14

ALC Oscillators

14.1 INTRODUCTION

As pointed out in Section 3.10, the frequency is a function of the crystal current, I_x (and power). Therefore, a major cause of instability and aging in high-performance oscillators is the variation in crystal current. All the oscillators considered thus far have the common characteristic that the crystal current is determined by the relatively noncontrolled nonlinear action of semiconductors and associated components. The action changes with time, temperature, and many other factors. It therefore follows that the most stable oscillators are most likely to be of the type wherein the crystal current is monitored and constrained to have a high degree of constancy as measured by $\partial I_x/I_x$. The type of oscillator performing the latter process is the Automatic Level Control oscillator. The word *level,* in this case, is synonymous with *crystal current.*

The single transistor oscillator performs the frequency generation and limiting in the same circuit. As a result, the frequency generation function cannot be optimized without destroying the limiting function. The same is true, but to a smaller extent, of the two-transistor oscillators of Chapter 13. However, it is extremely desirable to be able to design circuits in which the frequency generation and limiting are performed by separate relatively independent portions of the complete oscillator so that the operation of each can be optimized or modified without grossly affecting the other function. Again, the oscillator circuit, having this feature, is the ALC type.

The ALC oscillator is used only where the *highest* long- and short-term frequency stability is required. Usually, the oscillator is placed in a closely temperature controlled oven to minimize the instability due to varying temperature.

The price paid for the superior performance of the ALC oscillator is the complexity of the circuitry and the consequent poorer reliability and higher cost.

14.2 GENERAL DESCRIPTION OF THE ALC OSCILLATOR

This oscillator may be treated as a system composed of many components or subsystems, each of which can be specified and designed as necessary for the desired overall system performance.

Figure 14.1*a* shows the block diagram of the typical ALC oscillator. Figure 14.1*b* is Fig. 14.1*a* redrawn to demonstrate that the block diagram of the regulator type of control system is also applicable to the ALC oscillator. The tremendous amount of theory[14.1–14.3] developed to analyze and design this control system is therefore completely usable for the ALC oscillator.

14.2.1 The Steady-State g_m Loop and the Calculation of V_f

Figure 14.1*b* is somewhat incomplete in that it does not take into account the value of V_C requird by the g_m versus V_C characteristic of the oscillator. This is analogous to the voltage required to make the average frequency of the controlled oscillator equal to that of the reference oscillator in a phase-lock

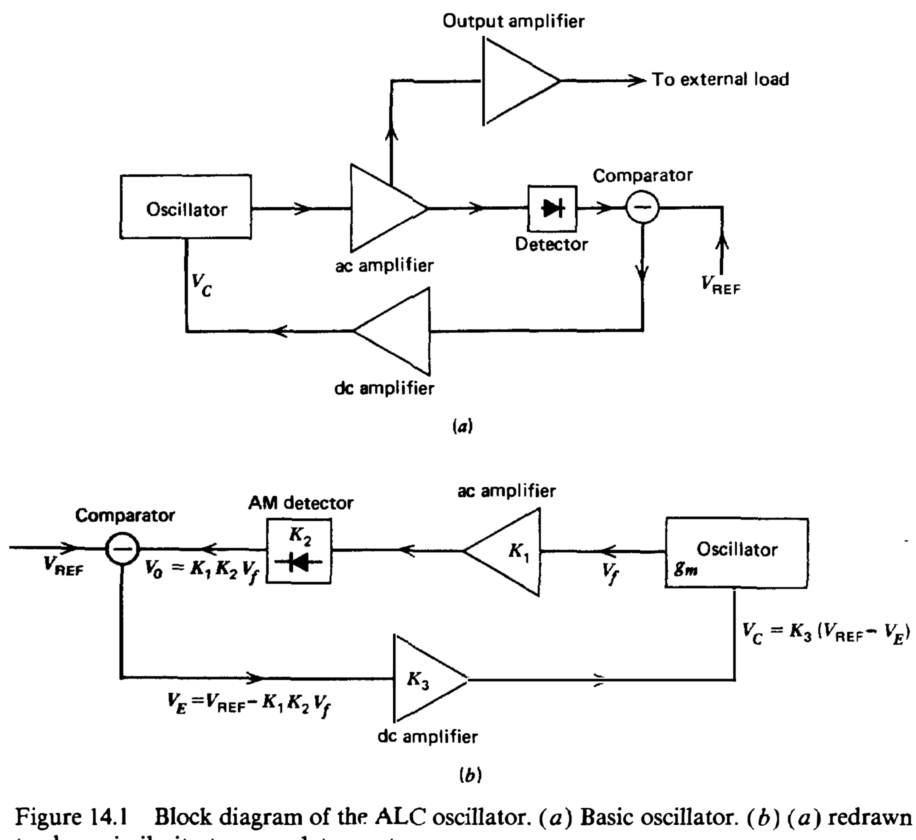


Figure 14.1 Block diagram of the ALC oscillator. (*a*) Basic oscillator. (*b*) (*a*) redrawn to show similarity to a regulator system.

loop. In the phase-lock system the steady-state control voltage is usually not important as its major effect is to create a constant phase difference more or less proportional to the difference between the actual operating frequency and the frequency of the controlled oscillator at $V_C = 0$. However, it does greatly affect the acquisition process if the required V_C is in the neighborhood of the maximum voltage available from the phase detector and may even make the acquisition process impossible, which is also analogously true in the ALC oscillator.

As seen from Chapters 1 and 5, every oscillator has a g_m or equivalent parameter which must have a unique value, g_{me}, to satisfy the conditions for equilibrium oscillation. In an ALC oscillator where all the frequency generating components are designed to operate in a completely linear fashion, the output amplitude *is independent of* g_{me}.

g_{me} is determined by the oscillator components as described in Chapter 5. This, in turn, fixes the value of V_C because of the relationship

$$g_m = F(V_C) \tag{14.1}$$

where

$$\frac{\partial g_m}{\partial V_C} \text{ is always positive} \tag{14.1a}$$

From Fig. 14.1*b*,

$$V_C = K_3(V_{\text{REF}} - K_1 K_2 V_f) \tag{14.2}$$

or

$$V_f = \frac{1}{K_1 K_2}\left(V_{\text{REF}} - \frac{V_C}{K_3}\right) \tag{14.2a}$$

Also, let

$$V_f = r_{\text{osc}} I_x \tag{14.3}$$

It is interesting to observe the oscillator amplitude behavior in accordance with Eqs. (14.1) to (14.2a).

When the oscillator is first started, $V_f = 0$ and $V_C = K_3 V_{\text{REF}}$. Therefore, $g_m \gg g_{me}$ and V_f increases. As V_f increases, V_C decreases in accordance with Eq. (14.2) and g_m decreases in accordance with Eq. (14.1). Eventually g_m reaches g_{me} and the oscillator assumes the equilibrium state. To ensure good oscillator starting performance g_m at $V_f = 0$ should be at least $4g_{me}$.

If V_{REF} is varied, V_f also varies in accordance with Eq. (14.2a) since V_C remains constant.

From the above it is seen that it is very important that K_1, K_2, and K_3, and r_{osc} remain constant in order to maintain a constant I_x.

14.2.2 Requirements for K_3 and V_C

The values of K_3 and V_C required to achieve a specified contribution, $\partial(\Delta f/f)$, by the variation of I_x, to the frequency instability, is now computed. From Section 3.10.2

$$\frac{\Delta f}{f} = aI_x^2 \tag{14.4}$$

where a is a constant depending upon the crystal cut, from which,

$$\partial\left(\frac{\Delta f}{f}\right) = \frac{2a\,\partial I_x}{I_x} I_x^2 \tag{14.5}$$

From Eqs. (14.2a) and (14.3),

$$I_x = \frac{1}{r_{osc}K_1K_2}\left(V_{REF} - \frac{V_C}{K_3}\right) \tag{14.6}$$

from which

$$\partial I_x = -\frac{\partial V_C}{K_1K_2K_3r_{osc}} \tag{14.7}$$

Assume that due to instability of components or environment, V_C shifts by an amount ∂V_C; then from Eqs. (14.6) and (14.7)

$$\frac{\partial I_x}{I_x} = -\frac{\partial V_C}{K_3V_{REF} - V_C} \tag{14.8}$$

From Eqs. (14.5) and (14.8)

$$\partial\left(\frac{\Delta f}{f}\right) = -\frac{2aI_x^2\,\partial V_C}{K_3V_{REF} - V_C} \tag{14.9}$$

For the case where $K_3 = 1$

$$\partial\left(\frac{\Delta f}{f}\right) = -\frac{2aI_x^2\,\partial V_C}{V_{REF} - V_C} \tag{14.10}$$

For the case where $K_3 \gg 1$ and $V_{\text{REF}} > V_C$,

$$\partial\left(\frac{\Delta f}{f}\right) = -\frac{2aI_x^2\,\partial V_C}{K_3 V_{\text{REF}}} \tag{14.11}$$

Consider the following example:

From Ref. 3.50, $a = 2 \times 10^{-8}$ for the fifth overtone best SC-cut crystal at 5 MHz. Let

$$I_x = 0.1 \text{ mA}$$

then Eq. (14.9) becomes

$$\partial\left(\frac{\Delta f}{f}\right) = -\frac{4 \times 10^{-10}\,\partial V_C}{K_3 V_{\text{REF}} - V_C}$$

$$\textit{make } K_3 = 1 \text{ (No } K_3 \text{ amplifier)}$$

$$V_{\text{REF}} = 3 \text{ V}$$

$$V_C = 1 \text{ V}$$

then for $|\Delta(\Delta f/f)| = 10^{-12}$, ΔV_C cannot exceed 5 mV which is a very small quantity. The advantage of having a large value for K_3 becomes obvious.

Since ΔV_C is caused by the necessity for a change in $g_m = \Delta g_m$, it follows that

$$\Delta V_C = \left(\frac{\partial V_C}{\partial g_m}\right)\Delta g_m \tag{14.12}$$

So that for any Δg_m, ΔV_C is smallest when $(\partial g_m/\partial V_C)$ is largest.

In connection with the above, it should be noted that V_C is an excellent monitoring point. If V_C changes, g_m changes in accordance with Eq. (14.1). The reason for the change of g_m should be investigated, as it may be accompanied by relatively significant frequency shifts. For example, frequency aging of the crystal will not change the g_m. However, a change in the crystal resistance and changes in the other circuit parameters will change the g_m.

14.2.3 The Approximate Noise Performance of the ALC Loop

From Fig. 14.1*b*, one may intuitively surmise that since V_{REF} is assumed noise-free, then V_0 is also noise-free. If K_2 and K_1 are also noise-free, then V_f must also be noise-free. Furthermore, if r_{osc} is noise-free, then I_x must also be noise-free.

It is implied in the above that since the detector is responsive only to AM signals, V_f will be only AM noise-free. However, Kulagin[14.4] has shown that the control loop also tends to reduce the FM noise because of the interdependence of AM and FM noise.

14.3 DESCRIPTION OF THE COMPONENTS OF THE ALC OSCILLATOR

This section will consider those characteristics of the components of the ALC oscillator which must be optimized for their proper functioning in the ALC circuit.

14.3.1 The Oscillator Circuit Requirements

The oscillator circuit may be any of those discussed previously. It must have the following properties:

1. It must have an input terminal to which a dc signal is fed for varying the g_m.
2. Equation (14.3) must be satisfied, and r_{osc} must be as noise-free as possible.
3. The value of crystal current is set at the compromise value most suitable for the desired short-term and long-term frequency performance.
4. The electronic circuitry is designed for lowest noise and maximum stability.

14.3.2 The Pierce Oscillator

A circuit very often used in ALC oscillators is the Pierce oscillator discussed in Chapters 5 and 7; the schematic diagram is shown in Fig. 14.2.

It is seen that it is exactly the same as that previously described except that the transistor has been replaced by an amplifying system, the g_m of which is controlled by V_C. Since the limiting function is mainly provided by circuitry external to the Pierce oscillator, this oscillator can be designed for lowest noise output and highest stability. Some of the measures that should be taken to achieve the desired performance are as follows:

1. Make X_1 very small compared to the input impedance at point 1. This is tantamount to minimizing the source resistance seen at point 1.
2. Similarly, make X_2 small compared to the output impedance at point 2.
3. Make the g_m versus V_C characteristic of Y_A as stable as possible.
4. Make the active device operate in the linear mode to minimize phase noise.

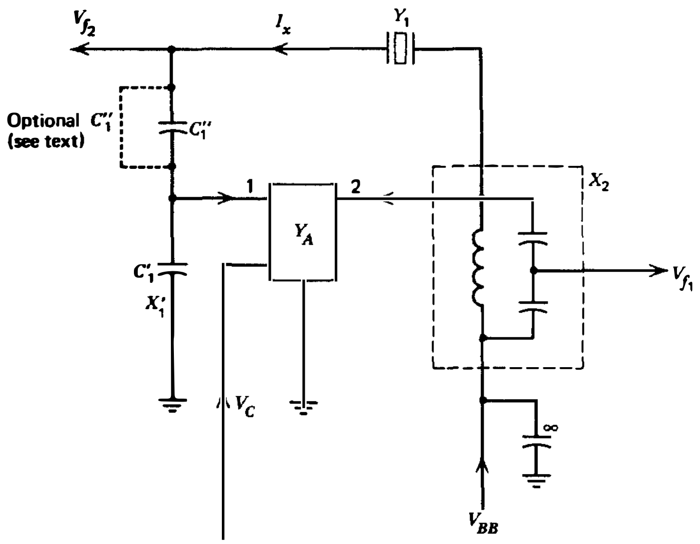


Figure 14.2 The general Pierce oscillator.

Figure 14.2 shows the output as V_{f_1} which is in the output circuit of Y_A and which is the output normally used. However, V_{f_2} has the advantage that at high Fourier frequencies, its noise content can be considerably less than that of V_{f_1}. This is due to the filtering action of the crystal at Fourier frequencies outside of the crystal bandwidth.

Figure 14.3*a* shows the theoretical phase noise characteristic of V_{f_1} and V_{f_2} in the frequency domain (see Section 14.4). The figure also shows the phase noise of the circuitry less the crystal $S_{\phi_\delta}(\mathbf{f})$. $\mathbf{f}_{\delta_1}$ is the Fourier frequency where the flicker noise of the active device is equal to the white noise.

Figure 14.3*b* is the same as Fig. 14.3*a* except that the operating Q is larger. Both Figs. 14.3*a* and 14.3*b* are somewhat unrealistic in that they do not include the contribution in the buffer and final amplifiers additive noise, $S_{\phi_{\delta a}}(\mathbf{f})$, to the overall noise. Figure 14.3*c* is Fig. 14.3*b* modified to show the effect of the amplifier noise.

To increase the V_{f_2} output, a relatively small-value capacitor, C_1'', is placed in series with C_1' and the voltage drop across it, produced by the crystal current, supplies the V_{f_2} output. This becomes a problem because as this capacitor approaches the value of the tuning capacitance it severely restricts the tuning range. This problem is discussed and solved by Burgoon and Wilson in Ref. 14.6.

Output V_{f_2} also has the advantage that the tuning in the Y_A output circuit does not affect the value of r_{osc}. Very often the output circuit tuning is used as a means of fine frequency tuning the oscillator and will incidentally influence the value of r_{osc}.

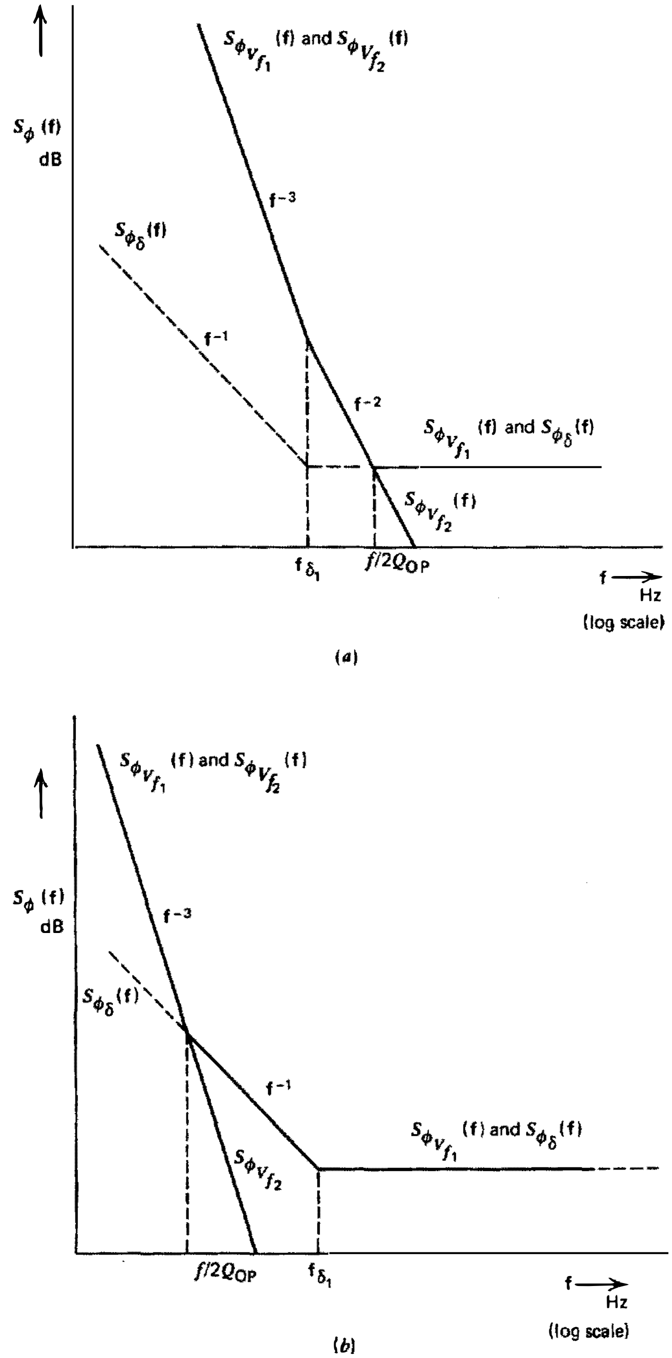
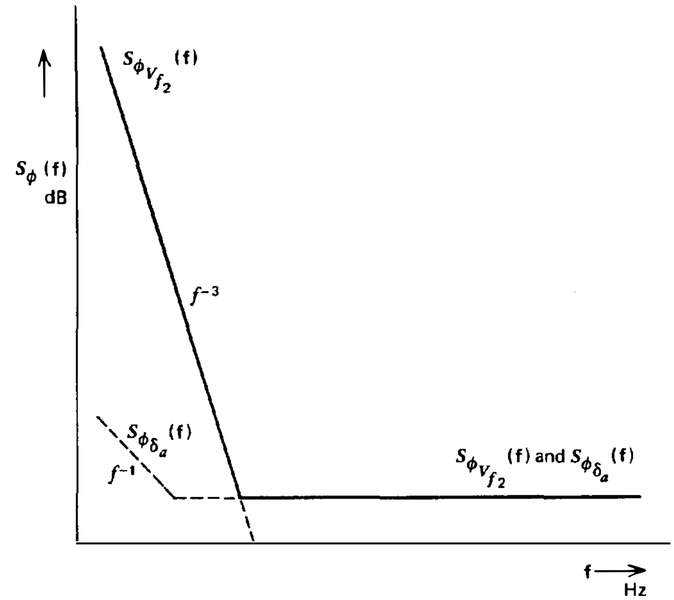


Figure 14.3 Phase noise characteristics of V_{f_1} and V_{f_2}. *Note*: All scales log. (*a*) Medium Q_{op}. (*b*) High Q_{op}. (*c*) (*b*) modified to include additive amplifier noise.



(c)

Figure 14.3 (*Continued*).

14.3.2.1 Y_A as a Voltage Variable g_m Transistor

Figure 14.4a shows the case where Y_A in Fig. 14.2 is made to be a single voltage variable g_m transistor. This is a very frequent realization of Y_A. R_E is a low-value resistor which reduces the transistor flicker phase noise.[14.10] r_2 serves to stabilize the g_m versus V_C characteristics but at the cost of a considerable reduction in the value of dg_m/dV_C.

It is interesting to compute the value of dg_m/dV_C.

$$I_E = \frac{V_C - 700}{r_2 + R_E}$$

From Eqs. (2.44) and (2.45a),

$$g_{m_0} \approx 0.038 I_E$$

where g_{m_0} is the small-signal g,

$$g_{m_0} \approx \frac{\quad}{R_E + r_2} \ ℧ \tag{14.13}$$

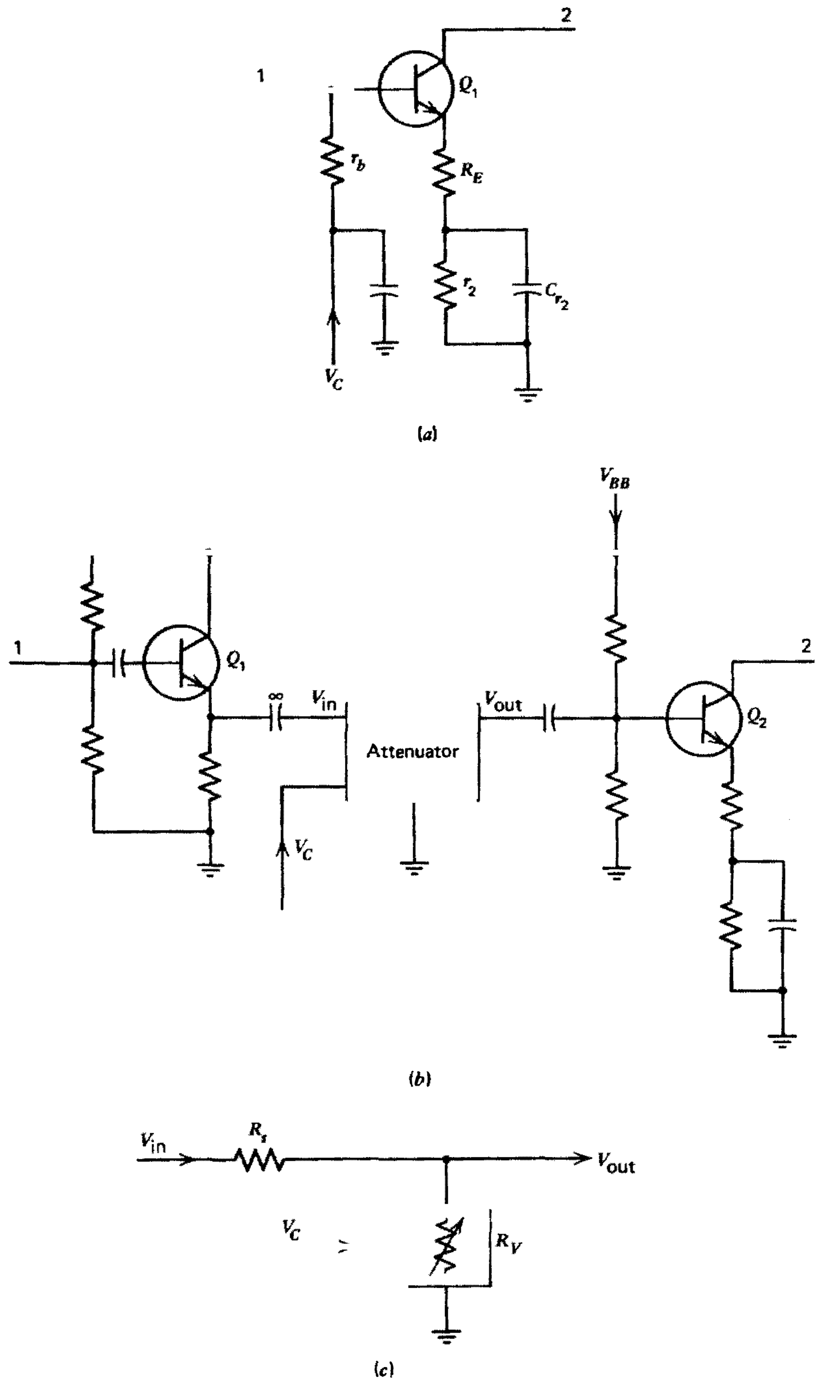


Figure 14.4 Realization of Y_A in Fig. 14.2. (a) Voltage variable g_m transistor. (b) Constant g_m transistors with auxiliary voltage variable attenuator. (c) Attenuator diagram.

V_C is in mV, I_E is in mA and;

$$g_m = \frac{g_{m_0}}{1 + g_{m_0} R_E}$$

$$\approx \frac{0.038(V_C - 700)}{R_E[1 + 0.038(V_C - 700)] + r_2} \tag{14.14}$$

also

$$\frac{dg_m}{dV_C} \approx \frac{0.038}{R_E[1 + 0.038(V_C - 700)] + r_2} - \frac{(0.038)^2(V_C - 700) R_E}{[R_E[1 + 0.038(V_C - 700)] + r_2]^2} \tag{14.15}$$

$$\approx \frac{0.038}{R_E[1 + 0.038(V_C - 700)] + r_2} \tag{14.15a}$$

for most circuits.

Equations (14.14) and (14.15) are useful for circuit analysis and troubleshooting.

Section 14.3.2, item 4 states that the transistor should operate in the linear mode. Figure 2.12*a* shows that when $R_E = 0$, $V_b \lesssim 10$ mV. Figure 2.13*b* shows that higher values of V_b are permissible for the linear mode depending upon the value of R_E and I_E. However, if the signal is increased beyond these limits by increasing the crystal drive, the white phase noise will decrease but the noise at low Fourier frequencies may increase. Thus if low phase noise at Fourier frequencies greater than 10 Hz is desired, high crystal drive should be employed.

Evidence exists that for minimum noise the β_o and the f_T of the transistor should be as high as possible (see Healey, Ref. 14.5). This also has the effect that the input impedance of the transistor is increased so that the requirement of Section 14.3.2, item 1, should be easier to satisfy. The main disadvantage of using extremely high f_T transistors is that they are more prone to spurious oscillations, which may require suppression and consequently greater circuit complexity.

FETs also have desirable noise properties and consideration should be given to their use. FETs, however, have the disadvantage of having low values of dg_m/dV_C, which is not important when the FET is used as a fixed gain device.

When Q_1 is a bipolar transistor, consideration should be given to making it a PNP type as there is some evidence that *PNP* transistors have less flicker phase noise than *NPN* transistors.

14.3.2.2 Y_A as Constant g_m Transistors with Auxiliary Voltage Variable Attenuator

Figure 14.4*b* shows another realization of Y_A. It consist of an emitter follower feeding a voltage variable attenuator. Examples of suitable attenuators are an FET or PIN diode used as a voltage variable resistor, R_v, in series with a fixed resistor, R_s, as shown in Fig. 14.4*c*.

The output of the attenuator feeds a constant gain amplifying network, Q_2. This network can be designed for maximum performance since it is operating in a fixed mode.

If the voltage gain of Q_1 is assumed to be 1, then

$$g_m = \frac{V_{\text{out}}}{V_{\text{in}}} g_{m_{Q2}} \tag{14.16}$$

An example of an oscillator using the attenuator type of Y_A is discussed by Babbitt in Ref. 14.7.

14.3.3 The ac Amplifier K_1 in Fig. 14.1*b*

This amplifier has the dual function of amplifying the oscillator level to a magnitude sufficient to drive the rectifier in its linear region and to supply a relatively large signal to the output amplifier.

The necessary characteristics of this amplifier are very high gain stability and noise freedom. A large amount of both local and overall negative feedback will achieve both of these characteristics. If the operating linearity of this amplifier is extremely good, it need not be tuned, provided the frequency is low enough to permit the necessary gain without tuned circuits.

The magnitude of K_1K_2 is determined by Eqs. (14.2) and (14.6). Since, as shown in Section 14.3.4, K_2 has a relatively small range of values, Eq. (14.2) also determines the value of K_1.

Pustafari describes a typical amplifier circuit in Ref. 14.8.

14.3.4 The Detector K_2 in Fig. 14.1*b*

The detector is a rectifier which converts the ac output of K_1 to a dc signal. Its necessary characteristics are that it must be noise-free and its conversion ratio K_1 must be highly stable. To achieve both these characteristics, the circuit is usually a voltage doubler operating in its linear region, thus necessitating an input voltage of 1 V rms, minimum. The Schottky diode is recommended for best noise performance. The output must be well filtered to prevent the ac input from supplying an undesirable and unintentional coupling to the oscillator circuitry.

14.3.5 The dc amplifier K_3 in Fig. 14.1*b*

The gain K_3 of this amplifier is calculated as described in Section 14.2.2. Its necessary characteristics are noise freedom, gain constancy, and very low drift. It is usually a high-performance operational amplifier with massive feedback. Its output saturation voltage must be sufficiently large to ensure good oscillator starting.

14.3.6 The Reference Voltage V_{REF} in Fig. 14.1*b*

As seen in Section 14.2, this voltage plays a very important role. In many ALC oscillators, V_{REF} is not explicitly identifiable, but it must exist and very often it is not constant; for example, the turn-on voltage (or contact potential) of a transistor which is relatively ill-defined. It is therefore extremely desirable that V_{REF} be explicit and capable of being intentionally varied to produce the desired crystal current. Obviously, V_{REF} must be noise-free and extremely stable as already pointed out.

Felch and Israel, in Ref. 14.9, provide an analysis of the ALC system wherein V_{REF} is not explicit. They also describe a precision oscillator using this type of ALC system.

14.3.7 The Comparator

The comparator in Fig. 14.1*b* is a simple subtraction circuit. The subtraction function is usually obtained by making the polarity of the diodes in the detector such that its output polarity is opposite to that of V_{REF} and the sum of the two voltages is then obtained.

14.3.8 The Output Amplifier in Fig. 14.1*a*

The functions of the output amplifier are to

1 Provide power gain.
2 Not add discernable noise.
3 Provide isolation so that variation of the external load circuit does not produce a frequency change.

All the above functions are admirably performed by the Cascode Amplifier Circuit discussed in Section 13.3.

14.3.9 Noise Due to Components

The reader is alerted that where the best stability is required, due consideration should be given to the noise present in all components, including resistors, capacitors, inductors, transistors, and diodes. It is very difficult to make generalizations about the magnitude of the noises contributed by the components, but they may be significant. The noise magnitude depends upon the type of component, the individual component in a given lot, and upon the manufacturer. As the manufacturing technology changes, the noise also changes. For the best noise performance it is necessary to resort to selection. (See also Ref. 14.10.)

14.3.10 Output Crystal Filters

Where the noise, at Fourier frequencies relatively distant from the carrier, is considered excessive, it may be reduced by suitable crystal filters. However, care should be taken that the filter does not introduce noise due to excessive drive in the crystals in the filter.

14.4 THE PREDICTION OF THE APPROXIMATE PHASE NOISE PERFORMANCE OF THE OSCILLATOR

Figures 14.3 show the approximate phase noise performance in the frequency domain of the Pierce oscillator of Section 14.3.2. This section presents the theory on which the figures are based.

Leeson[14.11] has proposed a heuristically derived oscillator phase noise model. The validity of this model has been theoretically confirmed by Sauvage.[14.12]

According to this model,

$$S_{\phi_{V_{f_1}}}(\mathbf{f}) = S_{\phi_8}(\mathbf{f})\left(1 + \left(\frac{f}{2Q_{\mathrm{op}}\mathbf{f}}\right)^2\right) \tag{14.17}$$

Obviously,

$$S_{\phi_{V_{f_2}}}(\mathbf{f}) = S_{\phi_{V_{f_1}}}(\mathbf{f})G_{Y_1}(\mathbf{f}) \tag{14.18}$$

where $G_{Y_1}(\mathbf{f})$ is the transfer function of the Y_1, X_1 network. But

$$G_{Y_1}(\mathbf{f}) \approx \frac{1}{1 + (2Q_{\mathrm{op}}\mathbf{f}/f)^2} \tag{14.19}$$

therefore

$$S_{\phi_{V_{f_2}}}(\mathbf{f}) \approx S_{\phi_8}(\mathbf{f})\left(\frac{f}{2Q_{\mathrm{op}}\mathbf{f}}\right)^2 \tag{14.20}$$

which is an extension of Leeson's model.