13

Other Oscillator Configurations

13.1 INTRODUCTION

This chapter treats oscillator configurations not considered in the previous chapters. The treatment, of necessity, will be qualitative rather than quantitative because space limitations do not permit a full exposition of this extremely extensive subject. Also, because of the enormous number of available circuits, only a sampling can be made.[13.7]

The chapter is concerned with the following:

1 Single transistor oscillators not covered in the previous chapters.
2 Multiple transistor circuits in which a transistor is added to the circuits of Chapters 5 to 11 to improve their performance.
3 Multiple transistor circuits which are substantially different from the basic circuits of Chapters 5 to 11. ALC type oscillators are discussed in Chapter 14.

13.2 SINGLE TRANSISTOR OSCILLATORS

13.2.1 Low-Frequency Crystal Oscillators

As the frequency decreases, the crystal resistance increases markedly, and the circuits previously described are either unsuitable or must be modified to accommodate the higher resistance. The FET is particularly useful because of its very high input impedance. Figures 13.1, 13.2, and 13.3 show oscillator circuits using FETs as the active device. These circuits are suitable for

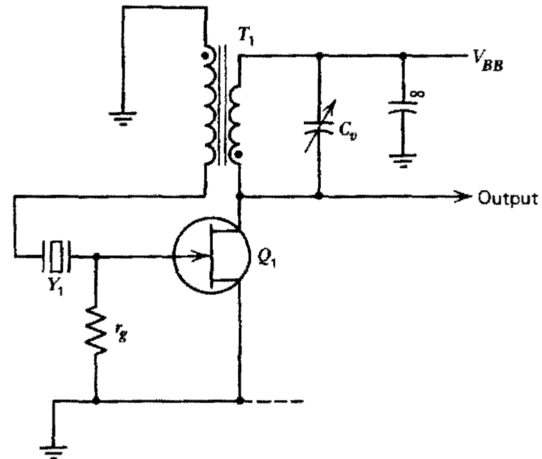


Figure 13.1 Series resonance low-frequency oscillator.

low-frequency crystals, from 16 kHz to above 1 MHz. The amplitude limiting process is the clamp biased limiting described in Sections 6.4 and 6.2.3.

13.2.1.1 Oscillator of Fig. 13.1

The crystal operates near series resonance, so that the operating Q, $Q_{op} \approx Q_x R_1/(R_1 + r_g)$. It is therefore seen that for small deteriorations of Q, R_1

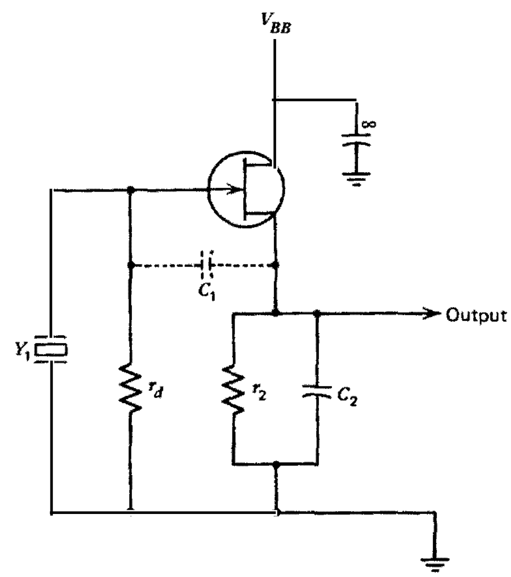


Figure 13.2 Colpitts low-frequency oscillator.

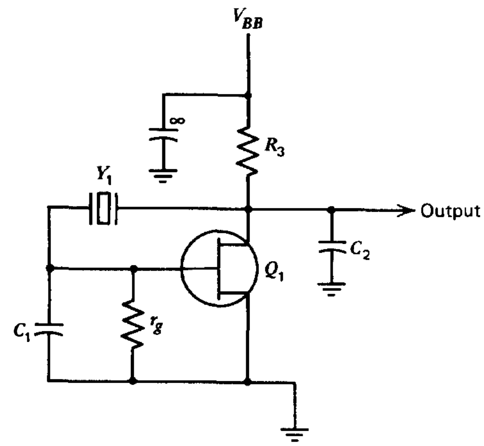


Figure 13.3 Pierce low-frequency oscillator.

must be comparable to or larger than r_g and the circuit is therefore suitable for crystals having very large R_1.

The circuit has the obvious disadvantage that it employs transformer T_1 to provide the 180° phase shift necessary for oscillation. The transformer may be a bulky and relatively expensive device and is therefore undesirable.

If the crystal is short-circuited, the circuit will oscillate at the resonant frequency of T_1 and C_V. When the crystal is used, T_1 and C_V serve to choose the correct response of the crystal.

13.2.1.2 Oscillator of Fig. 13.2

Figure 13.2 shows the Colpitts oscillator described in Chapter 9, adapted for a FET. This circuit is suitable for frequencies from 100 kHz up. C_1, shown in dashed lines, is often omitted and the transistor gate drain capacitance then supplies the functions of C_1. However, this is very poor practice as the transistor capacitance varies widely from transistor to transistor. Also, the circuit performance depends excessively on the transistor characteristics.

13.2.1.3 Oscillator of Fig. 13.3

This is the Pierce oscillator equivalent of the Colpitts oscillator of Fig. 13.2, and its performance is quite similar.

13.2.1.4 Oscillator of Fig. 13.4

This oscillator is basically the isolated Pierce oscillator of Chapter 8. It is suitable for frequencies from 50 kHz up. Resistor R_s is set for the proper crystal drive level and for dependable oscillator starting.

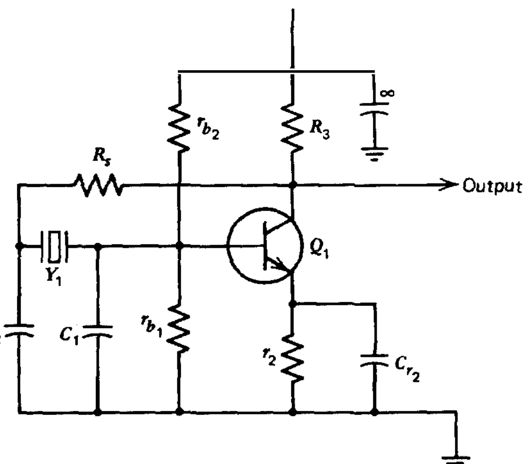


Figure 13.4 Low-frequency version of isolated Pierce oscillator.

13.2.2 Other Oscillator Variations

13.2.2.1 Oscillator of Fig. 13.5

This oscillator is basically the Colpitts oscillator of Chapter 9. The difference is the point of extracting the power output. Because C_L is in series with the crystal, the output is harmonic-free and the noise floor is lower than that of the normal Colpitts oscillator. However, the power output is made very small, in order not to excessively deteriorate the oscillator operating Q.

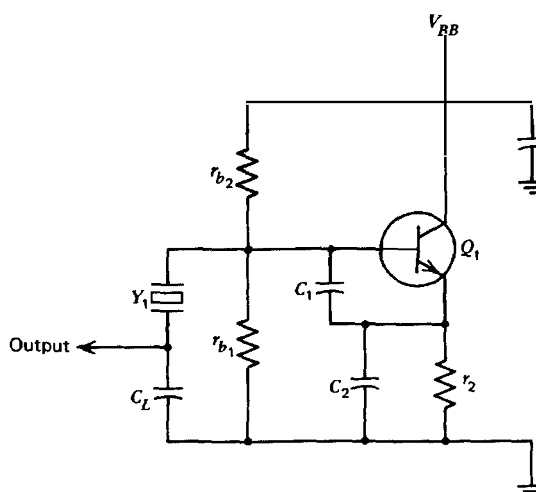


Figure 13.5 Low-noise and harmonic output Colpitts oscillator.

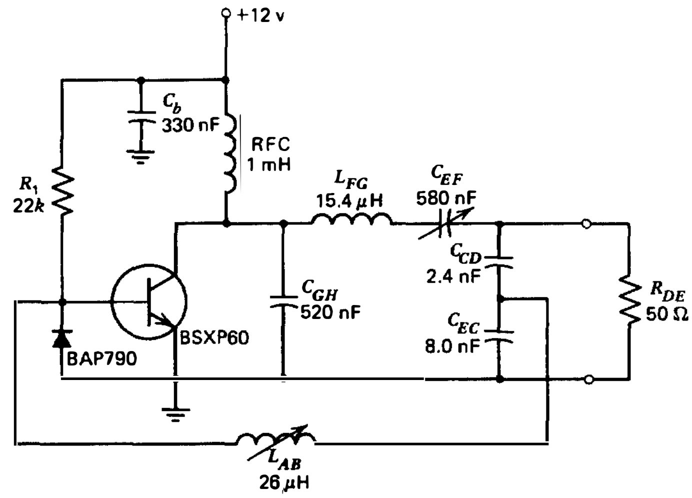


Figure 13.6 Class E high-efficiency tuned power oscillator (from Ebert, J. and Kazimierczuk, M., *IEEE J. Solid-State Circuits* **SC-16**, No. 2, 65 (April 1981).

13.2.2.2 Oscillator of Fig. 13.6 [13.1]

This oscillator is basically a Pierce oscillator but it is unique in that the transistor operates in Class E. By Class E operation is meant that the transistor operates as a switch with a duty cycle of 50%. This results in extremely high conversion efficiency. For example, the subject oscillator provides 3 W output at 2 MHz at a collector efficiency of over 95%. The output wave form is relatively sinusoidal.

13.2.2.3 Oscillator of Fig. 13.7

This oscillator may be considered as a variation of the Butler oscillator discussed in Chapter 11. However, in this case it is more convenient to consider it as a Pierce oscillator with the crystal providing degenerative emitter feedback. To emphasize the similarity between this circuit and the Pierce oscillator (see Fig. 5.1*a*), the like components have been assigned the same symbol number.

The oscillator is another member of the family of oscillators in which the emitter current is also the crystal current. As a result, the output has good wave form and a low noise floor, which is considerably below that of the normal Pierce oscillator. A further advantage is that considerable power output can be obtained without the serious deterioration of the operating Q that exists in the normal Pierce oscillator. However, at low power output, the normal Pierce oscillator has a better operating Q.

The circuit as shown is not popular because it requires more components than the Pierce or Butler oscillator and it does not have superior performance.

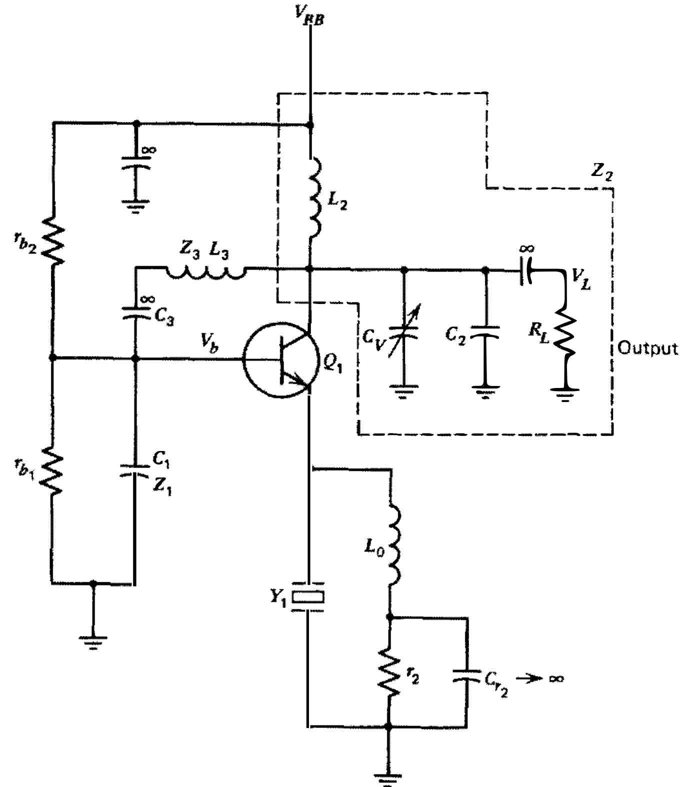


Figure 13.7 Pierce oscillator representation of the Butler oscillator.

However, it can be modified, as explained in Section 13.5.1, to have some very desirable characteristics.

A major problem in this circuit at high P_L/P_x is that V_L/V_b can be large and, therefore, Miller effects play an important role.

13.3 THE DARLINGTON TRANSISTOR PAIR[2.4, 2.5]

Figure 13.8*a* shows two transistors, Q_1 and Q_2, connected in cascade, called the Darlington pair. In many cases the pair can be replaced by an equivalent single transistor Q_{eq}. The approximate relationships between the properties of the pair and the equivalent transistor are now developed.

Q_1 has the parameters listed below. Each of these parameters is defined and calculated in Chapter 2.

$$\beta_{o_1}$$

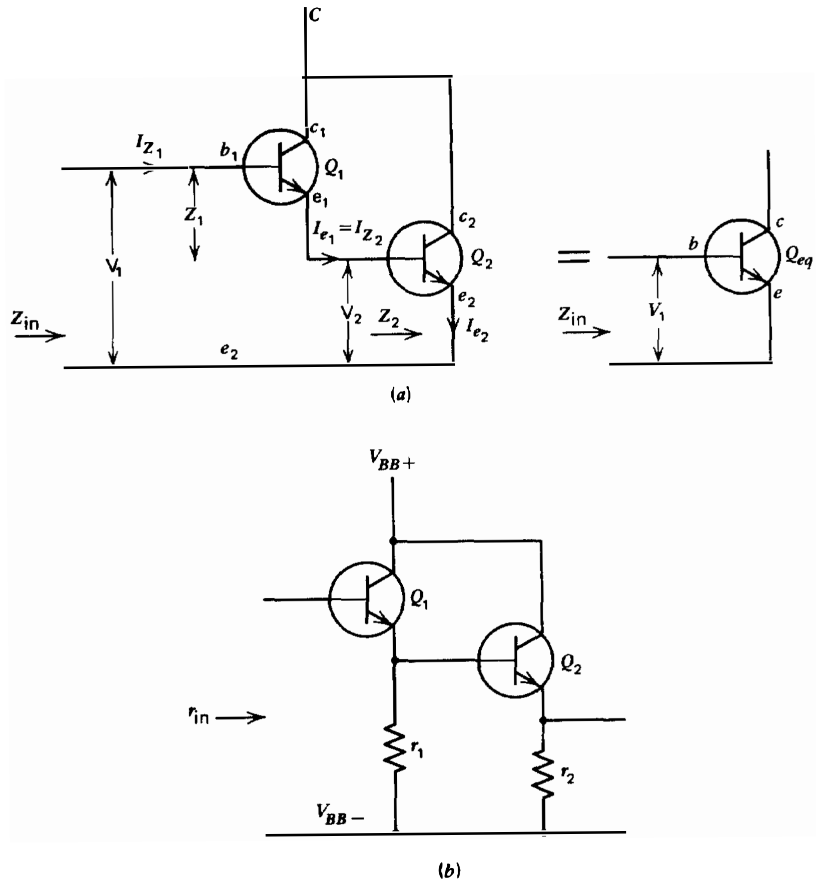


Figure 13.8 The Darlington pair. (a) ac equivalent circuit. (b) dc circuit.

The impedance between b_1 and $e_1 = Z_1$
The capacitance $C_{c_1 b_1}$
The capacitance $C_{c_1 e_1}$

$$g_{m_1} = \frac{I_{e_1}}{V_{b_1}} \qquad \text{assuming } i_C \approx i_E$$

Similarly, Q_2 has the parameters

$$\beta_{o_2}, \quad Z_2, \quad C_{c_2 b_2}, \quad C_{c_2 e_2}, \quad g_{m_2}$$

and Q_{eq} has the parameters

$$\beta_o, \quad Z_{in}, \quad C_{cb}, \quad C_{ce}, \quad g_m$$

13.3.1 Calculation of the Equivalent ac Parameters

From inspection of Fig. 13.8*a*,

$$V_1 = Z_1 I_{Z_1} + Z_2 I_{Z_2} \tag{13.1}$$

$$Z_{\text{in}} = \frac{V_1}{I_{Z_1}} = Z_1 + Z_2 \frac{I_{Z_2}}{I_{Z_1}} \approx Z_1 + \beta_1 Z_2 \tag{13.2}$$

At low frequencies

$$Z_2 \approx \frac{\beta_{o_2}}{g_{m_2}} \tag{13.3}$$

$$Z_{\text{in}} \approx \frac{\beta_{o_1}\beta_{o_2}}{g_{m_2}} \tag{13.4}$$

Also, from Eq. (13.1)

$$g_m \approx \frac{I_{e_2}}{V_1} = \frac{I_{e_2}}{V_2} \cdot \frac{V_2}{V_1} = g_{m_2} \frac{V_2}{V_1} \tag{13.5}$$

It is thus seen that, since $V_1 > V_2$,

$$g_m < g_{m_2} \tag{13.6}$$

Usually,

$$V_2 \to V_1$$

Therefore,

$$g_m \to g_{m_2} \tag{13.7}$$

Also from inspection,

$$C_{cb} \approx C_{c_1 b_1} \tag{13.8}$$

$$C_{ce} \approx C_{c_2 e_2} \tag{13.9}$$

13.3.2 Calculation of the Equivalent dc Parameters

Figure 13.8*b* shows the equivalent dc circuit which includes bias emitter resistors r_1 and r_2. Bias resistor r_1 is not required for the dc function, but is supplied to increase I_{E_1} and f_{T_1}.

$$r_{\text{in}} \approx \beta_{o_1}\left[r_1 \| (\beta_{o_2} r_2)\right] \tag{13.10}$$

If $r_1 \to \infty$,

$$r_{\text{in}} \approx \beta_{o_1}\beta_{o_2} r_2 \tag{13.11}$$

r_{in} is always much larger than the r_{in} of the single transistor circuit with the same I_E.

13.3.3 Darlington Pair Colpitts Oscillator

Figure 13.9 is the schematic of a Colpitts oscillator using two transistors in the Darlington connection. Because of this connection, r_{b_1} and r_{b_2} are very high-value resistors, and therefore do not deteriorate the circuit operating Q, even for crystals having low values of R_{df}.

R_E is provided to reduce the $1/f$ noise of Q_2.

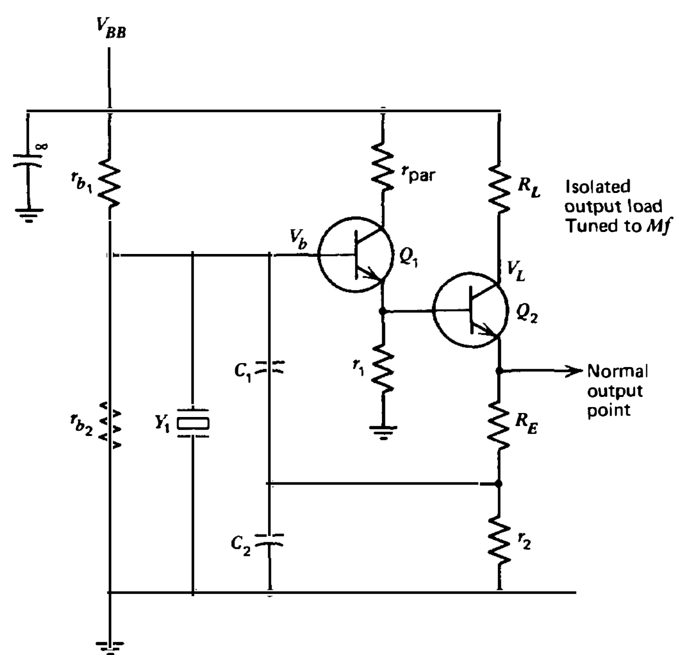


Figure 13.9 Darlington pair Colpitts oscillator.

In this circuit, the normal power output point is at the emitter of Q_2. The power, available at this point without excessive deterioration of the circuit operating Q, is quite small. However, power can be extracted from the collector circuit of Q_2 at fairly high levels.

Because of the excellent isolation provided by the Darlington pair, as stated in Eq. (13.8), V_L can be large compared to V_b without significant Miller effects. Also, the output circuit can be tuned to a multiple, M, of the oscillator frequency. It will be noted that this circuit has performance superior to that of the semi-isolated Colpitts oscillator of Chapter 10, but at the expense of one additional transistor and two resistors.

13.4 THE CASCODE AMPLIFIER[2.4, 2.5]

A circuit configuration which has proven to be useful in the field of oscillators is the well-known Cascode amplifier shown in Fig. 13.10. Its most important

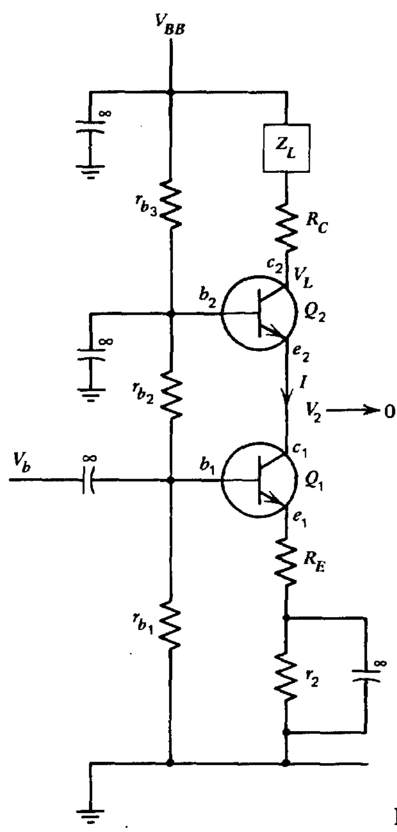


Figure 13.10 Cascode amplifier circuit.

property is that V_L, which may be high, is isolated from the base of Q_1. Instead, C_{cb} of Q_2 is lumped into the load Z_L.

The voltage at the emitter of Q_2, V_2, is equal to Ir_e where r_e is the dynamic emitter resistance of Q_2 and is very small. Therefore, the voltage at that point is very small and the C_{ce} of Q_2 is therefore also effectively lumped into Z_L. For the same reason, C_{be} of Q_2 is also rendered ineffective.

Because V_2 is very small, C_{cb} of Q_1 is effectively lumped into the source impedance.

It is thus seen that all the transistor capacitances which may give rise to the harmful Miller effects, which seriously limit the performance of the $M = 1$ semi-isolated Colpitts oscillator and the Butler oscillator, are rendered ineffective.

R_c and R_E are provided to reduce the phase noise and to minimize the tendency for spurious oscillations.

An example of the use of the cascode circuit is the semi-isolated Colpitts oscillator of Figure 13.11, which is the Colpitts oscillator of Fig. 5.7b to which has been added a cascode type load isolator. As a result, this circuit will work well for all values of M. It is interesting to note that the Q_1 oscillator circuit operates identically to the normal Colpitts oscillator without external load.

The algorithm for the semi-isolated Colpitts oscillator can be easily modified to include this type of oscillator.

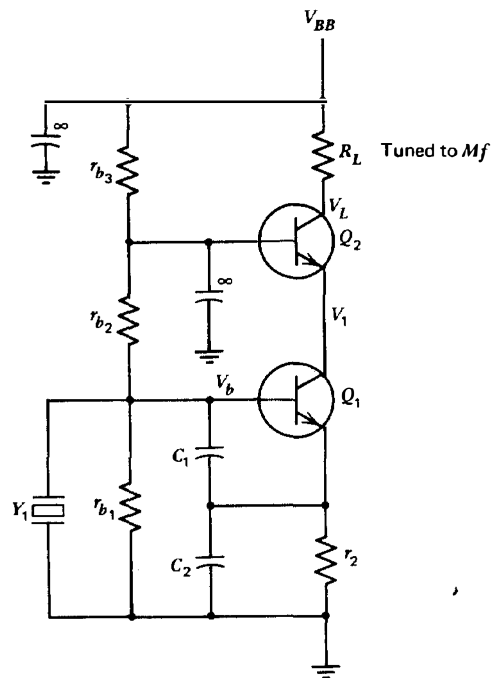


Figure 13.11 Semi-isolated Colpitts oscillator with cascode output circuit.

13.5 OTHER OSCILLATOR CIRCUITS

This section discusses miscellaneous oscillator circuits, other than ALC types, which are substantially unlike those treated in Chapters 7 to 11. Some of the circuits are quite similar except for the limiting means.

13.5.1 The Two-Transistor Oscillator and Limiter Configuration[13 2]

Figure 13.12 shows the oscillator of Fig. 13.7 combined with the cascode limiting stage of Fig. 13.11. This circuit has the advantage that the oscillator

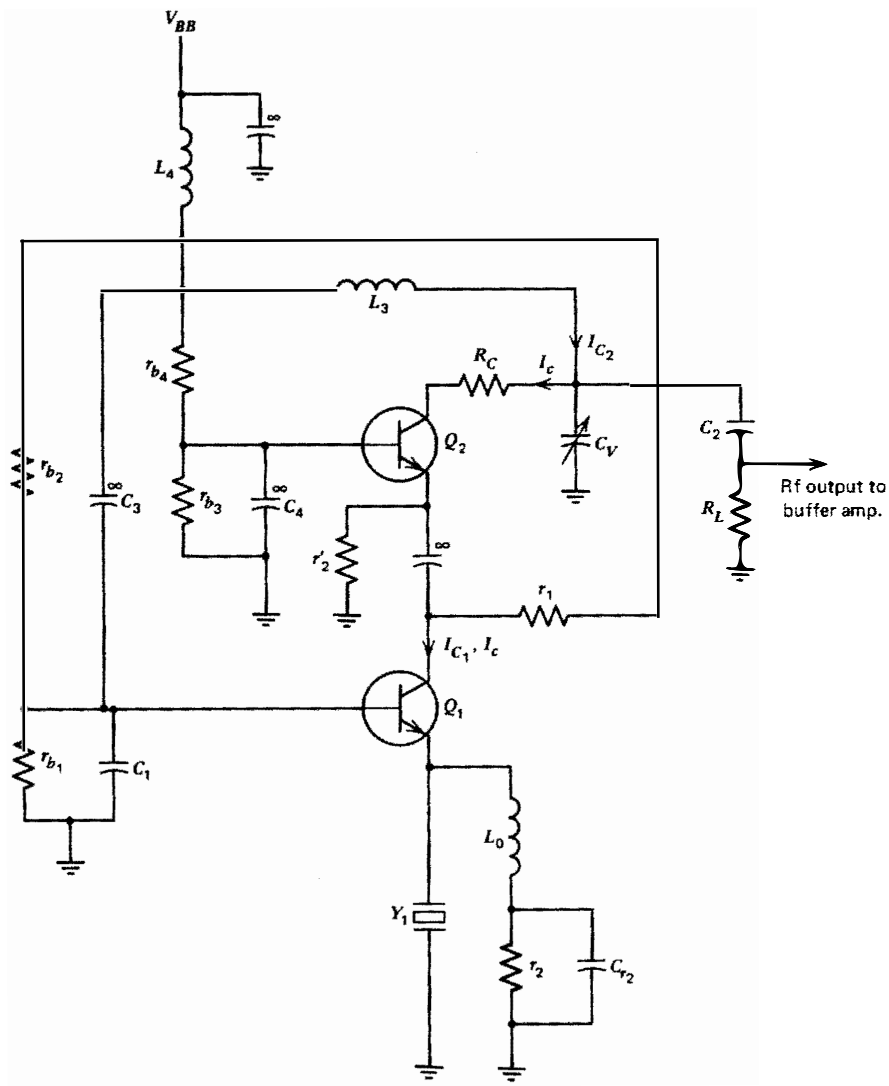


Figure 13.12 The two-transistor oscillator and limiter configuration.

function is separated from the limiting function. Q_1 functions as the oscillator and operates in a linear mode. Q_2 functions as the limiter and its operation is highly nonlinear. I_{C_1} is large to make R_{IN_1} small so as to produce a larger operating Q. I_{C_2} is made $\approx 1.4I_c$ to fix the current I_c in accordance with Eq. (6.23). It should be noted that the isolation provided by Q_2, as described in Section 13.4, is partially, but not seriously, negated by the nonlinear operation of Q_2.

13.5.2 Two-Transistor Diode Limiting Oscillator[13.3]

Figure 13.13*a* shows the oscillator of Fig. 13.7 combined with the cascode isolating stage of Fig. 13.11. Both stages operate in essentially linear mode. Limiting is provided by the diodes CR_1 and CR_2 which function as described in Section 6.5, except that the limiting level is set by the values of $-V_C$ and $+V_C$. Two diodes are provided to produce symmetrical limiting action, which results in slightly less noise, but a single diode limiter is often more convenient as it does not require a $-V_C$ supply and does not markedly increase the noise. The single diode arrangement is shown in Fig. 13.13*b*.

13.5.3 Two-Transistor Oscillator with Separate Variable Gain Type Limiter[13.4]

Figure 13.14 shows the ac circuitry of the oscillator of Fig. 13.7 combined with the cascode isolating stage of Fig. 13.11. This oscillator is designed for very low-noise output and is capable of operation at very high frequencies (100 MHz or more).

Both Q_1 and Q_2 operate in essentially linear mode. The limiting function is provided by IC_1 which acts as a differential limiter and is connected in the feedback path between the output circuitry of Q_2 and the input to Q_1. The differential limiter operates as a variable gain amplifier, the gain of which decreases as the amplitude increases. The differential configuration has excellent phase noise properties as also pointed out by Baugh.[13.5] The amplitude may be conveniently set by changing the dc current in CCG_2.

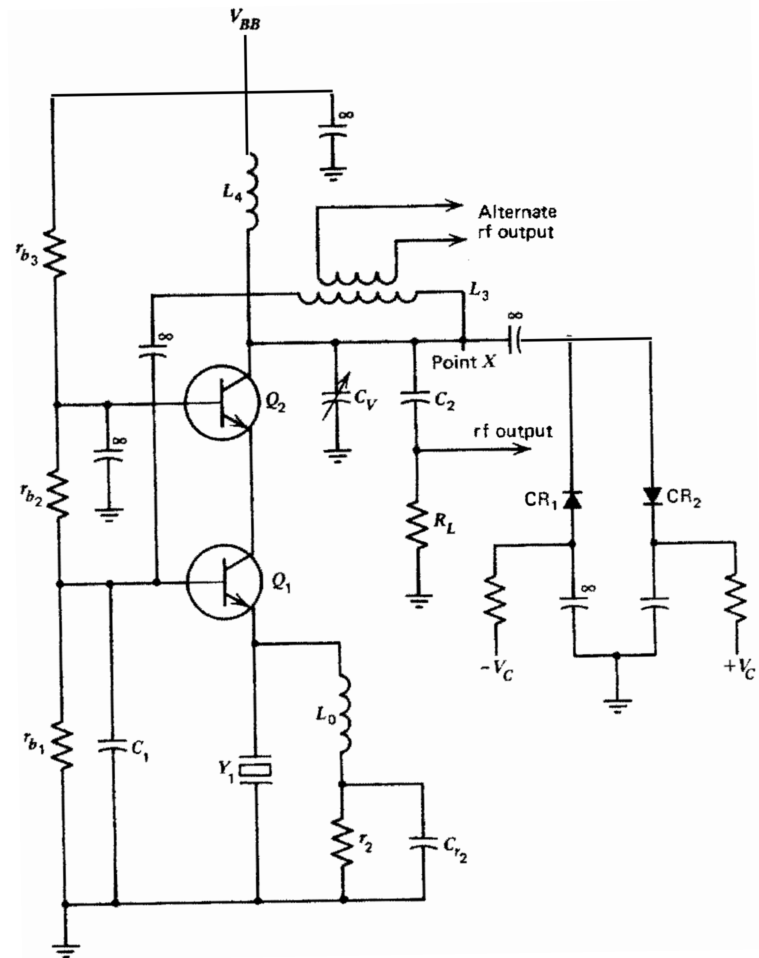
IC_2 functions as a buffer output amplifier. IC_1 and IC_2 are low-noise integrated circuits suitable for operation at very high frequencies, and include their respective constant current generators CCG_2 and CCG_3.

The remaining circuitry is composed of discrete components.

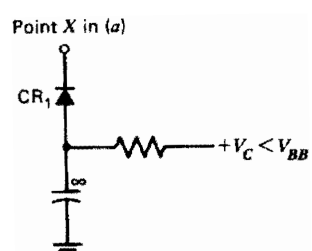
The function of the constant current generators is to maintain the total I_C of each stage at a closely regulated fixed value. It should be noted that their ac output impedance is also very high to prevent loading of Y_1 and to ensure highly balanced operation of the limiting and output amplifiers for best noise operation.

13.5.4 Two-Transistor Emitter Coupled Oscillator

Figure 13.15 shows an oscillator circuit suitable for a wide frequency range, provided that the crystal resistance is not too large or too small. It has been popular for many years and was formerly known as the Butler oscillator.



(a)



(b)

Figure 13.13 Two-transistor oscillator and limiter circuit. (*a*) With two diode limiter. (*b*) Single diode arrangement.

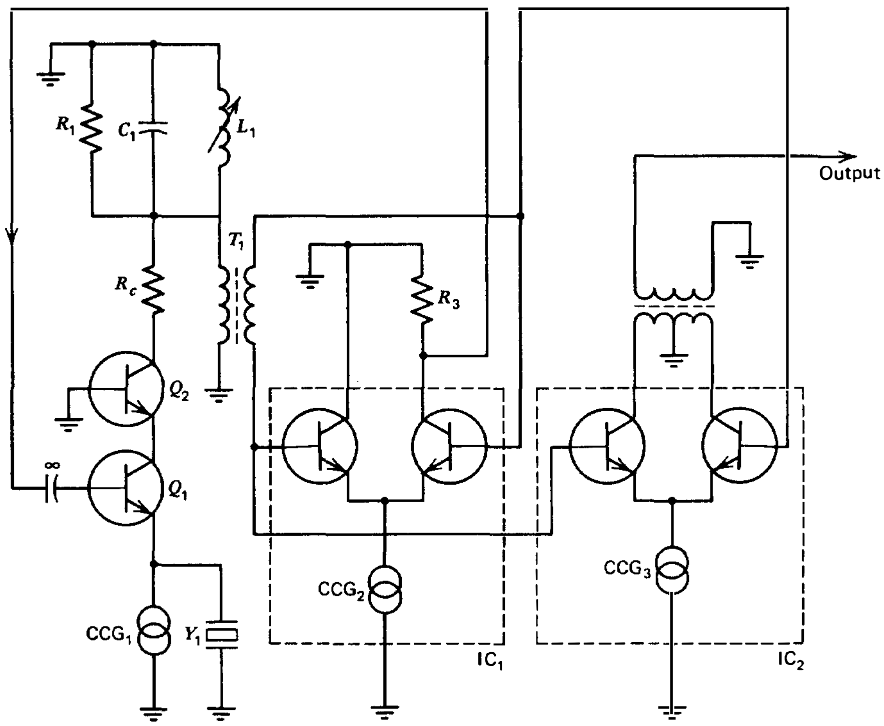


Figure 13.14 Two-transistor oscillator with separate variable gain limiter. (From Rhode, V. L., *Proceedings of the 32nd Annual Symposium on Frequency Control*, p. 418, 1978.)

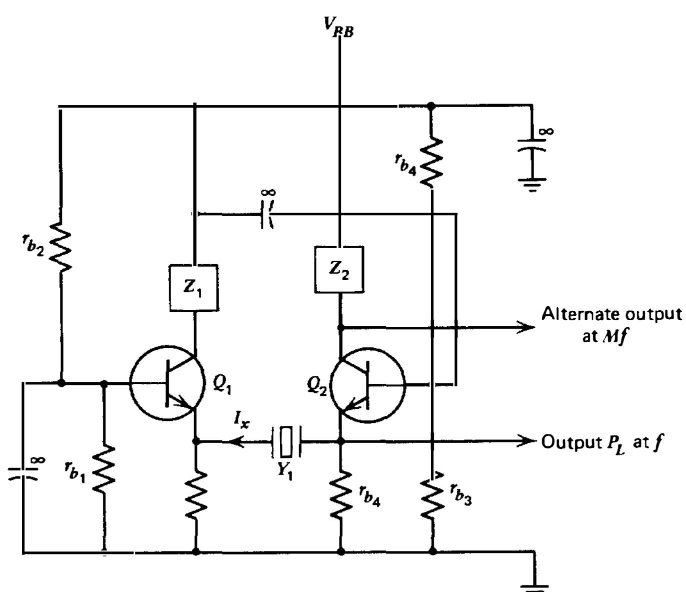


Figure 13.15 Two transistor emitter coupled oscillator (old Butler circuit).

The crystal operates near series resonance. Z_1 is a network for selecting the desired crystal response or overtone. In the simplest case, Z_1 can be a resistor. The limiting takes place in Q_1 and may be either the R_{IN} variation or the collector base voltage type. The operating Q is thus somewhat deteriorated. However, it has excellent noise performance at Fourier frequencies not too near the carrier.

For fundamental output $Z_2 = 0$. For harmonic output Z_2 is a circuit tuned to Mf and the output taken out at the "Alternate Output" point.

Groslambert *et al.*[13.6] describe an oscillator using this circuit.