^2 V_1}{s^3 + (\frac{1}{T_1})s^2 + (\frac{1}{T_1T_3})s + \frac{1}{T_1T_1T_3}}$$

$$= \frac{(s + \frac{1}{T_1}) V_1}{(s^2 + \frac{1}{T_1T_3})}$$

The solution as a function of time is found by taking the inverse laplace transformation. Thus we obtain

$$e_o(t) = V_1\sqrt{\frac{T_3}{T_1} + 1}\ \sin\left(\frac{1}{\sqrt{T_1T_3}}\ t + \psi\right)$$

where $\psi = \tan^{-1} \frac{T_1}{\sqrt{T_1\ T_3}}$

if $T_1 = T_3$

then $e_o\ (t) = V_1.\sqrt{2}\ \sin\left(\frac{1}{T_1}\ t + 45°\right)$

and $e_3(t) = V_1\ .\ \sqrt{2}\ .\ \cos\left(\frac{1}{T_1} t + 45°\right)$

Then the output $e_o(t)$ and e_3 (t) have the same amplitude with 90° phase difference, in this case the frequency of the oscillator is

$$fo = \frac{1}{2\pi RC}$$

where $R = R_1 = R_2 = R_3$

$C = C_1 = C_2 = C_3$

In a practical circuit, slight mismatching of components will cause the circuit to slowly converge or diverge, if $R_1 C_1$ is deliberately made slightly greater than R_2C_2 the oscilator output amplitude will diverge [11].

But if limiters clip the output of A_2, the output amplitude will be stabilized. The output distortion will be roughly proportional to the degree of mismatch between R_1C_1 and R_2C_2 the distortion will generally be lower at the output e_o than at the output e_3. [11]

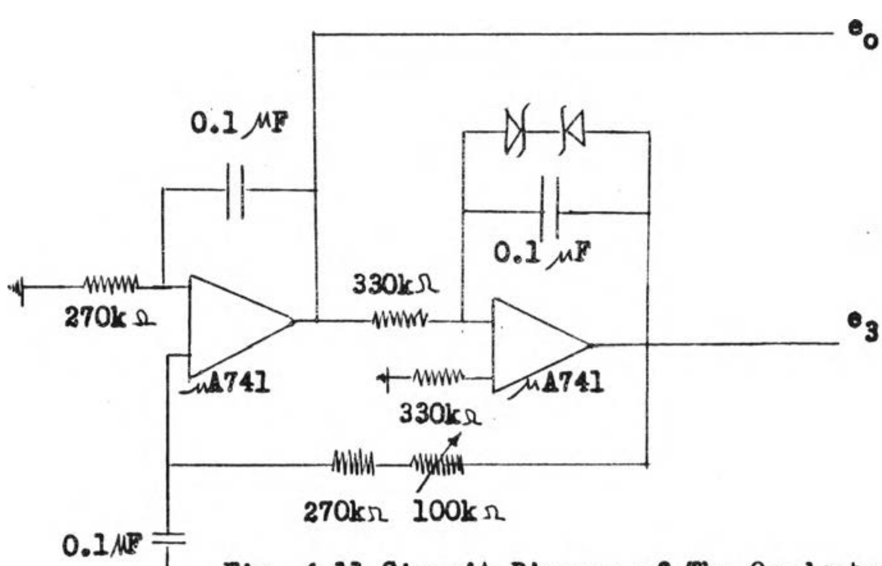

Fig. 4.11 Circuit Diagram of The Quadrature Signal Generator

Figure 4.11 shows the practical circuit using two μA741 operational amplifiers to generate two shifting frequency signals having 90 degrees phase difference between each other.

The frequency of the quadrature signal generator is between 4 Hz and 12 Hz [5], see page 11. For our application, we choose $f_s = 6$ Hz which corresponds to the following practical values

- of compoonents R=270 kΩ C= 0.1 μF as shown by the calculation :

$$f = \frac{1}{2\pi RC}$$

$$= \frac{1}{2.\pi .\ 270x\ 10^{3}.\ 0.1x\ 10^{-6}}$$

$$= 5.89 \text{ Hz}$$

The block diagram for the quadrature signal generator is shown in Fig. 4.12

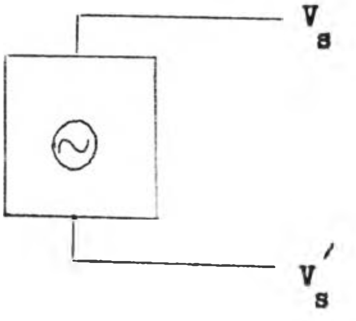

Fig. 4.12 Block Diagram of The Quadrature Signal Generator

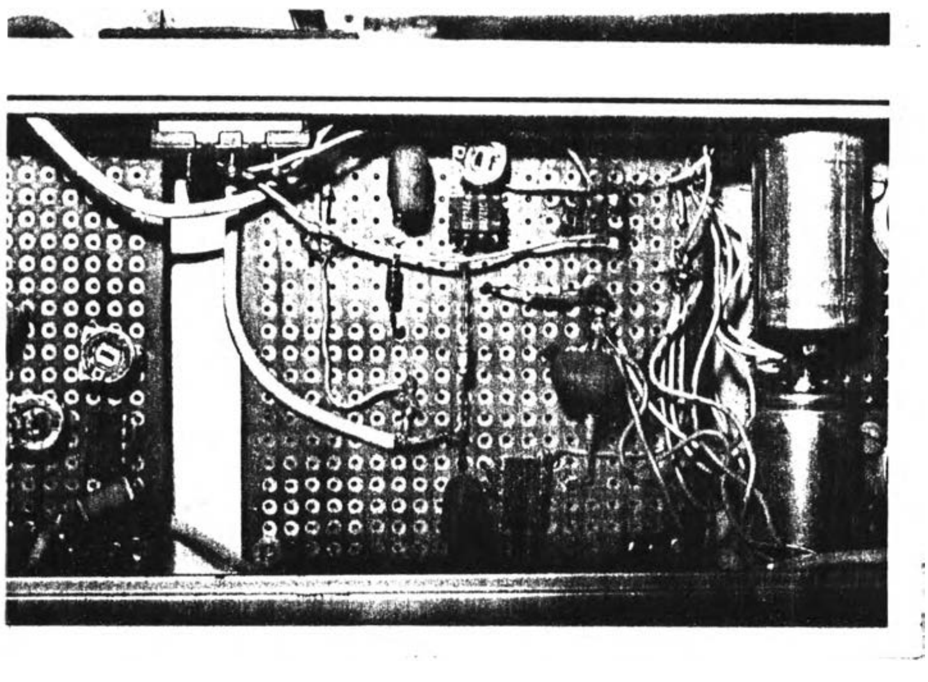

Fig. 4.13 Photo of The Quadrature Signal Generator

4.4 Analog Signal Multiplier

There are many methods in the application of operational amplifier for multiplication of analog signals. We will choose the current ratioing method, not only for its accuracy, speed and cost, but also for its excellent linearity and temperature stability. Fig 4.1 shows a circuit diagram of a current ratio multiplier.

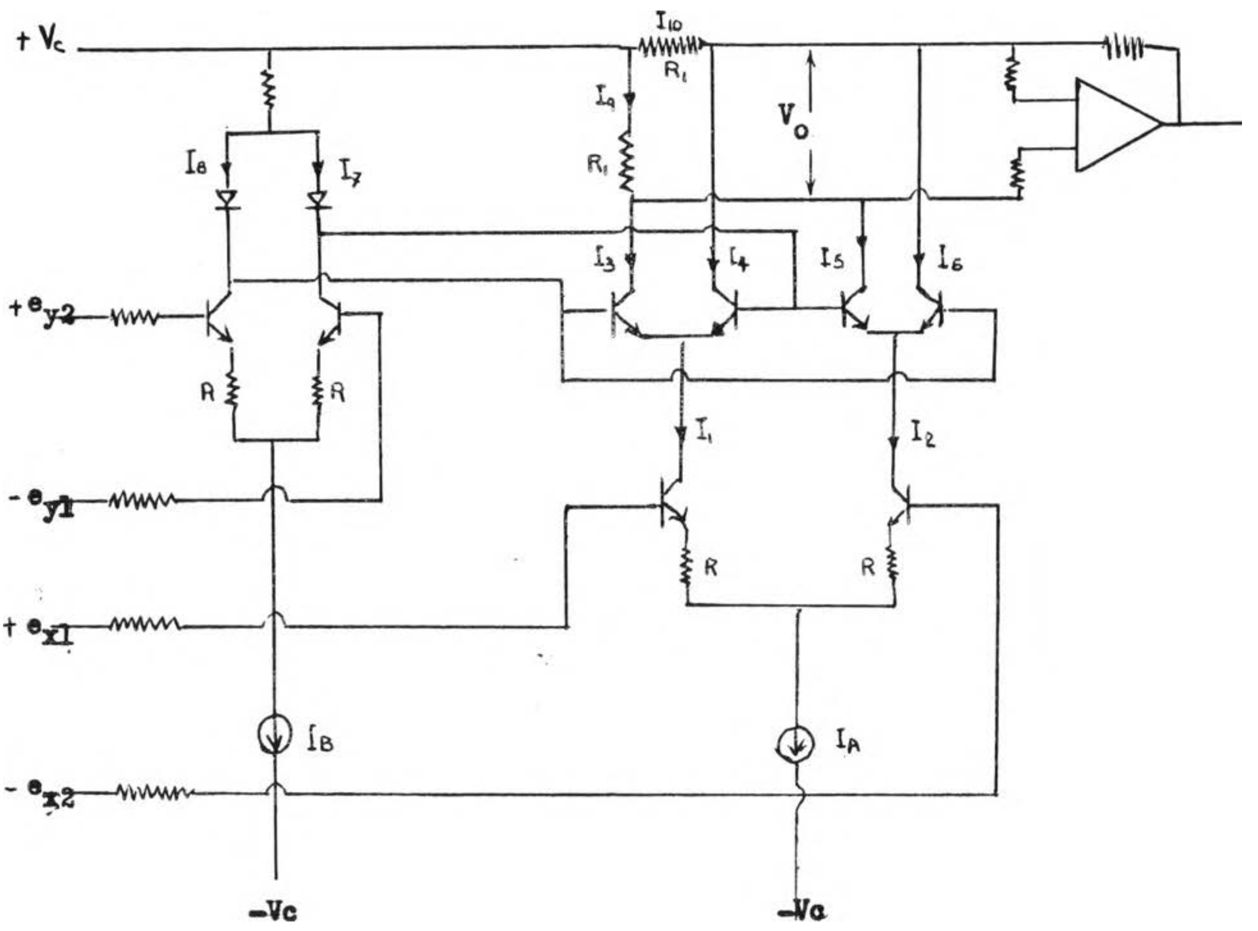

Fig. 4.14 Current Ratio Multiplier Circuit Diagram.

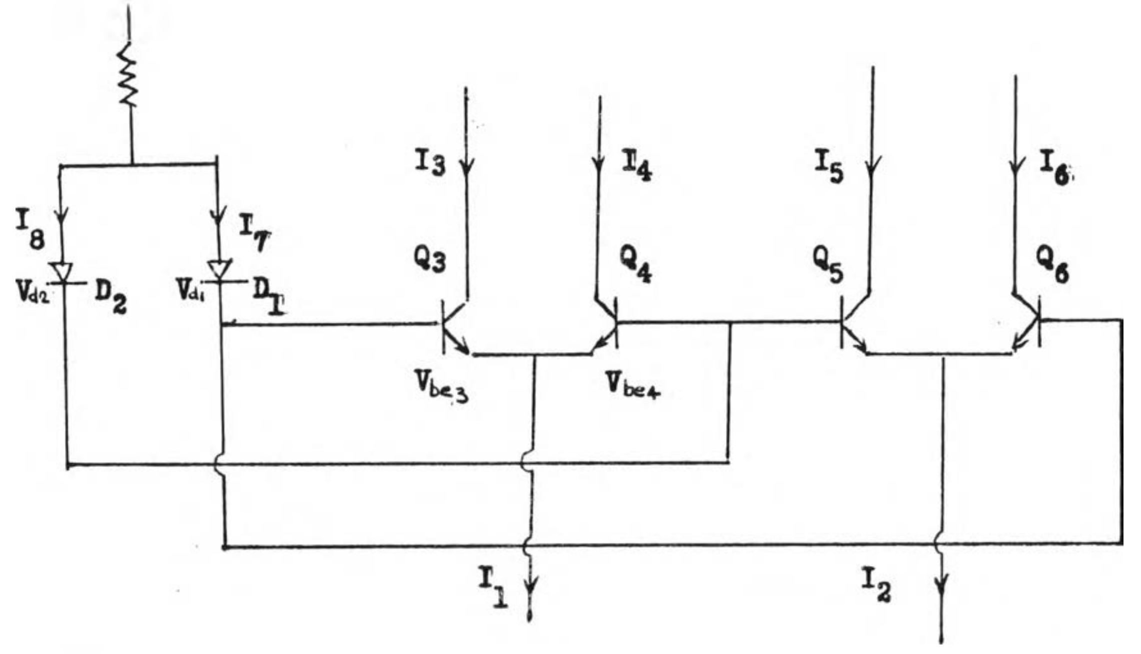

Fig. 4.15 Principle of The Multiplier Circuit Diagram

The main part of this multiplier circuit is shown in the figure 4.15 . In this circuit the currents I_3, I_4 in the collectors of transistors Q_3, Q_4 remain in a constant ratio equals to the ratio of the external current I_7 and I_8

The current I_1, I_7 and I_8 are generated by constant current sources. The currents and voltages of the circuit are related by the following equations (10)

$$I_7 = K_1 \, e^{\lambda_1 V_{d1}}$$

$$I_8 = K_2 \, e^{\lambda_1 V_{d2}}$$

$$I_3 = K_3 \, e^{\lambda_3 V_{be3}}$$

$$I_4 = K_4 \, e^{\lambda_4 V_{be4}}$$

where K_1, K_2, K_3, K_4 are the current when voltage across junction=0

$$\lambda = \frac{q}{k.T^{\circ}} \qquad \text{------------------------(4.20)}$$

q = Electronic charge

k = Boltz mann's constant

T° = Absolute temperature

and V_{be} = Voltage across base and emitter of transistor

V_d = Voltage across diode

If the transistors and diodes are matched to made

$$\lambda = \lambda_1 = \lambda_2 = \lambda_3 = \lambda_4$$

$$K_1 = K_2 = K_3 = K_4$$

then

$$\frac{I_7}{I_8} = \frac{K_1 e^{\lambda_1 V_{d1}}}{K_2 e^{\lambda_2 V_{d2}}}$$

$$\frac{I_7}{I_8} = e^{\lambda (V_{d1} - V_{d2})} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots (4.21)$$

and

$$\frac{I_4}{I_5} = e^{\lambda (V_{be4} - V_{be3})} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots (4.22)$$

The loop equation can be written

$$V_{d1} + V_{be3} = V_{be4} + V_{d2}$$

$$V_{d1} - V_{d2} = V_{be4} - V_{be3} \quad \ldots\ldots\ldots\ldots (4.23)$$

By substituing equation (4.23) into the equation (4.21) , (4.22).

then

$$\frac{I_7}{I_8} = \frac{I_4}{I_3}$$

similarly

$$\frac{I_7}{I_8} = \frac{I_5}{I_6}$$

then

$$\frac{I_7}{I_8} = \frac{I_4}{I_3} = \frac{I_5}{I_6} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots (4.24)$$

From Fig. **4.14**

$$I_1 = I_3 + I_4$$

$$I_2 = I_5 + I_6$$

$$I_9 = I_3 + I_5$$

$$I_{10} = I_6 + I_4$$

$$I_A = I_1 + I_2$$

$$I_B = I_7 + I_8$$

then

$$\frac{I_5}{I_4} = \frac{I_6}{I_3} = \frac{I_2}{I_1} \qquad \text{..............(4.25)}$$

$$I_7 = \frac{I_4}{I_1} \cdot I_B \qquad \text{..............(4.26)}$$

$$I_8 = \frac{I_3}{I_1} \cdot I_B \qquad \text{..................(4.27)}$$

from Fig 4.14

$$e_x = R(I_1 - I_2) \qquad \text{................(4.28)}$$

$$e_y = R(I_8 - I_7) \qquad \text{..............(4.29)}$$

$$V_o = R_1 (I_9 - I_{10})$$

from Eq.(4.24),(4.25)

$$= R_1 (I_1 - I_2) . \frac{(I_3 - I_4)}{I_1}$$

then from Eqs.(4.26), (4.27), (4.28),(4.29) we have

$$V_o = R_1 . \frac{e_x}{R} . \frac{e_y}{R. I_B}$$

$$= \frac{R_1}{R^2 . I_B} . (e_x . e_y)$$

Because R_1 and R are constant, I_B is a constant current source, so V_0 is proportional to e_x multiplied by e_y.

The above current ratioing multiplication method will function accurately if the transistors and diodes are dynamically matched, that requirement makes monolithic construction attractive for this type of multiplier. Then in our study we used a monolithic four-quadrant multiplier chip MC 1495 which has an excellent linearity 1% error on X input, 2% error on Y input which has a circuit diagram almost the same as that discussed above except for the addition of constant current sources I_A and I_B .

The practicle circuit and block diagram are shown in Fig. 4. 16 and Fig. 4.17

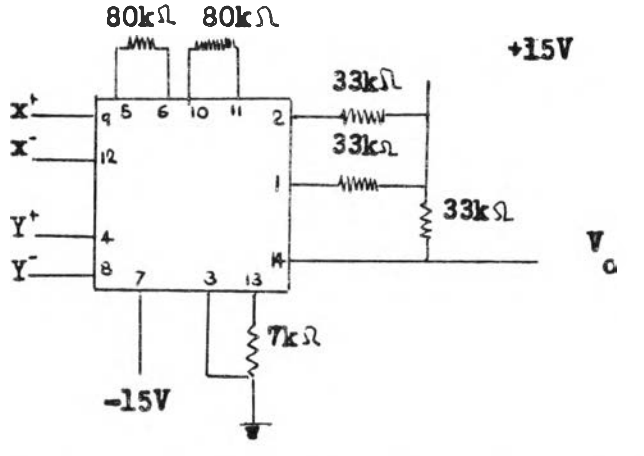

Fig. 4.16 Practical Circuit of MC 1495.

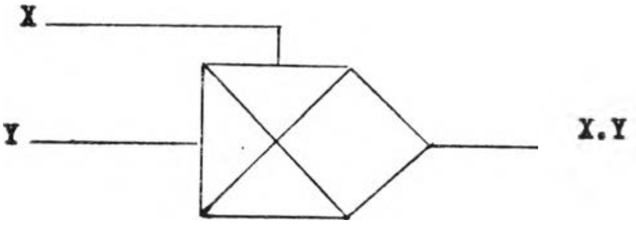

Fig. 4.17 Block Diagram of The Multiplier Circuit.

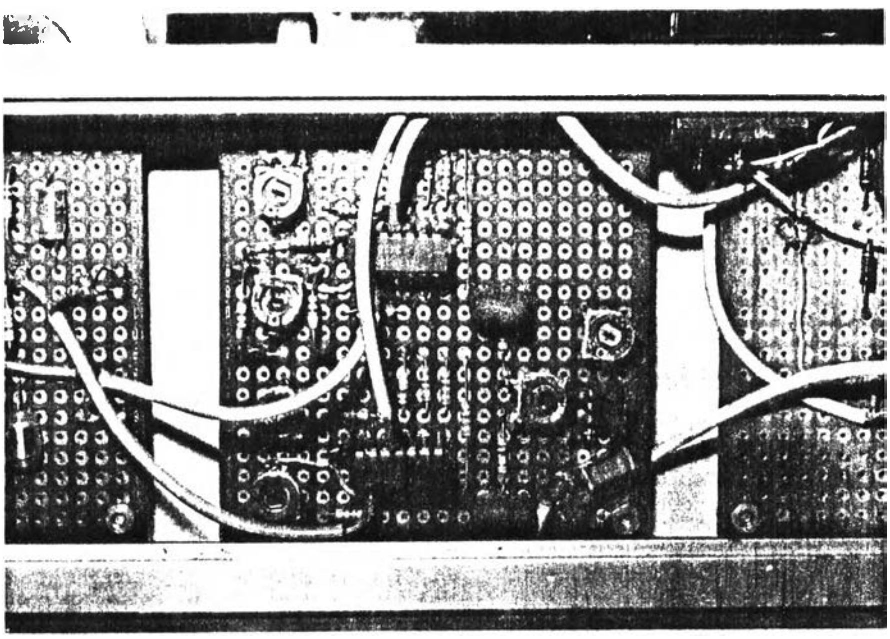

Fig. 4.18 Photo of The Multiplier Circuit.

4.5 The over all system.

The over–all system that can shift the input signal frequency by an amount of Δf which is generated by a quadrature signal generator are shown below in Fig. 4.19 and Fig. 4.20

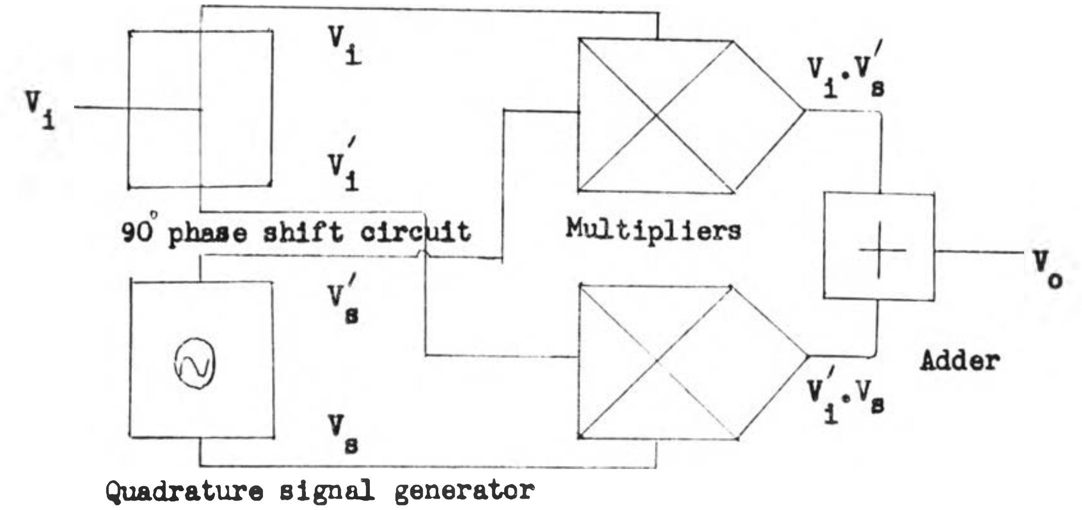

Fig. 4.19 Block Diagram of Frequency Shifting Circuit

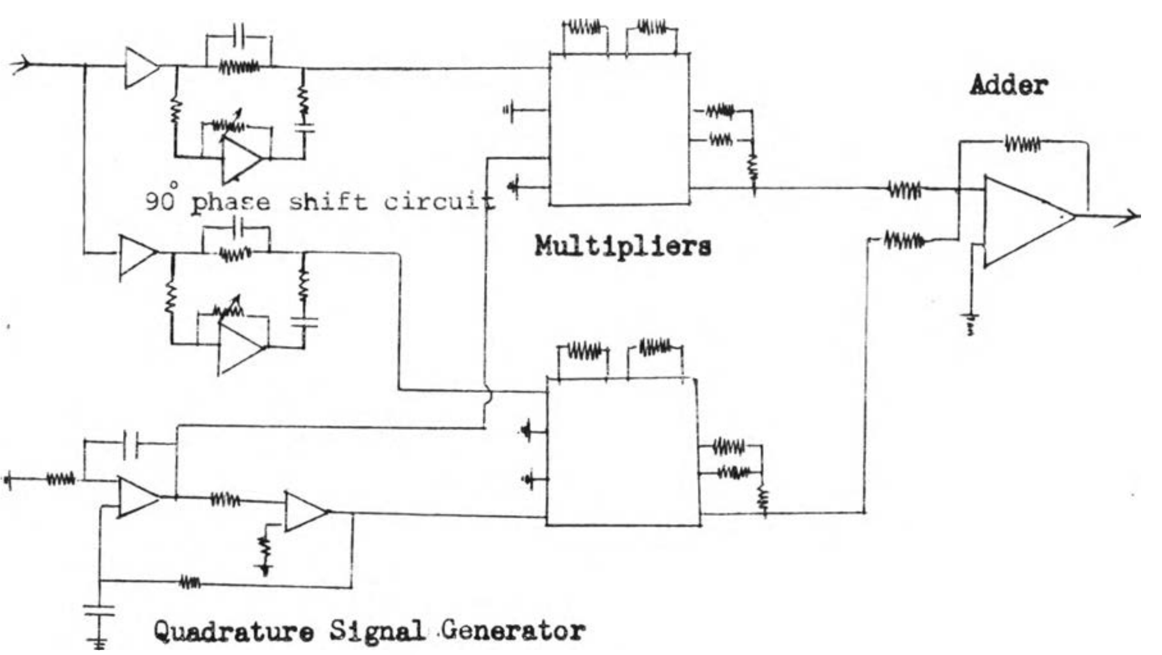

Fig. 4.20 Circuit Diagram of Frequency Shifting Circuit