CHAPTER - VIII

TWO SPECIAL OSCILLATORS

8.1 INTRODUCTION

The conventional phase-shift oscillator has a minimum of three RC sections in cascade with a phase inverting amplifier. An additional phase shifting amplifier can reduce the gain condition and decrease the number of RC sections to two. The frequencies obtainable in these oscillators are strictly dependent on the RC elements and therefore the lowest frequency cannot be made very small without making an unduly large RC product. An oscillatory system consisting of an RC differentiator, a new type of an RC network and two transactors in cascade, can, however, be designed whose frequency of oscillation is variable from a very low value covering a wide range by the variation of the ganged condensers in the RC networks - the R and C having reasonable values throughout the range of frequencies. The system is described in section 8.2.

One of the transactors is a voltage controlled voltage source (VCVS) and the other a voltage controlled current source (VCIS). The system is found to generate sinusoidal oscillation when the product of the gain and the transconductance of the controlled sources has a specific value determined by the circuit elements of the passive networks. Expression for the frequency of oscillation is derived from the chain matrix analysis and experimental support to the analysis has been made.

Section 8.3 deals with an RC oscillator using a modified form of the Wien network. The voltage transmittance form of the Wien network has been in commercial use since long. Lately its current dual has been used in a transistorized selective amplifier and an oscillator. A selective amplifier circuit realized with the admittance equivalent form of this network has also
appeared in the literature. But the so-called admittance equivalent form is basically a current transmittance type when selective amplification is derived in the circuit. A transistorized circuit incorporating this network has been analyzed with the help of the equivalent circuit and the circuit determinants.

The circuit is found to be readily usable both as a selective amplifier and an oscillator. One of the passive elements in the active circuit is kept adjustable and by varying the element the selectivity of the circuit can be varied when the circuit is used as an amplifier. Beyond a critical value of this element the system is found to generate sinewave oscillation the frequency of which is very close to the resonance frequency of the filter network. Conditions of oscillation are determined from the analysis simultaneously determining the frequency of oscillation and the critical value of the adjustable element used. Once this critical value is crossed the variation of the frequency of oscillation with the variation of the adjustable element becomes very small.

8.2 A SINEWAVE OSCILLATOR

8.2.1 The system

The system is shown in schematic block diagram in fig.8.1, where \( T_1 \) and \( T_2 \) are two transactors and \( N_1 \) and \( N_2 \) are the RC networks. Network \( N_2 \) is an RC differentiator (fig.8.3) and \( N_1 \) is a network of the configuration shown in fig.8.2. Transactor \( T_2 \) is a simple adjustable positive gain voltage amplifier whereas \( T_1 \) is a voltage to current source. The components are cascaded as shown in fig.8.1 and the output of the differentiator is feedback to the transactor \( T_1 \).

8.2.2 Analysis

The chain matrices of the networks \( N_1 \) (fig.8.2) and \( N_2 \) (fig.8.3) and
FIG. 8-1 SYSTEM BLOCK DIAGRAM.

FIG. 8-2. R.C. NETWORK $N_1$

FIG. 8-3. R.C. DIFFERENTIATOR $N_2$

FIG. 8-4. OPEN LOOP EQUIVALENT SYSTEM SHOWN WITH THE CHAIN MATRIX
of the transactor $T_1$ and $T_2$ are given respectively by

$$[a_{N_1}] = \begin{bmatrix} 1 & 0 \\ Y_1 & 1 \end{bmatrix} \quad \text{(8.1)}$$

$$[a_{N_2}] = \begin{bmatrix} 1 + Y_2 Z_2 & Z_2 \\ Y_2 & 1 \end{bmatrix} \quad \text{(8.2)}$$

$$[a_{T_1}] = \begin{bmatrix} 0 & \frac{1}{g_m} \\ 0 & 0 \end{bmatrix} \quad \text{(8.3)}$$

and

$$[a_{T_2}] = \begin{bmatrix} \frac{1}{A} & 0 \\ 0 & 0 \end{bmatrix} \quad \text{(8.4)}$$

where,

$$Y_1 = \frac{p^2 c_{o} c_{1} R_1 + p(c_{o}^2 + c_{1})}{p c_{1} R_1 + 1} \quad \text{(8.5)}$$

$$Y_2 = \frac{1}{R_2} \quad \text{(8.6)}$$

$$Z_2 = \frac{1}{pc_2} \quad \text{(8.7)}$$

$q_m$ is the transconductance of $T_1$ and $A$ is the gain of the voltage amplifier.

If $T_1$, $N_1$, $T_2$ and $N_2$ are cascaded as shown in fig.8.1, and the feedback loop is open, the overall chain matrix is given by

$$[a] = \begin{bmatrix} 0 & \frac{1}{g_m} & 1 & 0 & \frac{1}{A} & 0 & 1 + Y_2 Z_2 & Z_2 \\ 0 & 0 & Y_1 & 1 & 0 & 0 & Y_2 & 1 \end{bmatrix} \quad \text{(8.8)}$$

$$= \begin{bmatrix} \frac{1 + Y_2 Z_2}{A q_m} Y_1 & Y_1 Z_2 \\ 0 & 0 \end{bmatrix} \quad \text{(8.9)}$$
The equivalent system in the open loop condition is represented in fig. 8.4, and the transmission equations are given by

\[
\begin{bmatrix}
V_1 \\
I_1
\end{bmatrix} = \begin{bmatrix}
a_{11} & a_{12} \\
2a_{21} & a_{22}
\end{bmatrix} \begin{bmatrix}
Y_1 \\
I_1
\end{bmatrix}
\]

... (8.10)

Under the closed loop conditions, the relation

\[
a_{11} + a_{22} = \det \begin{bmatrix} a \end{bmatrix} = 1 \quad \text{... (8.11)}
\]

holds for oscillation. Combining relation (8.9) with eqn. (8.12), one gets

\[
\begin{bmatrix}
V_2 \\
I_2
\end{bmatrix} = \begin{bmatrix}
\frac{1 + \gamma_2 Z_2}{A g_m} Y_1 & \frac{\gamma_1 Z_2}{A g_m} \\
0 & 0
\end{bmatrix} \begin{bmatrix}
V_2 \\
I_2
\end{bmatrix}
\]

... (8.11)

Substituting \(Y\)'s and \(Z\)'s from eqns. (8.5) through (8.7) in eqn. (8.13) and rearranging, the following equation is derived:

\[
p^2 \gamma_0 C_1 C_2 R_1 R_2 + p \left[ \gamma_0 C_1 R_1 + C_2 R_2 (C_0 + C_1) - A g_m C_1 C_2 R_1 R_2 \right]
+ \left[ C_0 + C_1 - A g_m C_2 R_2 \right] = 0 \quad \text{... (8.14)}
\]

Equation (8.14) shows that a sinewave oscillation occurs in the system for the following conditions:

\[
C_0 C_1 R_1 + C_2 R_2 (C_0 + C_1) - A g_m C_1 C_2 R_1 R_2 = 0 \quad \text{... (8.15)}
\]

and

\[
C_0 + C_1 > A g_m C_2 R_2 \quad \text{... (8.16)}
\]

Using the value of \(A g_m\) from eqn. (8.15) in eqn. (8.14), one obtains after simplification

\[
p^2 \gamma_0 C_1 C_2 R_1 R_2 + \left[ C_1 - (C_0 + C_1) C_2 R_2 / (C_1 R_1) \right] = 0 \quad \text{... (8.17)}
\]
and condition (8.16) becomes

\[ C_1 > (C_0 + C_1) \frac{C_2 R_2}{C_1 R_1} \]  \hspace{1cm} ...(8.18)

This condition can always be satisfied by proper choice of the RC elements in the networks \( N_1 \) and \( N_2 \).

The frequency of oscillation, under this condition, is obtained from eqn. (8.17)

\[ \omega = \sqrt{C_1 - \frac{(C_0 + C_1)C_2 R_2}{C_1 R_1}} \]  \hspace{1cm} ...(8.19)

Relation (8.19) can be simplified if the following arbitrary choice is made without violation of the condition (8.18)

\[ C_0 = C_1 = C \]  \hspace{1cm} \hspace{2cm} ...(8.20)
\[ R_1 = R_2 = R \]

Equation (8.19) is, therefore, modified to

\[ \omega = \frac{\sqrt{C - 2}}{C R} \]  \hspace{1cm} ...(8.21)

In consequence of assumption (8.20), eqn. (8.15) is also modified to yield the value of the product of \( A \) and \( g_m \). Thus

\[ A g_m = \frac{2}{R} + \frac{C}{C_2 R} \]  \hspace{1cm} ...(8.22)

Since it is convenient to keep the transactors transmittances constant, any change in the value of \( C \) to obtain a variation in the frequency of oscillation should be associated with a change in the value of \( C_2 \) in the same ratio so that \( A g_m \) remains constant. For convenience \( R \) has been assumed constant. The value of \( \frac{\frac{C}{C_2}}{} \), however, determines the frequency range for the limits of the value of \( C \).
The lowest value of \( \frac{C_1}{C_2} \) is obtained from condition (8.18), which under assumption (8.20) is written as

\[
\frac{C_1}{C_2} > 2 \quad \ldots (8.23)
\]

As \( \frac{C_1}{C_2} \) is held constant the frequency relation (eqn.8.21) can be put in a more convenient form

\[
\omega = \frac{K}{CR} \quad \ldots (8.24)
\]

where

\[
K = \sqrt{\frac{C_1}{C_2}} - 2 \quad \ldots (8.25)
\]

It can be seen from eqn.(8.24) that without having to choose a large value for \( C \) or \( R \), \( \omega \) can be made very small if \( \frac{C_1}{C_2} \) is very close to 2 satisfying the inequality (8.23). The three capacitors used in the networks \( N_1 \) and \( N_2 \) may be ganged properly to realize the condition.

Graphs showing the relation between frequency and \( C \) for different values of \( K \) are shown in fig.8.5 when \( R \) is taken as 4.7 k.ohm.

8.2.3 Circuit and Results

The scheme of the experimental circuit is given in fig.8.6. The transactors \( T_1 \) and \( T_2 \) are shown in the equivalent form. \( T_2 \) is a standard high positive gain voltage amplifier, when \( T_1 \) is simulated with a common base configuration of transistor circuit amplifiers. The circuit for \( T_1 \) along with the appropriate isolation stage is shown in fig.8.7. The three condensers in \( N_1 \) and \( N_2 \) are ganged and the resistances are identical with a value 4.7 k.ohm. The gain of the voltage amplifier is kept adjustable for realizing the condition given by eqn.(8.22) for the obtained value of \( g_m \) of the transactor \( T_1 \). Experimental plot for \( K = 0.07 \) is shown in fig.8.5 in dashed line.
FIG. 8-5. FREQUENCY VERSUS CAPACITANCE PLOTS

FIG. 8-6. THE SCHEMATIC CIRCUIT OF THE OSCILLATOR
8.2.4 Remarks -

The sinewave oscillatory system with a large frequency ratio described above has variable frequency ranges which are easily realized by properly ganged condensers. The system produces low frequency sine wave oscillation without requiring unduly large capacitance and resistance values. The shift in the experimental curve is due to the difficulty in realizing the ideal circuit conditions with the available resources especially for that of the voltage controlled current source.

8.3 AN RC SELECTIVE AMPLIFIER AND AN OSCILLATOR

8.3.1 The filter network -

The passive filter network under consideration is shown in fig.8.8. This is the admittance equivalent form of the Wien network. When incorporated in the transistorized active network coupling the unbypassed resistors of the two emitters it acts as a feedback network. For the purpose of analysis the equivalent \( \pi \) - form of the \( T \)-filter is considered for convenience. This conversion is shown in fig.8.9, where,

\[
Y_A = \frac{pCR}{R(p^2C^2R^2 + 3pCR + 1)} \quad \ldots(8.26)
\]

\[
Y_B = \frac{pCR(1 + pCR)}{R(p^2C^2R^2 + 3pCR + 1)} \quad \ldots(8.27)
\]

and

\[
Y_C = \frac{p^2CR}{R(p^2C^2R^2 + 3pCR + 1)} \quad \ldots(8.28)
\]
FIG. 3-7. THE SCHEMATIC CIRCUIT OF THE TRANSACTOR T1.

FIG. 3-8. ADMITTANCE EQUIVALENT FORM OF THE WIEN NETWORK.

FIG. 3-9. T TO π CONVERSION OF THE FILTER.
The circuit of the system is shown in fig. 8.10, where the equivalent \( \pi \) - form replaces the actual filter network. The filter couples the unbypassed parts of the emitter resistors and the positive feedback is thus obtained. The element \( R_{E2} \), the unbypassed resistor in the emitter of the second transistor \( T_2 \), is kept variable. Adjusting its value the degree of feedback and consequently the Q-factor is altered. Subsequent analysis shows that the filter faces a low admittance source in this resistance in the operating frequency range of the circuit when used as an amplifier. This becomes a zero admittance source when oscillation starts in the system. Considering the low frequency \( y \)-parameter model of the transistors the ac equivalent circuit of the system is shown in fig. 8.11, where impedances of fig. 8.10 appear in the admittance forms and the following simplifications have been made:

\[
\begin{align*}
y_{E1} + y_C &= y_x \quad \ldots(8.29) \\
y_{E2} + y_B &= y_y \quad \ldots(8.30) \\
y_A &= y_z \quad \ldots(8.31)
\end{align*}
\]

It has been further assumed without loss of generality that the transistors are nearly identical so that

\[
\begin{align*}
y_{1e1} &\approx y_{1e2} = y_{1e} \quad \ldots(8.32) \\
y_{fe1} &\approx y_{fe2} = y_{fe} \quad \ldots(8.33)
\end{align*}
\]

Condition (8.33) is almost always true and condition (8.32) can be satisfied by choosing appropriate transistor pair.

The circuit can be solved with the help of the following set of equations derived from the nodal analysis of the circuit of fig. 8.11:
FIG. 8-10 THE CIRCUIT OF THE SYSTEM

FIG. 8-11. THE AC EQUIVALENT CIRCUIT
The admittance parameter circuit determinant is obtained from eqns. (8.34) through (8.38) as

\[
\Delta = \begin{vmatrix}
(y_s + y_{ie}) & -y_{ie} & 0 & 0 & 0 \\
-(y_{ie} + y_{fe}) & (y_{ie} + y_{fe} + y_x + y_z) & 0 & -y_z \\
y_{fe} & -y_{fe} & y_{L2} + y_{ie} & -y_{ie} & 0 \\
0 & -y_z & -(y_{ie} + y_{fe}) & (y_{ie} + y_{fe} + y_y + y_z) & 0 \\
0 & 0 & y_{fe} & -y_{fe} & y_{L2} \\
\end{vmatrix}
\]  

...(8.39)

8.3.3 The selective amplifier

The overall circuit gain considered at the emitter of the transistor \( T_2 \) is given by (8.38)

\[
A = \frac{Y_s}{\Delta} \triangleleft_{14} 
\]  

...(8.40)

where \( \Delta \) is the circuit determinant given by relation (8.39) and \( \Delta_{14} \) is the determinant obtained after deleting the first row and fourth column of the circuit determinant. In terms of the admittance parameters eqn. (8.40) can be written as:
Now as $R_{E2}$ and in consequence $Y_{E2}$ is variable while all other circuit parameters are assumed constant, gain-frequency relation can appropriately be derived for different values of $Y_{E2}$ till for a certain value of $Y_{E2}$ the system starts oscillating. Using eqns. (8.26) through (8.31) and writing

$$\omega CR = \frac{\omega}{\omega_0} = x = \text{normalized angular frequency} \quad \ldots(8.42)$$

eqn. (8.41) can be written as

$$A = \frac{a_4x^4 - a_2x^2 + a - j(a_3x^3 - a_1x)}{(y_{E2}b_4 + c_4)x^4 - (y_{E2}b_2 + c_2)x^2(y_{E2}b_0 + c_0) - j(y_{E2}b_3c_3)x^3 - (y_{E2}b_1 + c_1)x) \quad \ldots(8.43)$$

where $a$'s, $b$'s and $c$'s are functions of the constant parameters $\gamma_{1e}$, $\gamma_{fe}$, $\gamma_s$, $\gamma_L$, and $\gamma_{E1}$ only. Knowing the values of these functions eqn. (8.43) can be solved to obtain the gain-normalized frequency $(A - x)$ curves for different values of $Y_{E2}$. The curves are shown in fig. 8.12. As $Y_{E2}$ decreases the selectivity increases. The $Y_{E2} - Q$ curve is shown in fig. 8.13. The phase-frequency curve drawn for a single value of $Y_{E2}$ (fig. 8.14) is seen to be quite steep around zero phase condition, and the circuit shows potential characteristic of a stable oscillator. An interesting feature is noted from the gain-frequency curves is that with decreasing $Y_{E2}$ frequency at the peak gain also decreases although the decrease is quite small.

8.3.4 The oscillator

As $Y_{E2}$ is decreased below a critical value the system starts oscillating generating sinewave oscillations. This can be derived from eqn. (8.41) or (8.42) when the overall gain of the system equals $\infty$. Equation (8.40) indicates that
The gain normalized frequency curves of the amplifier for different values of \( y_2 \).
this condition also means that the circuit determinant should be zero. Thus the condition of oscillation is written as

\[ \Delta = \nabla_{\text{Re}} + \nabla_{\text{Im}} = 0 \quad \cdots \text{(8.44)} \]

where \( \nabla_{\text{Re}} \) and \( \nabla_{\text{Im}} \) represent the real and imaginary parts of the determinant \( \Delta \). Equating the imaginary part to zero the frequency condition is obtained as

\[ x^2 = \frac{A + B \gamma_{E2}}{C + D \gamma_{E2}} \quad \cdots \text{(8.45)} \]

where \( A, B, C \) and \( D \) are dependent only on \( \gamma_{le}, \gamma_{fe}, \gamma_{le}', \gamma_{L1} \) and \( \gamma_{El} \), and are constant. Thus \( x^2 \) is a bilinear function of the parameter \( \gamma_{E2} \). This bilinearity is clearly indicative of the relation between \( \gamma_{E2} \) and \( x^2 \). However the constant circuit parameters may be chosen such that with the change in the value of \( \gamma_{E2} \), \( x \) does not materially shift from unity and the choice becomes automatic for sustained oscillation. The variation of \( x \) with \( \gamma_{E2} \) is plotted in fig. 8.15.

Using this value of \( x \) (eqn. (8.40)) in \( \nabla_{\text{Re}} = 0 \) the critical value of \( \gamma_{E2} \) can be evaluated. For all practical purposes, i.e. when \( x \approx 1 \), \( \gamma_{E2} \) is given by

\[ \gamma_{E2} = \frac{1}{\gamma_{L1} + \gamma_{le}} \left[ \frac{1}{3R} \left\{ \frac{3R \gamma_{fe}(\gamma_{le} + \gamma_{fe}) + 4(\gamma_{le} + \gamma_{fe})(\gamma_{L1} + \gamma_{le})}{3R \left[ \gamma_{o} (\gamma_{le} + \gamma_{fe}) + (\gamma_{le} + \gamma_{fe}) \gamma_{El} + (\gamma_{le} + \gamma_{fe}) \right]} \right\} - 2(\gamma_{KL} + \gamma_{le}) \right] \gamma_{L1}(\gamma_{le} + \gamma_{fe}) \quad \cdots \text{(8.46)} \]

8.3.5 The admittance of the exciter source for the filter in the feedback path—

The admittance at the ports 22 in fig. 8.11 is given by the relation

\[ Y_{dd} = \frac{\Delta}{\Delta_{44}} \quad \cdots \text{(8.47)} \]

where \( \Delta_{44} \) is the determinant obtained after deleting the 4th row and the 4th column from the circuit determinant \( \Delta \). When the system starts to oscillate the circuit determinant is zero. This means that the admittance is also zero. When
FIG. 8-14.  THE PHASE FREQUENCY CURVE OF THE AMPLIFIER FOR A SINGLE VALUE OF $Y_{E2}$
the circuit is used as a selective amplifier, and $y_{o4}$ can be shown to be quite low in the selective range of frequencies. However, solving eqn.(8.47) transforms to

$$y_{o4} = \frac{(y_{L1} + y_{ie})(y_{ie} + y_{fe} + y_{ie} + y_{z}) - y_{ie}(y_{ie} + y_{fe})}{y_{L1} + y_{ie}}$$

$$- \frac{y_{z}^2 \left[ y_{fe} (y_{ie} + y_{fe}) + y_{L1} y_{ie} \right]}{(y_{L1} + y_{ie}) \left[ (y_{ie} + y_{fe}) \right]^2} \left( y_{ie}^2 + y_{fe}^2 + y_{ie}^2 + y_{ze} - y_{ie} (y_{ie} + y_{fe}) \right)$$  

(8.48)

When this equation is solved for specific values of the parameters, within the selective frequency range of the amplifier $y_{o4}$ does not show any marked increase in value. Outside the selective range, transmission through the filter is very small (fig.8.12) and hence the admittance value at port 22' is of no practical consequence.

8.3.6 Circuit and Results

The circuit of the experimental setup is shown in fig.8.16, with the passive elements marked. The measured $y$-parameters of the transistors used are $y_{ie1} = y_{ie2} = 3.53 \times 10^{-4} \text{ mho}$ and $y_{fe1} = y_{fe2} = 285 \times 10^{-4} \text{ mho}$. The experimental values of the gains at different frequencies with the different settings of $R_{E2}$ are found to be in good agreement with the theoretical results. The experimental points are shown by crosses in fig.8.12. Oscillation started in the system when $y_{E2}$ reached a critical value of $4.45 \times 10^{-4} \text{ mho}$. The variation of the frequency of oscillation by decreasing $y_{E2}$ was practically negligible. The admittance at port 22' (fig.8.16) was measured just below the oscillatory condition and was found to be nearly zero.
FIG. 8-15 THE $Y_{E_2} - X$ CURVE OF THE OSCILLATOR.

FIG. 8-16 THE EXPERIMENTAL CIRCUIT
8.3.7 Remarks

A modified form of the Wien network has been used to form a selective amplifier and an oscillator. In the transistorized circuit this passive filter introduces frequency selective regenerative current feedback. The convenient output terminal is marked \( Q \) in fig. 8.16. A higher gain is possible at this point depending on the collector resistance of \( T_2 \) and also the input impedance of the subsequent coupled stages does not materially affect the feedback circuit. Variation of the circuit elements especially the admittance \( y_s \) affects the gain and frequency conditions to certain extent which can be checked from the relations deduced above. Widely different \( y_{ie} \)'s of the two transistors bring in complexity in the analysis, nevertheless with a little effort similar results are arrived at.

The practical amplifier circuit is designed to obtain a preselected value of the Q factor only by adjusting \( y_{E2} \) which is not possible when the signal generator admittance contributes to the \( y_s \) parameter.
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* Material in Chapter VIII is based mainly upon the results
given in these publications.