QUARTERLY REPORT NO. 1
PROJECT NO. 236-206

TRANSISTOR OSCILLATORS OF
EXTENDED FREQUENCY RANGE

By

B. J. DASHER, D. L. FINN,
W. B. JONES, JR., and WAI MUN SYN

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CONTRACT NO. DA-36-039-sc-42712

DEPARTMENT OF THE ARMY PROJECT: 3-99-11-022
SIGNAL CORPS PROJECT: 142B

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SEPTEMBER 30, 1953
<table>
<thead>
<tr>
<th>QUARTERLY REPORT NO.</th>
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<th>AUTHOR</th>
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<tbody>
<tr>
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B. J. DASHER
PROJECT DIRECTOR

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SEPT. 30, 1953-JUNE 30, 1955
TABLE OF CONTENTS

<table>
<thead>
<tr>
<th>Section</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>I. PURPOSE</td>
<td>1</td>
</tr>
<tr>
<td>II. ABSTRACT</td>
<td>2</td>
</tr>
<tr>
<td>III. CONFERENCES</td>
<td>3</td>
</tr>
<tr>
<td>IV. FACTUAL DATA</td>
<td>4</td>
</tr>
<tr>
<td>A. Introduction</td>
<td>4</td>
</tr>
<tr>
<td>B. Frequency Stability.</td>
<td>4</td>
</tr>
<tr>
<td>C. Nonlinear Elements</td>
<td>7</td>
</tr>
<tr>
<td>D. Isocline Diagrams.</td>
<td>15</td>
</tr>
<tr>
<td>E. Measurement Techniques and Circuit Representation</td>
<td>21</td>
</tr>
<tr>
<td>V. CONCLUSIONS</td>
<td>27</td>
</tr>
<tr>
<td>VI. PROGRAM FOR THE NEXT INTERVAL</td>
<td>28</td>
</tr>
<tr>
<td>VII. PERSONNEL</td>
<td>29</td>
</tr>
<tr>
<td>VIII. BIBLIOGRAPHY</td>
<td>31</td>
</tr>
</tbody>
</table>

This Report Contains 31 Pages
LIST OF FIGURES

<table>
<thead>
<tr>
<th>Figure</th>
<th>Description</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>Basic Two-Terminal Oscillator</td>
<td>5</td>
</tr>
<tr>
<td>2</td>
<td>Equivalent Circuit of Negative Resistance Oscillator</td>
<td>10</td>
</tr>
<tr>
<td>3</td>
<td>Locus of $R_{2R} + jR_{21}$ for Variable Frequency</td>
<td>12</td>
</tr>
<tr>
<td>4</td>
<td>Equivalent Circuit of Grounded-Base Transistor Oscillator</td>
<td>13</td>
</tr>
<tr>
<td>5</td>
<td>Transistor Oscillator Circuits</td>
<td>16</td>
</tr>
<tr>
<td>6</td>
<td>Idealized Volt-Ampere Characteristic</td>
<td>17</td>
</tr>
<tr>
<td>7</td>
<td>Isocline Diagram for a Small L/C Ratio for the Circuit</td>
<td>18</td>
</tr>
<tr>
<td></td>
<td>in Figure 5</td>
<td></td>
</tr>
<tr>
<td>8</td>
<td>Isocline Diagram for a Large L/C Ratio for the Circuit</td>
<td>18</td>
</tr>
<tr>
<td></td>
<td>in Figure 5</td>
<td></td>
</tr>
<tr>
<td>9</td>
<td>Amplitude and Frequency as Functions of Collector Voltage</td>
<td>20</td>
</tr>
<tr>
<td>10</td>
<td>The Effects of Changes in $Q$ and Changes in Inductance on Frequency Variations</td>
<td>20</td>
</tr>
<tr>
<td>11</td>
<td>Equivalent Circuit Involving $h$-Parameters</td>
<td>23</td>
</tr>
<tr>
<td>12</td>
<td>Equivalent Circuits and Important Relations of a Grounded-Base Stage</td>
<td>25</td>
</tr>
</tbody>
</table>
I. PURPOSE

The purpose of this project is to evolve data and information regarding transistor oscillator circuitry and the degree of reliability and reproducibility attained with transistor parameter variations. It includes both analytical and experimental studies of crystal-controlled and free-running oscillators employing transistors exclusively as active elements.

The work is divided into three principal tasks as follows:

(1) the evaluation of crystal-controlled and free-running transistor oscillators from the point of view of reliability, size, weight, circuit requirements, frequency limitations, and other similar considerations;

(2) the development of new circuits and/or techniques specifically to take advantage of the unique characteristics of transistors regarding size, weight, power requirements, etc.; and

(3) the development of practical engineering data and information to serve as a basis for designing free-running and crystal-controlled junction- and point-contact transistor oscillators in the range 0.1 to 15 mc.
II. ABSTRACT

Some basic factors that determine the frequency stability of series-mode oscillators are reviewed. It is shown that frequency stability with respect to an unstable impedance, $Z_2$, depends on the ratio $(\Delta X_1/\Delta Z_2)/(\Delta X_0/\Delta \omega)$, in which $X_0$ is the reactance of the resonator and $X_1$ is the reactance of the circuit external to the resonator. First-order effects of nonlinearity are investigated, and attention is called to the fact that $L/C$ is fundamentally more important than $Q$ as a figure of merit of the resonator in this type of oscillator. This conclusion is confirmed by both graphical analysis and laboratory experiments. Certain stability coefficients are derived for a particular transistor circuit.

Techniques for measuring transistor parameters are reviewed. A simple test set for use in the frequency range 1 kc to 1 mc has been constructed.
III. CONFERENCES

Dr. D. L. Finn, Dr. W. B. Jones, Jr., and Dr. B. J. Dasher of Georgia Tech visited the Frequency Control Branch, Squier Signal Laboratories, on July 16 and 17. Various activities of the Frequency Control Branch were observed and discussed. Basic technical requirements of the Signal Corps relative to the use of transistors in oscillators were discussed with Mr. E. Gonzales. Discussions with Mr. W. L. Doxey led to the conclusion that it would benefit both parties if the period of the contract were extended so as to permit more technical information to be included in the first reporting period, since unavoidable circumstances had delayed the project. Action on this proposal has been taken, and the period of the present contract was extended 2-1/2 months with corresponding adjustments of reporting dates.
IV. FACTUAL DATA

A. Introduction

The transistors that are presently available are subject to wide variations in characteristics from unit to unit. They are also greatly affected by changes in temperature, especially at temperatures near 50°F and above. It is generally believed that improved manufacturing techniques eventually will largely eliminate these problems. In the meantime, it is necessary to determine how these limitations affect circuit performance and to devise means to ensure stability and reproducibility of circuits employing transistors. This project is concerned with the performance and design of transistor oscillator circuits in the frequency range 0.1 mc to 15 mc.

Since there is no way of knowing what spread of transistor parameters may be expected at some particular time in the future, it is not worth much in the long run to say that a given oscillator performs well with one transistor but not with another unless the cause of failure is understood. Accordingly, during the period covered by this report, the major emphasis has been placed on a study of the fundamental factors involved in oscillator performance.

B. Frequency Stability

The frequency of an oscillator is usually determined primarily by either a resonant circuit (resonator) or a phase-shifting network. For purposes of analysis it may be assumed that the parameters of these circuits are sufficiently independent of their environment that, if they alone determined the oscillator frequency, an adequate stability would be obtained. Of course, this is not necessarily true in practice, but it is certainly reasonable to suppose the resonator to be the most stable part of a transistor oscillator. Hence, failure of an oscillator to maintain satisfactory stability is due principally either
to nonlinear effects or to changes in the parameters of circuits external to the resonator, or to both. In either case, the difficulties are usually associated with comparatively isolated portions of the circuit, for example, with the amplifier or with the load. Accordingly, an oscillator whose frequency is fixed by a resonator may be supposed to consist of three distinct parts as indicated in Figure 1.

![Figure 1. Basic Two-Terminal Oscillator.](image)

Here, \( Z_0 \) represents the resonator; \( Z_2 \) represents an element that is supposed variable; \( N \) is a coupling network; and \( Z_1 \) is the impedance of the network terminated in \( Z_2 \). Since any portion of a complete circuit may be designated \( Z_2 \), various components may be investigated one at a time and their effects combined to obtain the total effect of all variations.

In the case of a crystal-controlled oscillator for example, \( Z_0 \) represents the crystal regardless of what the rest of the circuit is like. The advantage of this two-terminal-oscillator point of view is that it separates the crystal, whose characteristics are known and supposed constant, from the part of the circuit which is considered to be inherently unstable. Thus, it is clear that, given the variable element \( Z_2 \), the essential problem is to design the network, \( N \), so as to realize the best use of the intrinsic stability of the crystal.

Or, carrying this idea further, given transistors with parameters varying over a specified range, design the network so that the over-all stability is as near as possible to that of the crystal.
The criterion for sustained oscillations in the circuit of Figure 1 is

\[ Z_0 + Z_1 = 0 \]  \hspace{1cm} (1)

and, consequently,

\[ \text{Re}[Z_0] = - \text{Re}[Z_1], \] \hspace{1cm} (2)

\[ \text{Im}[Z_0] = - \text{Im}[Z_1]. \] \hspace{1cm} (3)

Since some form of amplitude limiting is always present, equation 2 determines the amplitude of oscillations and equation 3 determines the frequency.

Equations 1, 2 and 3 are strictly correct only for a linear system because the ordinary concept of impedance is valid only for linear systems. However, if the system is almost linear, a negligible error results from their use. In one of the simplest situations, \( Z_0 \) consists of a series RLC circuit operating near its resonant frequency. Then, since the \( \text{Re}[Z_0] \) is constant, the frequency stability may be investigated by considering the imaginary components only.

Suppose that \( Z_2 \) is changed by some external means. This will result in a change of \( Z_1 \), but because of the amplitude limiting, only the imaginary part of \( Z_1 \) changes. Equilibrium must be restored according to the relation

\[ \Delta X_0 + \Delta X_1 = 0, \] \hspace{1cm} (4)

in which \( X_0 \) and \( X_1 \) are the imaginary parts of \( Z_0 \) and \( Z_1 \), respectively. In terms of the assumed variables, equation 4 becomes

\[ \frac{\partial X_0}{\partial \omega} \Delta \omega + \frac{\partial X_1}{\partial Z_2} \Delta Z_2 + \frac{\partial X_1}{\partial \omega} \Delta \omega = 0 \] \hspace{1cm} (5)

from which it follows that

\[ \frac{\partial X_1}{\partial Z_2} = \frac{-\frac{\partial X_1}{\partial \omega}}{\frac{\partial X_0}{\partial \omega} + \frac{\partial X_1}{\partial \omega}} \] \hspace{1cm} (6)
Let $D_2 = \frac{\Delta \omega}{\Delta Z_2}$, which is a measure of the absolute frequency stability (more precisely, the absolute frequency instability) with respect to changes of $Z_2$.

Also, let $K_{12} = \frac{\partial X_1}{\partial Z_2}$, $K_{01} = \frac{\partial X_0}{\partial \omega}$, $K_{11} = \frac{\partial X_1}{\partial \omega}$. In terms of these definitions, equation 6 becomes

$$D_2 = \frac{-K_{12}}{K_{01} + K_{11}}.$$  

This result emphasizes the fact that the over-all stability depends jointly on the properties of the resonator and the remainder of the circuit. In the situation of interest here, $K_{01} > K_{11}$ and, therefore,

$$D_2 \approx \frac{K_{12}}{K_{01}}.$$  

Moreover, $K_{01}$ must be regarded as fixed, and so it becomes evident that the circuit design problem is to make $K_{12}$ small without upsetting equation 2.

If a certain circuit modification multiplies $K_{12}$ by some factor, then, so far as $D_2$ is concerned, this is equivalent to dividing $K_{01}$ by the same factor. For example, the bridge arrangement of the Meacham "bridge-stabilized oscillator" is usually considered to produce an increase of the "effective Q" of the crystal. Physically, however, the crystal is not changed, and according to the above argument, it should be quite as satisfactory to think of the bridge as improving the amplifier instead of the crystal. It remains to be seen whether the latter point of view will prove really useful.

C. Nonlinear Elements

The frequency stability of an oscillator is significantly affected by the nonlinear behavior of elements in the oscillator circuit. Therefore, it is to be expected that a linear analysis of an oscillator circuit cannot provide a complete description of the variation of operating frequency. In spite of this fact, it is found that the first-order effects on frequency stability as...
described by a linear analysis are adequate for most oscillators which are highly stable in frequency and nearly sinusoidal in operation.

The first step in the linear analysis of an oscillator is the selection of linear parameters which give an adequate mathematical description of the performance of the oscillator. It is often convenient, but not necessary, to draw an equivalent circuit consisting of various resistances, inductances, capacitances and sources to represent these parameters. The linear parameters are defined in terms of the voltages and currents of fundamental frequency in the oscillator. Thus, the driving-point impedance of a two-terminal element is defined as the complex ratio of fundamental voltage to fundamental current. This impedance is a complicated function of both the voltage and current waveshape and the frequency for the element. It can be specified and measured only under operating conditions for the circuit.

According to the definition of impedance given above, a nonlinear pure resistor may be described by a linear impedance having both a real and an imaginary part. This is possible because the fundamental component of voltage in a nonlinear resistor may be out of phase with the fundamental component of current. As mentioned above, this impedance is a complicated function of both the voltage and the current of the resistor.

If the nonlinear element being analyzed has more than two terminals, the linear parameters selected to describe the operation cannot, in general, be specified uniquely because the principle of superposition is no longer applicable. In this case, the selection of the parameters may be based on measurements taken when the element is operating under as nearly linear conditions as possible.
The linear parameters which are selected may be used to write a set of homogeneous algebraic equations describing the steady-state performance of the oscillator system. The condition for self-sustained oscillations may be obtained by setting the determinant of this set of equations equal to zero. Two equations of equilibrium result from this procedure. One is obtained by setting the real part of the determinant equal to zero; the other is obtained by setting the imaginary part of the determinant equal to zero. The oscillator frequency may be obtained from these equations of equilibrium if all other parameter values are known. (However, no information may be derived concerning the amplitude of the oscillations.) The closeness to which the predicted frequency approximates the actual frequency is dependent upon the accuracy with which the linear mathematical system describes the performance of the oscillator system.

The variation of frequency produced by a variation in a given parameter T of the circuit may be obtained from the two equations of equilibrium. All parameters other than the frequency of oscillation, \( \omega \), the real part, \( T_r \), of the variable parameter, and the imaginary part, \( T_i \), of the variable parameter are held fixed. The two equations of equilibrium may then be written as

\[
\begin{align*}
f_1(\omega, T_r, T_i) &= 0, \\
f_2(\omega, T_r, T_i) &= 0.
\end{align*}
\]

The frequency variation of the circuit may be obtained through determining the locus of \( T_r + jT_i \) for variable frequency. The equation of this locus may be obtained by eliminating \( \omega \) from equations 9 and 10.

The relative frequency stability of the oscillator with respect to variations in T may be defined as

\[
S_f = \frac{\omega}{|T|} \lim_{\Delta \omega \to 0} \frac{\Delta T}{\Delta \omega} = \frac{\omega}{|T|} \frac{d|T|}{d\omega} = \frac{d|T|}{\frac{dT}{d\omega}}. \tag{11}
\]
Here, \( \frac{d|T|}{d\omega} \) is the rate of change of arc length along the locus of the complex variable \( T \) with respect to frequency \( \omega \). The relative frequency stability, \( S_f \), is taken to be a measure of the tendency of the oscillator system to resist changes in frequency.

The simple negative resistance oscillator shown in Figure 2 may be analyzed to provide a demonstration of the procedure described above.

![Figure 2. Equivalent Circuit of Negative Resistance Oscillator.](image)

The elements \( R_1, L \) and \( C \) are assumed to be linear in operation, and the linear parameters describing their operation are assumed to be the impedances \( R_1j\omega L \) and \(-j/\omega C\), respectively. It is assumed that the element \( R_2 \) may be nonlinear and may vary in a random manner. The linear parameter describing the operation of the element \( R_2 \) is taken to be the impedance \( R_{2r} + jR_{2i} \). Any possible relationship between the fundamental voltage and fundamental current of \( R_2 \) may be described by proper values of the real and imaginary parts of this impedance.

The equation describing the operation of this circuit is

\[
I \left( R_{2r} + jR_{2i} + R_1 + j\omega L - j/\omega C \right) = 0. \tag{12}
\]

The condition for sustained oscillations is

\[
R_{2r} + jR_{2i} + R_1 + j\omega L - j/\omega C = 0. \tag{13}
\]
The real and imaginary parts of equation 13 may be set equal to zero to provide two equations of equilibrium:

\[ R_{2r} + R_1 = 0, \quad (14) \]
\[ R_{2i} + \omega L - 1/\omega C = 0. \quad (15) \]

Finally, equations 14 and 15 may be rewritten as

\[ R_{2r} = -R_1; \quad (16) \]
\[ \omega = \sqrt{\frac{1}{LC} + \left(\frac{R_{2i}}{2L}\right)^2} - \frac{R_{2i}}{2L}. \quad (17) \]

It is evident from equation 16 that the real part of the impedance describing \( R_2 \) must be negative and exactly equal in magnitude to \( R_1 \). If the element \( R_2 \) is not able to adjust itself to meet this condition by varying the waveforms and frequency of the voltage and current in the circuit, then the oscillator will cease operation. The operating frequency is identical with the series resonant frequency of the \( R_1LC \) combination only if \( R_{2i} = 0 \). This special case requires that the fundamental component of current in the element \( R_2 \) be exactly 180° out of phase with the fundamental component of voltage.

The locus of the variable parameter \( R_{2r} + jR_{2i} \) as found by use of equations 14 and 15 is shown in Figure 3. This locus is exactly the negative of the locus of the impedance of the \( R_1LC \) series combination for variable frequency.

The relative frequency stability of the oscillator, evaluated at the series resonant frequency of the \( R_1LC \) combination, is found by differentiating equation 15:

\[ \frac{\omega}{R_2} \frac{d|R_2|}{d\omega} = \frac{1}{R_2\sqrt{LC}} \left( L + \frac{1}{\omega^2 C} \right) \omega = \frac{1}{\sqrt{LC}} \]

or,

\[ S_f = \frac{\omega}{R_2} \frac{d|R_2|}{d\omega} = \frac{2}{R_1} \frac{\sqrt{L}}{\sqrt{C}} = 2Q. \quad (19) \]
Equation 18 apparently shows that frequency stability does not depend on the value of the resonator resistance $R_1$. Equation 19 seems to indicate that frequency stability is very much dependent on the value of $R_1$. The question as to which result gives a more accurate picture depends on the nature of the response of the nonlinear circuit to the basic variation, such as temperature or supply voltage, which is actually causing the frequency deviation. Equation 19 should be used if a given increment in the primary variable, such as temperature, etc., tends to produce a constant phase shift in the impedance describing the fundamental components of the nonlinear resistance $R_2$. Equation 18 should be used if this given increment in the primary variable produces a given change in the imaginary component of the impedance describing the resistor $R_2$, regardless of the value of the resistance $R_2$. In some situations perhaps neither definition gives an accurate description of the nature of the nonlinear element. However, it is very important to note that in both of these cases the frequency
stability can be improved by increasing the L/C ratio. Therefore, it seems reasonable to suppose that an increase in this ratio results in an increase in frequency stability regardless of the nature of the nonlinear resistor used in the circuit. It seems just as important to emphasize that a decrease in the value of the resonator resistance, or a corresponding increase in the value of Q, does not necessarily result in an increase in frequency stability. Whether the stability improves or not depends upon the type of negative resistance element used in the oscillator and the conditions under which it is operated.

A more complicated analysis problem is the grounded-base transistor oscillator circuit shown in Figure 4.

Figure 4. Equivalent Circuit of Grounded-Base Transistor Oscillator.

A set of homogeneous equations describing the steady-state performance of this circuit is

\[ 0 = I_c (r_b + R_b) + I_e (j\omega L - j/\omega C + r_b + R_b + r_e + R_e) \]
\[ 0 = I_c (r_c + R_c + r_b + R_b) + I_e (r_b + R_b + ar_c) \]  \( (20) \)

The condition for sustained oscillations is obtained by setting the determinant of this set of equations equal to zero.

\[ 0 = (j\omega L - j/\omega C + r_e + R_e)(r_c + R_c + r_b + R_b) + (r_b + R_b) [R_c + (1 - a)r_c] \]  \( (21) \)
For perfectly linear operation, setting the real and imaginary parts, respectively, equal to zero yields the following two constraints on the operation of the circuit.

\[
\begin{align*}
\omega &= \frac{1}{\sqrt{LC}} \\
a &= \left(1 + \frac{r_e + R_e}{r_b + R_b}\right) \left(1 + \frac{R_c}{r_e}\right) + \frac{r_e + R_e}{r_c}
\end{align*}
\]  \hspace{1cm} (22)

The relative frequency stability of the oscillator with respect to variation in any parameter may be determined by use of equation 21. For instance, suppose that all parameters remain fixed except \(r_b\). The parameter \(r_b\) is allowed to assume complex values \(r_{br} + jr_{bi}\) in keeping with the constraints imposed by equation 21. These constraints may be stated explicitly by setting the real and imaginary parts of equation 21 equal to zero.

\[
0 = (r_{br} + R_b)[r_e + R_e + R_c + (1 - a)r_c] + (r_e + R_e)(r_c + R_c) - r_{bi}(\omega L - \frac{1}{\omega C}) \hspace{1cm} (23)
\]

\[
0 = r_{bi}[r_e + R_e + R_c + (1 - a)r_c] + (r_c + R_c + r_{br} + R_b)(\omega L - \frac{1}{\omega C}) \hspace{1cm} (24)
\]

The equation of the locus of \(r_{br} + jr_{bi}\) for variable frequency may be found by eliminating \(\omega\) from equations 23 and 24.

\[
0 = r_{bi}^2[r_e + R_e + R_c + (1 - a)r_c] + \left\{(r_{br} + R_b)[r_e + R_e + R_c + (1 - a)r_c] + (r_e + R_e)(r_c + R_c + r_{br} + R_b)\right\}(r_c + R_c + r_{br} + R_b) \hspace{1cm} (25)
\]

This locus is seen to be an ellipse with focii on the \(r_{br}\) axis.

The relative frequency stability with respect to variations in \(r_b\) calculated at the frequency \(\omega = \frac{1}{\sqrt{LC}}\) is found from equation 23.

\[
S_r = \frac{\omega}{r_b} \frac{d|r_b|}{d\omega} = \frac{1}{\sqrt{LC}} \frac{1}{(r_e + R_e)} \left(1 + \frac{R_b}{r_b}\right) \left(1 + \frac{r_b + R_b}{r_c + R_c}\right) \hspace{1cm} (26)
\]
A similar analysis may be carried out to determine the relative frequency stability with respect to variations in the other parameters. The results of an analysis of this type are shown in Table I.

TABLE I

NORMALIZED RELATIVE FREQUENCY STABILITY OF THE TRANSISTOR OSCILLATOR SHOWN IN FIGURE 4 CALCULATED AT THE FREQUENCY $\omega = 1/\sqrt{LC}$

<table>
<thead>
<tr>
<th>Variable</th>
<th>$S_f$</th>
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<tr>
<td>$r_b$</td>
<td>$2\sqrt{L/C} \frac{1}{(r_e + R_e)} \left(1 + \frac{R_b}{r_b} \right) \left(1 + \frac{r_b + R_b}{r_c + R_c} \right)$</td>
</tr>
<tr>
<td>$r_c$</td>
<td>$2\sqrt{L/C} \frac{1}{(r_e + R_e)} \left(1 + \frac{r_c + R_c}{r_b + R_b} \right) \left(1 + \frac{r_b + R_b}{r_e + R_e} \right)$</td>
</tr>
<tr>
<td>$r_e$</td>
<td>$2\sqrt{L/C} \frac{1}{r_e}$</td>
</tr>
<tr>
<td>$a$</td>
<td>$2\sqrt{L/C} a(r_b + R_b) \left(1 + \frac{R_c}{r_c} + \frac{r_b + R_b}{r_c} \right)$</td>
</tr>
</tbody>
</table>

D. Isocline Diagrams

It was pointed out in the preceding section that, in so far as frequency stability alone is concerned, increasing the Q of the resonator of a series-mode oscillator does not necessarily result in improved performance. Rather, in this case, it is the L/C ratio that is basically more important. Although this fact has been recognized for a long time, there persists a popular notion that high
Q and a high degree of stability are inseparable. For this reason it is worthwhile to include here some further analytical and experimental results. A word of warning is likewise appropriate. High-Q resonators are often desirable for other reasons, and it must also be remembered that these results apply specifically only to circuits of the series-mode type.

It is shown in equation 22 of the preceding section that the circuit of Figure 5 will oscillate, provided

$$a > 1 + \frac{R + r_e}{R_b + r_b} \left[ \left( 1 + \frac{R_c}{r_c} \right) + \frac{r_e + R}{r_c} \right].$$

(27)

If $r_c$ is very large compared to the other resistances, equation 27 yields the approximation

$$a > 1 + \frac{R + r_e}{R_b + r_b}.$$

(28)

Thus, if the emitter-circuit resistance is sufficiently low or the base-circuit resistance sufficiently high, the circuit will oscillate.
The apparent resistance in series with the resonant circuit is negative when inequality 28 holds. An idealized static V-I characteristic of the circuit to the right of points 1 and 2 in Figure 5 is shown in Figure 6, in which the static operating point has been taken as the origin.

Figure 6. Idealized Volt-Ampere Characteristic.

When the resistance $R$ is taken into account, the combined characteristic appears as in Figure 7. If this characteristic has a region of negative slope, the circuit will oscillate at a frequency determined primarily by $L$ and $C$. Figure 7 also constitutes an isocline diagram by means of which the current and voltage waveshapes of the circuit may be found. For details of this procedure the reader is referred to the literature (for example, Edson, Vacuum Tube Oscillators, John Wiley, 1953). Briefly, the isocline diagram is a graphical, parametric representation of the equation for voltage equilibrium, which reads

$$F(i) + Ri + L \frac{di}{dt} + \int \frac{idt}{C} = 0 ,$$

(29)
Figure 7. Isocline Diagram for a Small L/C Ratio for the Circuit in Figure 5.

Figure 8. Isocline Diagram for a Large L/C Ratio for the Circuit in Figure 5.
F(i) representing the transistor characteristic. Eliminating t from equation 29 gives

\[
\frac{dv}{dt} + \frac{L}{C} v + \frac{1}{v + F(i) + Ri} = 0. \tag{30}
\]

A closed path that satisfies equation 30 is called a limit cycle and represents steady-state oscillations. The limit cycle may be determined from the isocline diagram, after which the desired waveshapes may be found.

Again referring to Figure 7, a limit cycle and the corresponding current wave are shown. Figure 8 shows the result of increasing the L/C ratio. Notice that the distortion is greatly reduced and that the operating path is more nearly confined to the region of negative slope. The relative frequency stability is not obvious, but a careful analysis shows that the system of Figure 7 is more sensitive to small changes of the F(i) characteristic than that of Figure 8. Also, it may be seen from equation 30 that decreasing R has the same qualitative effect as decreasing L/C. Thus, an increase of Q will result in an improvement or an impairment of frequency stability depending upon how the increase is obtained.

Some experimental data have been taken on laboratory oscillators of the type shown in Figure 5, using Western Electric type-1768 point-contact transistors. The influence of harmonic content on frequency stability is apparent from the curves of Figure 9. As the collector voltage, \( V_c \), is increased to approximately 5.5 volts, the amplitude of oscillation increases almost linearly. In this range the waveshape is nearly sinusoidal, and the frequency is affected very little by changes in amplitude. For higher values of \( V_c \), the circuit becomes more nonlinear, the harmonic content of the current increases sharply, and its effect on frequency is apparent. For still higher values of \( V_c \), the amplitude and harmonic content become less sensitive to small changes in \( V_c \).
Figure 9. Amplitude and Frequency as Functions of Collector Voltage

A: \( Q = 20, L = 10\text{mH} \)
B: \( Q = 20, L = 20\text{mH} \)
C: \( Q = 10, L = 20\text{mH} \)
\( f_0 = 50 \text{kc} \)

Figure 10. The Effects of Changes in \( Q \) and Changes in Inductance on Frequency Variations.
The frequency is therefore also less sensitive to changes of $V_c$ in this region. Figure 10, which shows the variation in frequency as $V_c$ is changed, also shows the relative effects of changing Q and changing the L/C ratio. These results entirely confirm the analysis presented above.

E. Measurement Techniques and Circuit Representation

One of the main objectives of this project is the design of oscillator circuits at high frequencies based upon known transistor parameters. Consequently, the problem of the choice and measurement of such parameters immediately arises. In general, the measurements of interest are the low- and high-frequency small-signal parameters, the static curves, and special characteristics which may be important for special transistor applications. A preliminary survey of existing measurement techniques shows that test sets of varying complexity have been designed to make these measurements. The small-signal parameters usually involve the current gain $\alpha$ and $(1 - \alpha)$ for junction transistors, and the open circuit impedances $Z_{11}$, $Z_{12}$, $Z_{21}$ and $Z_{22}$ of the transistor when treated as an active linear four-pole network. If preferred, the actual elements of the T (or $\pi$) equivalent circuit, namely, $Z_e$, $Z_b$, and $Z_c$ (or $Y_{eb}$, $Y_{ce}$, and $Y_{cb}$), may be measured directly. These measurements are relatively easy to make at audio frequencies, but they are difficult to make at high frequencies where the simple T and $\pi$ equivalent circuits no longer represent the transistor accurately, and transit time and internal reactance effects cannot be neglected.

In large-signal or nonlinear applications of the transistor, static curves are useful if not entirely necessary. Point-by-point methods of obtaining such curves have been replaced by rapid-sweep methods which present the desired information as a family of curves displayed on a CRO or as actual plots on a curve plotter.
The analysis of transistor circuits from the matrix viewpoint is becoming increasingly popular. This method conveniently eliminates the necessity of first replacing the transistor by its equivalent circuit. The transistor is treated as an active linear four-pole network so that its electrical behavior is completely described by two equations which relate the input and output currents and voltages. Since only two of the four variables are independent, the two equations can be written in a maximum of six permutations. Three of these—those involving the open-circuit impedance (z) parameters, the short-circuit admittance (y) parameters, and the hybrid (h) parameters—have been found useful in connection with transistor analysis and measurement.

The h-parameters of a grounded-base stage have been selected as the parameters to be used as the basis for circuit analysis and design. They are identified by the equations

\[
\begin{align*}
v_1 &= i_1(h_{11}) + v_2(h_{12}) , \\
i_2 &= i_1(h_{21}) + v_2(h_{22})
\end{align*}
\]

where

\[
\begin{align*}
h_{11} &= \frac{\partial v_1}{\partial i_1} v_2 = \text{constant} & h_{12} &= \frac{\partial v_1}{\partial v_2} i_1 = \text{constant} \\
h_{21} &= \frac{\partial i_2}{\partial i_1} v_2 = \text{constant} & h_{22} &= \frac{\partial i_2}{\partial v_2} i_1 = \text{constant} .
\end{align*}
\]

The h-parameters require both open-circuit input and short-circuit output conditions. This is advantageous from the measurement-and-circuit-operation point of view because the transistor is inherently a low-input impedance, high-output impedance device in the more common circuit applications. Therefore, the
required terminal conditions can be easily achieved so that the h-parameters can readily be measured with existing measuring equipment.

Figure 11 shows the equivalent circuit of a grounded-base stage whose identifying equations are

\[
\begin{align*}
\v_{eb} &= i_e(h_{11}) + v_{cb}(h_{12}), \\
\i_c &= i_e(h_{21}) + v_{cb}(h_{22}).
\end{align*}
\]

Figure 11. Equivalent Circuit Involving h-Parameters.

The h-parameters in equation 33 are related to the familiar z- and y-parameters in the following manner:

\[
\begin{align*}
h_{11} &= \frac{h_z}{z_{22}} = \frac{1}{y_{11}}, \\
h_{12} &= \frac{z_{12}}{z_{22}} = -\frac{y_{12}}{y_{11}}, \\
h_{21} &= -\frac{z_{21}}{z_{22}} = \frac{y_{21}}{y_{11}}, \\
h_{22} &= \frac{1}{z_{22}} = \frac{\Lambda_y}{y_{11}}.
\end{align*}
\]
where

\[ \Delta_z = z_{11}z_{22} - z_{12}z_{21} ; \Delta_y = y_{11}y_{22} - y_{12}y_{21} . \]

Also,

\[ a = -h_{21} = -\frac{y_{21}}{y_{11}} = \frac{z_{21}}{z_{22}} . \quad (35) \]

Examination of equations 33 and 34 reveals the fact that the h-parameters may be used directly in matrix analysis without conversion. Moreover, they provide information on the important factors affecting transistor performance. For example, \( h_{21} \) is the negative of the current amplification factor \( a \); \( h_{12} \) is the reverse voltage amplification factor and provides information on internal feedback performance; \( h_{11} \) and \( h_{22} \) are useful for impedance matching. Also, the h-parameters are said to be more nearly independent of each other than the more conventional equivalent circuit parameters. The so-called "active mutual" impedance \( z_m \) (or admittance \( y_m \)) found in the T (or \( \pi \)) equivalent circuit has no counterpart in the h-parameter equivalent circuit. Finally, the fact that the h-parameters can be readily transformed into the z- or y-parameters by means of simple algebraic relations adds to their usefulness. Figure 12 gives a tabulation of the important relations of the grounded-base z-, y-, and h-parameter equivalent circuits. Incidentally, the h-parameters have recently been adopted by Bell Telephone Laboratories, the General Electric Company and the Western Electric Company for the specification of their transistors.

The current amplification factor, \( a \), has been defined as

\[ a = -\frac{\partial i_c}{\partial i_e} \biggm|_{v_{cb} = \text{constant}} , \quad (36) \]

so that at any one operating point and frequency \( a \) remains constant regardless of the way in which the transistor is connected.
### Z-PARAMETER

- \( v_{eb} = i_e (z_e + z_b) + i_c (z_b) \)
- \( v_{cb} = i_e (z_b + z_m) + i_c (z_b + z_c) \)

### Y-PARAMETER

- \( i_e = v_{eb} (y_{eb} + y_{ce}) + v_{cb} (-y_{ce}) \)
- \( i_c = v_{eb} (y_{m} - y_{ce}) + v_{cb} (y_{cb} + y_{ce}) \)

### H-PARAMETER

- \( i_e = v_{eb} (h_{11}) + v_{cb} (h_{12}) \)
- \( i_c = i_e (h_{21}) + v_{cb} (h_{22}) \)

### GENERAL FORMULAE

- \( \alpha = \frac{\partial i_c}{\partial i_e} \ |_{v_{cb} = \text{constant}} \)
- \( \alpha = \frac{z_{21}}{z_{22}} = \frac{y_{21}}{y_{11}} = -h_{21} \)

### Z → H TRANSFORMATIONS

- \( z_{11} = h_{11} - (h_{21}) \frac{h_{12}}{h_{22}} \)
- \( z_{12} = \frac{h_{12}}{h_{22}} \)
- \( z_{21} = \frac{h_{21}}{h_{22}} \)
- \( z_{22} = \frac{1}{h_{22}} \)

### Y → H TRANSFORMATIONS

- \( y_{11} = \frac{1}{h_{11}} \)
- \( y_{12} = -\frac{h_{12}}{h_{11}} \)
- \( y_{21} = \frac{h_{21}}{h_{11}} \)
- \( y_{22} = h_{22} - (h_{21}) \frac{h_{12}}{h_{11}} \)

---

Figure 12. Equivalent Circuits and Important Relations of a Grounded-Base Stage.
Figure 13 shows the basic circuits of a test set constructed for measuring the small-signal $h$-parameters and $(1 - a)$ for frequencies from 1 kc to 1 mc.

![Circuit Diagram]

**A.** Measures $h_{11}$, $h_{21}$, $(1 - a)$.

**B.** Measures $h_{12}$, $h_{22}$.

Figure 13. Basic Circuits for Measuring $h$-Parameters.

The dc bias supply is not shown. Both circuits are incorporated into a single unit, and switching arrangements have been made to make measurements easily and quickly. The parameter values are read as voltage ratios across precision resistors $R_1$, $R_2$, and $R_3$, using a sensitive vacuum-tube voltmeter. It is felt that the work done so far on parameter measurements will lead to adequate measurements at much higher frequencies.
V. CONCLUSIONS

The frequency stability of even simple oscillator circuits depends jointly upon the resonator and the amplifier (negative impedance). Since stability coefficients computed as phase or reactance slopes do not take the amplifier into consideration, they do not give a complete representation of the stability problem. Further investigation is necessary to determine the significance of this observation.

The h-parameter equivalent circuit appears to give the most convenient representation of transistors. The h-parameters are relatively easy to measure, and the elements of the T or \( \pi \) equivalent circuits may be computed from them by means of simple conversion formulas.
VI. PROGRAM FOR THE NEXT INTERVAL

1. Analytical studies will be continued.

2. Several basic oscillator circuits will be designed and tested.

3. Studies of techniques for measurement of transistor parameters will be continued with special emphasis on high-frequency problems.
VII. PERSONNEL

The key personnel working on this project during the period covered by this report are:

<table>
<thead>
<tr>
<th>Name</th>
<th>Position</th>
<th>Approximate Hours Worked</th>
</tr>
</thead>
<tbody>
<tr>
<td>B. J. Dasher, Sc.D.</td>
<td>Project Director</td>
<td>325</td>
</tr>
<tr>
<td>D. L. Finn, Ph.D.</td>
<td>Research Associate</td>
<td>250</td>
</tr>
<tr>
<td>W. B. Jones, Jr., Ph.D.</td>
<td>Research Associate</td>
<td>175</td>
</tr>
<tr>
<td>Wai Mun Syn, M.S.</td>
<td>Research Assistant</td>
<td>350</td>
</tr>
</tbody>
</table>

Dr. Dasher is Professor of Electrical Engineering and Research Associate, Engineering Experiment Station, Georgia Institute of Technology. He received the degrees of B.S. and M.S. in E.E. from Georgia Tech in 1935 and 1945, respectively, and the Sc.D. in E.E. from MIT in 1952. From 1940 to 1946 he taught electrical engineering at Georgia Tech, and from 1946 to 1951 he was first Instructor and then Assistant Professor of Electrical Communications at MIT, returning to Georgia Tech in 1951. From 1949 to 1951 he was also a Group Leader of the Telemetering and Instrumentation Group, Project Meteor, Research Laboratory of Electronics, MIT. His fields of special interest include advanced network theory, vacuum tube electronics (circuit design and analysis), applied mathematics, and communication theory.

Dr. Finn is Associate Professor of Electrical Engineering and Research Associate, Engineering Experiment Station, Georgia Institute of Technology. He received the B.S. and M.S. degrees from Purdue University in 1943 and 1948, respectively, and the Ph.D. degree in E.E., also from Purdue University, in 1952. From 1943 to 1946 he was a radar technician in the A.A.F., and from 1948 to 1952 he was a Teaching Assistant in the Electrical Engineering Department, Purdue
University. He joined the faculty of Georgia Tech in 1952. His fields of special interest include advanced network theory, applied mathematics, electronic circuitry, and communication theory.

Dr. Jones is Associate Professor of Electrical Engineering and Research Associate, Engineering Experiment Station, Georgia Institute of Technology. He received his B.S. and M.S. degrees in E.E. from the Georgia Institute of Technology in 1945 and 1948, respectively, and the Ph.D. degree in E.E. in 1953. He has been engaged in teaching and research at Georgia Tech since 1948. His special interests include oscillators, electronic circuits, and digital computers.

Mr. Syn received the degrees B.E.E. and M.S.E.E. from the Georgia Institute of Technology in 1951 and 1952, respectively, and is currently continuing his graduate studies. He has had experience teaching electronics and mathematics. During 1952-1953 he worked as a Research Assistant at the Engineering Experiment Station on Project No. 171-118.
VIII. BIBLIOGRAPHY


QUARTERLY REPORT NO. 2
PROJECT NO. 236-206

TRANSISTOR OSCILLATORS OF EXTENDED-FREQUENCY RANGE

by
B. J. DASHER, D. L. FINN
and
W. B. JONES, JR.

CONTRACT NO. DA-36-039-sc-42712

DEPARTMENT OF THE ARMY PROJECT: 3-99-11-022
SIGNAL CORPS PROJECT: 142B

December 31, 1953
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- o - o - o - o -

December 31, 1953
TABLE OF CONTENTS

<table>
<thead>
<tr>
<th>Section</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>I. PURPOSE</td>
<td>1</td>
</tr>
<tr>
<td>II. ABSTRACT</td>
<td>2</td>
</tr>
<tr>
<td>III. CONFERENCES</td>
<td>3</td>
</tr>
<tr>
<td>IV. FACTUAL DATA</td>
<td>4</td>
</tr>
<tr>
<td>A. Introduction</td>
<td>4</td>
</tr>
<tr>
<td>B. Almost-Linear Series-Mode Oscillator</td>
<td>6</td>
</tr>
<tr>
<td>C. Analysis of Elementary Negative-Resistance Transistor Oscillator</td>
<td>18</td>
</tr>
<tr>
<td>D. Experimental Data</td>
<td>28</td>
</tr>
<tr>
<td>E. Measurement Techniques</td>
<td>34</td>
</tr>
<tr>
<td>V. CONCLUSIONS</td>
<td>37</td>
</tr>
<tr>
<td>VI. PROGRAM FOR THE NEXT INTERVAL</td>
<td>38</td>
</tr>
<tr>
<td>VII. PERSONNEL</td>
<td>39</td>
</tr>
<tr>
<td>VIII. BIBLIOGRAPHY</td>
<td>40</td>
</tr>
</tbody>
</table>

This Report Contains 40 Pages
<table>
<thead>
<tr>
<th>Figure</th>
<th>Description</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.</td>
<td>Basic Two-Terminal Oscillator</td>
<td>6</td>
</tr>
<tr>
<td>2.</td>
<td>Typical Negative-Resistance Characteristic</td>
<td>8</td>
</tr>
<tr>
<td>3.</td>
<td>Circuit for Obtaining Negative Resistance</td>
<td>8</td>
</tr>
<tr>
<td>4.</td>
<td>Circuit for Studying Input Impedance</td>
<td>9</td>
</tr>
<tr>
<td>5.</td>
<td>Variation of $R_1$ as a Function of Collector Supply Voltage</td>
<td>10</td>
</tr>
<tr>
<td>6.</td>
<td>Characteristics of the Oscillator in Figure 7</td>
<td>15</td>
</tr>
<tr>
<td>7.</td>
<td>Circuit for Checking Groszkowski's Equation</td>
<td>16</td>
</tr>
<tr>
<td>8.</td>
<td>Circuit for Reducing Second-Harmonic Distortion</td>
<td>16</td>
</tr>
<tr>
<td>9.</td>
<td>Characteristics of the Oscillator in Figure 8</td>
<td>17</td>
</tr>
<tr>
<td>10.</td>
<td>Basic Configuration for Negative-Resistance Oscillator Using Point-Contact Transistor</td>
<td>18</td>
</tr>
<tr>
<td>11.</td>
<td>Equivalent Circuit of Tuned-Emitter Oscillator</td>
<td>19</td>
</tr>
<tr>
<td>12.</td>
<td>Equivalent Circuit of Tuned-Collector Oscillator</td>
<td>21</td>
</tr>
<tr>
<td>13.</td>
<td>Tuned-Emitter Oscillator Circuit</td>
<td>25</td>
</tr>
<tr>
<td>14.</td>
<td>Tuned-Collector Oscillator Circuit</td>
<td>26</td>
</tr>
<tr>
<td>15.</td>
<td>Performance of the Circuit in Figure 13</td>
<td>26</td>
</tr>
<tr>
<td>16.</td>
<td>Performance of the Circuit in Figure 14</td>
<td>27</td>
</tr>
<tr>
<td>17.</td>
<td>The Effect of Signal Current on $R_1$ and $H_2$, $R_c = 1k$</td>
<td>30</td>
</tr>
<tr>
<td>18.</td>
<td>The Effect of Signal Current on $R_1$ and $H_2$, $R_c = 2.7k$</td>
<td>30</td>
</tr>
<tr>
<td>19.</td>
<td>An External Limiter Reduces Distortion</td>
<td>32</td>
</tr>
<tr>
<td>20.</td>
<td>Variation of $R_1$ and $H_2$ with Emitter Bias</td>
<td>33</td>
</tr>
<tr>
<td>21.</td>
<td>Variation of $R_1$ and $H_2$ with Collector Bias</td>
<td>33</td>
</tr>
<tr>
<td>22.</td>
<td>Vector Diagram Representing Equation 20</td>
<td>35</td>
</tr>
</tbody>
</table>
I. PURPOSE

The purpose of this project is to evolve data and information regarding transistor-oscillator circuitry and the degree of reliability and reproducibility attained with transistor parameter variations. It includes both analytical and experimental studies of crystal-controlled and free-running oscillators employing transitors exclusively as active elements.

The work is divided into three principal tasks as follows:

(1) The evaluation of crystal-controlled and free-running transistor oscillators from the point of view of reliability, size, weight, circuit requirements, frequency limitations and other similar considerations;

(2) the development of new circuits and/or techniques specifically to take advantage of the unique characteristics of transitors regarding size, weight, power requirements, etc.; and

(3) The development of practical engineering data and information to serve as a basis for designing free-running and crystal-controlled junction- and point-contact transistor oscillators in the range 0.1 to 15 mc.
II. ABSTRACT

The operation of series-mode oscillators of the negative-resistance type is discussed in detail. The relation between amplitude stability and distortion is considered, and it is shown that distortion is necessary to amplitude stability in this kind of oscillator. Frequency "pulling" due to distortion can be computed by means of Groszkowski's equation. Tests on simple transistor oscillators at 1,000 cps show a fairly good agreement between computed and measured frequency.

Frequency stability is also considered from the point of view of linear circuit theory. Equations relating the frequency change due to small changes in the reactance of various transistor parameters are derived. The results agree qualitatively with experiments.

Experimental curves showing the way in which some of the more significant parameters of series-mode circuits vary with supply voltages are presented.
III. CONFERENCES

Dr. B. J. Dasher attended the "Symposium on Applications of Transistors to Military Equipment" held at Yale University on September 2 and 3, 1953.

Professor K. S. Van Dyke, Mr. R. A. Sykes and Dr. E. A. Gerber visited Georgia Tech during September. The objectives of the project and the current status of the work were discussed with them.

Mr. Leyton, of Squier Laboratories, visited Georgia Tech in October, primarily in connection with another SCEL project. The work of this project was outlined briefly to him.
IV. FACTUAL DATA

A. Introduction

Some of the basic factors affecting the frequency stability of transistor oscillators were discussed in Quarterly Report No. 1. Because of the simplicity of two-terminal oscillators, the discussion was mostly confined to oscillators of this type. A simple two-terminal transistor oscillator consists of a point-contact transistor with a series-resonant circuit connected from emitter to ground. The base is grounded through a resistor and the collector is fed from a source having low impedance at the operating frequency. However, it is possible to treat any oscillator as a two-terminal oscillator, and it was suggested that this point of view may be especially useful in the case of crystal-controlled oscillators because the crystal is itself a two-terminal element that would, ideally, completely control the frequency. When this is done the circuit external to the crystal, whatever its arrangement, is treated as another two-terminal element. The outstanding characteristic of this element is that it behaves like a negative resistance over a certain range of currents and voltages. If the voltage is determined by the current, i.e., the voltage is a single-valued function of current, the negative resistance is said to be current controlled; and if the opposite is true, i.e., the current is a single-valued function of voltage, the negative resistance is said to be voltage controlled. An oscillator that employs a current-controlled negative resistance is usually associated with a series-resonant circuit and is called a series-mode oscillator, and one that uses a voltage-controlled negative resistance (more accurately, negative conductance) is usually associated with a parallel-resonant circuit and is called a parallel-mode oscillator.
The parallel-mode oscillator is the dual of the series-mode oscillator and vice versa. Consequently, the essential characteristics of parallel-mode oscillators may be deduced from those of series-mode oscillators by the principles of duality. Some circuits do not fall clearly into either class, and, of course, quartz crystals may be used in both types. Still other circuits are best treated as feedback or three-terminal oscillators. Three-terminal oscillators will be considered later in the course of the project.

The elementary theory of oscillators in terms of Barkhausen's condition for oscillation and similar treatments based on linear-circuit analysis are entirely satisfactory for many purposes; but when frequency stability is of first importance the nonlinear character of most circuits cannot be ignored. A complete and thorough treatment of oscillations in nonlinear circuits is beyond the scope of this project. However, a simplified treatment is not only useful but necessary to a clear understanding of the circuits to be studied.

At the time of writing the last report it was expected that tests on various circuits would begin immediately. However, it was found from preliminary experiments that performance often depends critically on what appear at first sight to be minor changes in circuit design, and it was therefore decided to postpone tests on practical circuits. For example, the exact manner of limiting the amplitude of oscillations in the circuit is very important. In fact, it is so important that it may be necessary to provide an external limiter in many cases in order to overcome the variations of transistor parameters. During the period covered by this report the nonlinear properties of point-contact transistors have been examined in some detail. Some carefully controlled experiments have been performed to check
the correlation between theory and practice. The objective of these experiments is to determine whether a simple theory can be used to estimate the importance of nonlinearity in limiting the stability of transistor-oscillator circuits.

B. Almost-Linear Series-Mode Oscillator

The circuit to be considered is shown schematically in Figure 1. Here $Z_o$ represents the resonator, $T$ represents some parameter that is assumed to vary ($T$ may be a circuit element, a battery voltage, a temperature or other quantity), $N$ is a coupling network and $Z_1$ is the impedance of the network including $T$.

![Figure 1. Basic Two-Terminal Oscillator.](image)

According to linear analysis, the condition for sustained oscillations in this circuit is simply

$$Z_o + Z_1 = 0,$$  \hspace{1cm} (1)

for if the impedance of the circuit is zero a current can flow without any source voltage.
If the system were linear and the condition (1) were met the circuit would not oscillate spontaneously; but once started by some external means, it would generate sustained oscillations of constant amplitude. If some parameter were changed so that the condition (1) could not be met at any frequency the oscillations would either die out altogether or increase in amplitude indefinitely. Such a pin-point condition would be exceedingly difficult to establish in practice and would be of little use. Practical oscillators are constructed so that a very small disturbance, even thermal noise, starts growing oscillations, but the growth causes the system to change in such a way that an equilibrium is reached for a finite amplitude of oscillations. If some parameter is changed, a new equilibrium results. Since the adjustment to equilibrium is automatic, practical oscillators are amplitude stable. Amplitude stability cannot be achieved without using some nonlinear element or device. Perhaps the simplest situation occurs when the volt-ampere characteristic of $Z_1$ is a single-valued function of the current. A typical characteristic of this kind is shown in Figure 2, which represents the emitter characteristic for a particular Western Electric type 1768 transistor when used as shown in Figure 3. The dynamic input resistance is given by the slope of this characteristic. For small variations of current near an operating point such as 0 in Figure 2 the dynamic resistance is negative and oscillations will begin if $R_1$, the magnitude of this resistance, is greater than $R_0$. Amplitude limiting depends on the fact that the effective value of $R_1$ decreases with increasing amplitude of oscillations. An exact analysis of this oscillator is quite involved; however, in practical situations simplifying approximations can be made which permit a satisfactory
Figure 2. Typical Negative-Resistance Characteristic.

Figure 3. Circuit for Obtaining Negative Resistance.
explanation of its behavior. The most important assumption is that the current in the circuit is nearly sinusoidal. For each value of current there is a corresponding value of the fundamental component of voltage. The ratio of this voltage to the fundamental component of current gives $R_1$, provided these two are in phase. Assuming for the moment that they are in phase, $R_1$ can be measured easily by means of the circuit in Figure 4.

![Circuit Diagram](image)

**Figure 4.** Circuit for Studying Input Impedance.

The operation of this circuit is explained later, in section D.

The variation of $R_1$ with current for a particular type 1768 transistor is shown in Figure 5 for several values of collector supply voltage. (Emitter bias current is constant.) These curves may be used to determine the amplitude of oscillations corresponding to a specified value of the tank-circuit resistance, $R_0$. In fact, they can be interpreted as representing the variation of oscillating current as $R_0$ is varied. In this sense they give only an approximation because the actual oscillating current is not quite sinusoidal.
However, when the distortion is small, the approximation is a very good one. Amplitude stability results from the fact that if the current is smaller than that corresponding to \( R_0 \), the net effective resistance of the circuit is negative and the oscillations therefore increase in amplitude, whereas the converse is true if the current is larger than the equilibrium value. Since the variation of \( R_1 \) is the result of distortion, the amplitude limiting is likewise regarded as a result of distortion.

The operating frequency of this circuit is determined by the condition implied by equation 1 that the net circuit reactance must be zero. If the current in the circuit is truly sinusoidal and if the volt-ampere characteristic of \( Z_\perp \) is truly single-valued, analysis shows that the fundamental component of the voltage across \( Z_\perp \) remains in phase with the current in spite of the distortion. This phase relationship still holds when the current wave is distorted, provided the current wave can be expressed as a cosine series. In
other words, if the reactance of $Z_o$ is zero not only at the fundamental frequency but also at its harmonics the operating frequency is not affected by the nonlinear nature of $Z_j$. If, as is usually the case, $Z_o$ has a reactive component at the harmonic frequencies the voltage across $Z_1$ is not in phase with the current at the fundamental frequency. Under these circumstances $Z_1$ may be regarded as having a reactive component at the fundamental frequency, and the distortion causes the frequency to be "pulled" away from the natural resonant frequency of the tank circuit.

In terms of the harmonic currents, the condition on the reactances is given by Groszkowski's equation,

$$\sum_{n=1}^{\infty} n^2 I_n x_n = 0, \quad (2)$$

in which $n$ is the order of the harmonic current $I_n$, and $x_n$ is the reactance of the passive circuit at this frequency. In terms of the harmonic voltages equation 2 becomes

$$\sum_{n=1}^{\infty} n V_n^2 \left(\frac{x_n}{|z_n|^2}\right) = 0, \quad (3)$$

in which $|z_n|$ is the magnitude of the impedance at the nth harmonic. Equation 3 may also be written in terms of susceptance as

$$\sum_{n=1}^{\infty} n V_n^2 b_n = 0, \quad (4)$$

since $x_n/|z_n|^2 = -b_n$. 

-11-
Equation 3 may be rearranged to give the following more convenient result.

\[ x_1 = - \sum_{n=2}^{\infty} n \left( \frac{V_n}{V_1} \right)^2 \left| \frac{z_n}{z_1} \right|^2 x_n \]  \hspace{1cm} (5)

This is the reactance that \( Z_0 \) must have at the operating frequency. As an example, let \( Z_0 \) be a simple series circuit \( R_0, L_0, C_0 \). Then

\[ Z_0 = R_0 + j(\omega L_0 - \frac{1}{\omega C_0}) = R_0 \left[ 1 + jQ_0 \left( \frac{\omega}{\omega_0} - \frac{\omega_0}{\omega} \right) \right], \]  \hspace{1cm} (6)

where

\[ \omega_0^2 = \frac{1}{L_0 C_0}, \]  \hspace{1cm} (7)

\[ Q_0 = \frac{\omega_0 L_0}{R_0}, \]  \hspace{1cm} (8)

and

\[ x_n = R_0 Q_0 \left( \frac{n\omega_0}{\omega_0} - \frac{\omega_0}{n\omega_0} \right) = R_0 Q_0 \left( n - \frac{1}{n} \right). \]  \hspace{1cm} (9)

Except for very small \( Q_0 \),

\[ z_1 \approx R_0 \]  \hspace{1cm} (10)

and

\[ z_n \approx x_n \text{ for } n \geq 2. \]  \hspace{1cm} (11)

Substituting these results into equation 5 gives

\[ X_0 = x_1 = \sum_{n=2}^{\infty} \left( \frac{V_n}{V_1} \right)^2 \left( \frac{\omega L_0}{Q_0^2} \right) \left( \frac{n^2}{n^2 - 1} \right). \]  \hspace{1cm} (12)
If $X_0$ is small, then to a good approximation

$$X_0 \approx 2\omega_0 L_0 \left(\frac{\omega - \omega_0}{\omega_0}\right) = 2L_0 \Delta \omega$$

and so equation 12 yields

$$\Delta \omega = -\frac{\omega_0}{2Q^2} \sum_{n=2}^{\infty} \frac{V_n^2}{V_1} \left(\frac{n^2}{n^2 - 1}\right).$$

According to equation 14, a circuit operating with 30 per cent second harmonic, 20 per cent third harmonic, and 10 per cent fourth harmonic and with a tank circuit $Q$ of 100 would be "pulled" away from the resonant frequency of the tank by about 0.001 per cent. It should be noted that these figures represent relatively little distortion of the current. The second-harmonic current, for example, would be only 0.2 per cent of the fundamental. Any change in the circuit external to the tank that tends to increase the amplitude will ordinarily increase the distortion as required by the limiting action and this in turn will cause the frequency to change. It is clear, therefore, that the frequency stability will be improved if amplitude control can be obtained without depending on distortion.

Unfortunately, practical circuits always contain parasitic reactances that may cause the impedance to harmonics to be much less than is indicated by equation 6. The resulting increase in harmonic currents will cause the actual frequency deviation to be greater than is given by equation 14. Parasitic reactances may also cause the volt-ampere characteristic of $Z_1$ to be a double-valued function of the current. In this case Groszkowski's equation does not apply, but it seems likely that the effect of distortion would be increased. Thus, distortion may play an important part in oscillator stability even at high frequencies.
The first step in finding out just how important distortion is at high frequencies is to verify the theory at low frequencies where parasitic reactances have very little influence. Even here circuit parameters cannot be measured with precision sufficient to enable the frequency to be precalculated with the required accuracy. However, if the actual operating frequency and the harmonic voltages are measured, the equations can be used to determine $\Delta \omega$ or $\Delta f$. It is not necessary to know the circuit Q exactly because, as may be seen from equation 14, $\Delta \omega$ can be computed to about the same accuracy as that of the measured values.

The results of an experiment of this kind are shown in Figure 6. The circuit used is shown in Figure 7. Note that the emitter bias supply does not affect the calculation because its current and voltage are in phase. The sharp drop in second-harmonic voltage at about 14-volts supply voltage is caused by the fact that the emitter current is reversed during part of the cycle. Comparison of the measured and computed frequency variation discloses a discrepancy that is too large to be ignored. The experiment is being repeated in an effort to determine the cause of this difficulty.

Figure 8 shows a modification of this oscillator that illustrates one method of improving the frequency stability through a reduction in distortion. A parallel circuit tuned to the second harmonic is placed in the collector circuit and a series condenser is added to restore the power factor to unity at the fundamental frequency. The curves in Figure 9 describe the performance of this circuit. Although the trap is not very effective after emitter limiting begins, at the lower collector voltages, near twelve volts for example, it produces a ten-to-one improvement over the first circuit.
Figure 6. Characteristics of the Oscillator in Figure 7.
Figure 7. Circuit for Checking Groszkowski's Equation.

Figure 8. Circuit for Reducing Second-Harmonic Distortion.
Figure 9. Characteristics of the Oscillator in Figure 8.
C. Analysis of Elementary Negative-Resistance Transistor Oscillator

An elementary negative-resistance transistor oscillator may be constructed by placing a series-mode LC tank in series with either the emitter or collector leads of the circuit shown in Figure 10. Satisfactory operation may also be obtained by placing a parallel-mode resonator in series with the base lead of the transistor. However, the last alternative is not discussed in this report. The bias supplies, which are not shown in Figure 10, may be placed to obtain either grounded-base, grounded-emitter or grounded-collector operation.

![Figure 10. Basic Configuration for Negative-Resistance Oscillator Using Point-Contact Transistor.](image)
When the operation of the oscillator is almost linear, the frequency of oscillation can be closely predicted by linear network theory. In this type of analysis, the condition for sustained oscillations is obtained by setting the determinant of the equilibrium equations of the equivalent circuit equal to zero. Performing this operation for the equivalent circuits of Figures 11 and 12 yields the following conditions for sustained oscillations.

For the tuned-emitter operation of Figure 11:

\[
0 = \begin{bmatrix} j\omega L - \frac{j}{\omega C} \end{bmatrix} \begin{bmatrix} r_c + R_c + r_b + R_b \end{bmatrix} + \begin{bmatrix} r_b + R_b \end{bmatrix} \begin{bmatrix} r_e + R_e + R_c + (1-a)r_c \end{bmatrix} 
+ \begin{bmatrix} r_e + R_e \end{bmatrix} \begin{bmatrix} r_c + R_c \end{bmatrix}. \tag{15}
\]

For the tuned-collector operation of Figure 12:

\[
0 = \begin{bmatrix} j\omega L - \frac{j}{\omega C} \end{bmatrix} \begin{bmatrix} r_e + R_e + r_b + R_b \end{bmatrix} + \begin{bmatrix} r_b + R_b \end{bmatrix} \begin{bmatrix} r_e + R_e + R_c + (1-a)r_c \end{bmatrix} 
+ \begin{bmatrix} r_e + R_e \end{bmatrix} \begin{bmatrix} r_c + R_c \end{bmatrix}. \tag{16}
\]
Setting the real and imaginary parts of these two expressions equal to zero at the natural resonant frequency of the LC tank provides the results

$$\omega = 1/\sqrt{LC}$$  \hspace{1cm} (17)

and

$$a = \left[ \frac{R_e + R_c}{r_b + R_b} \right] \left[ \frac{R_c}{r_c} \right] + \frac{R_e + R_c}{r_c}$$  \hspace{1cm} (18)

Equations 15 and 16 can be used to determine the relative frequency stability of the two equivalent circuits with respect to nonlinear variations in a single element. The relative frequency stability is defined as

$$S_f = \frac{|\omega\frac{dT}{dt}|}{T}.$$  \hspace{1cm} (19)

In this equation, $T$ is the parameter of the equivalent circuit that is allowed to vary. Of course, the variations in $T$ must be such that the constraints obtained by making the circuit determinant zero continue to be met.

For the equivalent circuits of Figure 11 and Figure 12 the relative frequency stability $S_f$, calculated at the natural frequency of the LC tank, is a measure of the insensitivity of the operating frequency to small angular variations in the parameter $T$. In other words, when the frequency stability $S_f$ is a small number a small phase shift in the parameter $T$ (due to nonlinearities) causes a relatively large variation in the operating frequency. If the frequency stability $S_f$ is a large number a small phase shift in the parameter $T$ causes a relatively small variation in the operating frequency. The linear analysis which has been used here does not provide any information about the magnitude of the phase shift that is imposed on the parameter $T$ due to nonlinear behavior.
Figure 12. Equivalent Circuit of Tuned-Collector Oscillator.

The relative frequency stability \( S_f \) of the circuits shown in Figures 11 and 12 may be obtained by the use of equations 15 and 16. These frequency stabilities, evaluated at the natural resonant frequency of the LC tank, are shown in Tables I and II. Table I, in a slightly modified form, is included in Quarterly Report No. 1 for this project.

The formulas shown in Tables I and II may be used to design and compare performance with respect to frequency stability of oscillator circuits of the type shown in Figures 11 and 12. When using these formulas, however, it should be remembered that the equivalent circuits that have been used do not provide an exact description of the performance of the physical oscillator. Furthermore, the frequency stability has been evaluated only for a small phase shift in a single element. In practice, variations of an external parameter such as bias voltage may cause variations in several or all of the parameters of the equivalent circuit.
TABLE I

RELATIVE FREQUENCY STABILITY OF THE TUNED-EMITTER OSCILLATOR
OF FIGURE 11 EVALUATED AT THE FREQUENCY \( \omega = 1/\sqrt{LC} \).

<table>
<thead>
<tr>
<th>Variable</th>
<th>Relative Frequency Stability</th>
</tr>
</thead>
</table>
| \( r_b \) | \[
\frac{2}{r_b} \sqrt{\frac{L}{C}} \left( \frac{r_b + R_b}{r_e + R_e} \right) \left( 1 + \frac{r_b + R_b}{r_c + R_c} \right) \]
| \( r_c \) | \[
\frac{2}{r_c} \sqrt{\frac{L}{C}} \left[ \frac{1 + r_c + R_c}{r_b + R_b} \right] \left[ 1 + \frac{r_e + R_e}{r_b + R_b} \right] \]
| \( r_e \) | \[
\frac{2}{r_e} \sqrt{\frac{L}{C}} \left[ \frac{1 + r_c + R_c}{r_b + R_b} \right] \]
| \( a \) | \[
\frac{2}{ar_c} \sqrt{\frac{L}{C}} \left[ \frac{1 + r_c + R_c}{r_b + R_b} \right] \]

In spite of these facts, this type of analysis may provide information which is useful in the design of transistor oscillators that are nearly sinusoidal in operation. In other words, an oscillator that is designed to be stable with respect to independent variations in all of the parameters of its equivalent circuit is expected to be stable with respect to collective variations in all parameters, assuming again that the operation is almost linear.

Frequency stability tests have been made for a negative-resistance oscillator using several transistors to test the applicability of the type of analysis presented here. The performance determined experimentally was found to agree qualitatively with the performance predicted by analysis. The results of one such test are included here to show the general nature of the agreement.
between the predicted performance of the oscillator and the experimentally determined performance. A low-frequency, low-Q circuit was used for ease in measurement. The analysis should be applicable over as wide a frequency range as the equivalent circuit is valid. At higher frequencies a more complicated equivalent circuit undoubtedly will have to be used.

Equal values of external resistance were used in the emitter, base and collector leads. Therefore, the only variation in the two circuits for this
test was the placing of the LC tank. The oscillations were maintained close to the threshold condition to insure as nearly sinusoidal operation as possible.

A type 1768 point-contact transistor having the following measured parameter values was used in these tests.

\[
\begin{align*}
    a &= 2.25 \\
    r_c &= 3,500 \text{ ohms} \\
    r_b &= 110 \text{ ohms} \\
    r_e &= 40 \text{ ohms}
\end{align*}
\]

Each of the external resistances \( R_b, R_c \) and \( R_e \) was 450 ohms. The tank coil \( L \) had an inductance of 0.5 henries, and the capacitor \( C \) had a capacitance of 0.0534 microfarads.

The relative frequency stabilities for the above parameter values, as calculated by use of Tables I and II, are shown in Table III.

**TABLE III**
CALCULATED RELATIVE FREQUENCY STABILITY OF ELEMENTARY NEGATIVE-RESISTANCE TRANSISTOR OSCILLATORS

<table>
<thead>
<tr>
<th>Variable</th>
<th>( r_b )</th>
<th>( r_c )</th>
<th>( r_e )</th>
<th>( a )</th>
</tr>
</thead>
<tbody>
<tr>
<td>Tuned Emitter ( S_f )</td>
<td>73</td>
<td>37</td>
<td>150</td>
<td>6.2</td>
</tr>
<tr>
<td>Tuned Collector ( S_f )</td>
<td>17</td>
<td>8.6</td>
<td>36</td>
<td>1.5</td>
</tr>
</tbody>
</table>

An inspection of Table III shows that the frequency of the tuned-emitter oscillator should be roughly four times as insensitive to phase shifts in the parameters of the equivalent circuit as the frequency of the tuned-collector oscillator.
oscillator. It should be remembered that this prediction is made only for an oscillator circuit using the particular parameter values shown above. Entirely different predictions might be expected to result from a different circuit design. An inspection of Table III also shows that the frequency of oscillation of both circuits is more sensitive to phase shifts in $a$ and $r_c$ than to phase shifts in $r_b$ and $r_e$.

Measurements of the variation of operating frequency due to variations in bias were made for the two oscillator circuits described above. Circuit diagrams for the circuits used in the measurements are shown in Figures 13 and 14. In all cases variations in bias were made from an operating point of $V_c = 14$ volts and $I_e = 2$ milliamperes. The results of these frequency-stability measurements are shown in Figures 15 and 16.

![Figure 13. Tuned-Emitter Oscillator Circuit.](image-url)
Figure 14. Tuned-Collector Oscillator Circuit.

Figure 15. Performance of the Circuit in Figure 13.
An inspection of the curves of Figures 15 and 16 shows that the frequency of the tuned-emitter oscillator, as predicted, is several times as stable as that of the tuned-collector oscillator. Similar results were obtained for several transistors used in the same type circuit, provided that the operation was maintained close to the threshold of oscillation.
D. Experimental Data

Several experiments have been run in order to study the input (emitter to ground) impedance of the grounded-base point-contact transistor. This impedance can be real and negative and can, therefore, be used to build a negative-resistance type oscillator.

It was desired to determine the magnitude of negative resistance obtainable, its nonlinear properties and the manner in which it changes with various parameters, such as bias and collector or base-circuit impedance. The nonlinear properties were determined by measuring the harmonics in the emitter-voltage wave when the emitter was driven by a high-impedance sinusoidal-current source.

The circuit used for these measurements was shown earlier in Figure 4. A harmonic-wave analyzer WA, and a precision 1,000-ohm potentiometer were used in the emitter circuit. The fundamental frequency was 1 kc.

To measure the magnitude of negative input resistance the wave analyzer is tuned to the fundamental frequency and the 1,000-ohm potentiometer is adjusted until a null is indicated. Then the total resistance across which the voltage is being measured is zero and the resistance of the potentiometer is, therefore, the negative of the input resistance of the transistor. The fact that the null obtained in this way was very good indicates that the input impedance is practically pure resistance at the fundamental frequency. It should be noted, however, that this is true only when the driving current is a pure sine wave.
By connecting the wave analyzer directly to the emitter terminal the harmonics of the emitter voltage can be measured.

A few selected curves showing the variation of negative input resistance and significant harmonic voltages with other circuit parameters are shown in Figures 17 through 21. All of these curves were taken using a single type 1768 point-contact transistor.

Figures 17 and 18 show the manner in which the input resistance $R_1$ and the second-harmonic voltage $H_2$ vary with the amplitude of the signal current in the emitter circuit. The curves in Figure 17 were obtained with $R_c$ much smaller than the transistor-collector resistance $r_c$, so that the collector current was almost independent of $R_c$. Under these conditions the input resistance is negative and only slightly smaller in magnitude than $R_b$. This magnitude is practically constant and little distortion is present for signals up to 1 ma. Reference to the static collector family curves for the type 1768 transistor indicates that the operation should be essentially linear in this case.

Conditions for the curves of Figure 18 are the same as for Figure 17 except that $R_c$ was increased to 2,700 ohms, which is only slightly smaller than $r_c$. The magnitude of $R_1$ is reduced to approximately $R_b/2$. However, $R_1$ decreases appreciably and the second harmonic component of the emitter voltage increases for larger signal amplitudes. The static curves show that in this case the collector is being driven into nonlinear region when the emitter current exceeds approximately 0.7 ma rms. This is in agreement with the data plotted in Figure 18.

In order for a negative-resistance oscillator to have good amplitude stability $R_1$ must decrease with increasing signal amplitude, and it is desirable
Figure 17. The Effect of Signal Current on $R_1$ and $H_2$, $R_c = 1\, k$. 

Figure 18. The Effect of Signal Current on $R_1$ and $H_2$, $R_c = 2.7\, k$. 

-30-
that the harmonic voltages remain small. The circuits corresponding to Figures 17 and 18 are not particularly good in this respect.

One way in which this characteristic can be improved is seen by comparing the curves in Figures 19A and 19B. The data in Figure 19A corresponds to an $R_c$ of 390 ohms. The circuit used for Figure 19B was identical except that a Mazda Lamp having a cold resistance of 400 ohms was used for $R_c$. The increase in the lamp resistance due to heating causes the input resistance to decrease much more rapidly with signal current than it does with a linear $R_c$. Notice that for the same magnitude of $R_1$, the harmonic content in Figure 19B is smaller than that in Figure 19A. For example, when $R_1 = 700$ ohms the per cent $H_2$ in Figure 19B is only half that in Figure 19A.

By careful design it should be possible to build oscillators using this type of nonlinear resistance so that the amplitude stability is good and the harmonic content is very small. The important property of this class of circuits is that amplitude limiting is accomplished by means of a slowly varying nonlinear resistance rather than by overloading the amplifier. The Meacham Bridge Oscillator is probably the best-known example of this general class of oscillators.

Figures 20 and 21 show typical variations of $-R_1$ and per cent $H_2$ when the signal amplitude is fixed but a d-c bias is changed. Both curves show regions of reasonably fast change of $R_1$ and small harmonic content. This suggests the possibility of using some kind of automatic gain control to limit the amplitude of oscillation.

For example, Figure 20 indicates that for $V_c = -5$ volts, $I_e = 2.5$ ma and $i_e = 0.3$ ma, $R_1 = -600$ ohms, and the harmonic voltage is less than five per cent of the fundamental voltage. If an AGC circuit can be used to increase
Figure 19. An External Limiter Reduces Distortion.
Figure 20. Variation of $R_1$ and $H_2$ with Emitter Bias.

Figure 21. Variation of $R_1$ and $H_2$ with Collector Bias.
I_e when i_e exceeds 0.3 ma the amplitude of oscillation may be stabilized at a value that produces small distortion. A similar result could be obtained by causing the collector-bias voltage to decrease when the amplitude of oscillation increases.

Oscillators employing AGC are similar to the lamp-controlled oscillators described above in that the amplifier can be made essentially linear and limiting can be accomplished by other means.

E. Measurement Techniques

A test set for measuring the small-signal h-parameters at low frequencies was described in Quarterly Report No. 1. This set has been used for measuring the small-signal parameters when they are real. However, since it measures only the magnitude of the parameters it does not give the complete data when a parameter is complex. Also, its useful frequency range is limited.

Preliminary plans have been made to supplement the existing test set in three ways:

1. The construction of a small-signal parameter test set, basically similar to the present low-frequency set, which can be used at considerably higher frequencies.

2. The use of an impedance bridge to measure complex input and output impedances at high frequencies.

3. The use of the Law of Cosines and other relations from which the complex parameters may be calculated from measurements that can be conveniently made with the equipment listed above.

If the magnitudes of the two complex quantities α|φ and (1 - a) |φ can be measured, the phase angles can be found from the relation

\[ α|φ + (1 - a) |φ = 1 \]  

(20)
This equation can be represented as a vector diagram which can be constructed without knowing the phase angles $\theta$ and $\phi$. An example of this construction is shown in Figure 22. The angle $\theta$ can be measured on the vector diagram or it can be calculated using the Law of Cosines.

\[
\cos \theta = \frac{1 + |a|^2 - |1 - a|^2}{2|a|}.
\]

Figure 22. Vector Diagram Representing Equation 20.

Thus,

\[
\cos \theta = \frac{1 + |a|^2 - |1 - a|^2}{2|a|}.
\]  

The complex value of $\alpha$ can be checked by measuring the input (emitter to base) impedance of the transistor with the collector shorted to the base. This impedance is

\[
Z_{zz} = r_e + (1 - a)r_b.
\] 

The resistances $r_e$ and $r_b$ can be measured at low frequencies. They are constant pure resistances throughout the range of frequencies for which simplified equivalent limits are valid.
If $r_c$ becomes complex at high frequencies this impedance can be measured with an impedance bridge. Various other open- and short-circuit impedance measurements can be used to measure and check the parameters in an equivalent circuit at high frequencies where any or all of these parameters are complex.

In a point-contact transistor $\alpha$ (or $a$) is the only parameter of the equivalent-$T$ circuit that varies with frequency until the frequency far exceeds the $\alpha$ cut-off frequency. In such cases the resistance parameter may be measured at low frequencies and $\alpha$ determined at the frequencies of interest.
V. CONCLUSIONS

Waveform distortion due to nonlinearity can cause the frequency of an oscillator to differ appreciably from that of its tuned circuits. Variation of distortion caused by changes in circuit parameters is one source of frequency variation. In the elementary oscillators considered here, distortion is necessary for amplitude stability. Frequency stability can be improved by means of auxiliary amplitude-control devices and the use of special circuits to suppress harmonics.

A qualitative measure of the sensitivity of a given circuit to parameter changes can be obtained by linear circuit analysis. This method appears to give a satisfactory comparison of various circuits and, consequently, it may prove useful for design purposes.
VI. PROGRAM FOR THE NEXT INTERVAL

1. Analytical studies will be continued.

2. Practical studies of crystal-controlled transistor oscillators will be made.
VII. PERSONNEL

The key personnel working on this project during the period covered by this report are:

<table>
<thead>
<tr>
<th>Name</th>
<th>Position</th>
<th>Approximate Hours Worked</th>
</tr>
</thead>
<tbody>
<tr>
<td>B. J. Dasher, Sc.D.</td>
<td>Project Director</td>
<td>260</td>
</tr>
<tr>
<td>D. L. Finn, Ph.D.</td>
<td>Research Associate</td>
<td>90</td>
</tr>
<tr>
<td>W. B. Jones, Jr., Ph.D.</td>
<td>Research Associate</td>
<td>200</td>
</tr>
<tr>
<td>Wai Mun Syn, M.S.</td>
<td>Research Assistant</td>
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Respectfully submitted:

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D. L. Finn  
Research Associate

J. E. Boyd, Head 
Physics Division

Herschel H. Cudd, Director  
Engineering Experiment Station
VIII. BIBLIOGRAPHY


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Atlanta, Georgia

QUARTERLY REPORT NO. 3
PROJECT NO. 236-206

TRANSISTOR OSCILLATORS OF
EXTENDED FREQUENCY RANGE

By
B. J. DASHER, W. B. JONES, JR.
and
S. N. WITT, JR.

CONTRACT NO. DA-36-039-sc-42712
DEPARTMENT OF THE ARMY PROJECT: 3-99-11-022
SIGNAL CORPS PROJECT: 142B

MARCH 31, 1954
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TABLE OF CONTENTS

<table>
<thead>
<tr>
<th>Section</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>I. PURPOSE</td>
<td>1</td>
</tr>
<tr>
<td>II. ABSTRACT</td>
<td>2</td>
</tr>
<tr>
<td>III. CONFERENCES</td>
<td>3</td>
</tr>
<tr>
<td>IV. FACTUAL DATA</td>
<td>4</td>
</tr>
<tr>
<td>A. Introduction</td>
<td>4</td>
</tr>
<tr>
<td>B. Effect of Distortion on Frequency</td>
<td>4</td>
</tr>
<tr>
<td>C. Parasitic Oscillations</td>
<td>12</td>
</tr>
<tr>
<td>D. Evaluation of Practical Circuits</td>
<td>13</td>
</tr>
<tr>
<td>V. CONCLUSIONS</td>
<td>24</td>
</tr>
<tr>
<td>VI. PROGRAM FOR THE NEXT INTERVAL</td>
<td>25</td>
</tr>
<tr>
<td>VII. PERSONNEL</td>
<td>26</td>
</tr>
<tr>
<td>VIII. BIBLIOGRAPHY</td>
<td>27</td>
</tr>
</tbody>
</table>

This Report Contains 27 Pages
LIST OF FIGURES

1. Harmonic Distortion and Frequency as Functions of Collector Supply Voltage .............................................. 7
2. Circuit Diagram of the Test Oscillator ................................................. 8
3. Comparison of the Computed Natural Frequency and the Actual Frequency .............................................. 10
4. Frequency Comparison Curves Including the Effects of the First Eight Harmonics .............................................. 10
5. Results Obtained with Battery Operation ................................................. 11
6. Free-Running Oscillator Using a Single Bias Source ................................................. 14
7. Tuned-Emitter Oscillator ................................................. 15
8. Crystal-Controlled Oscillator for High-Impedance Crystal ................................................. 16
9. Crystal-Controlled Oscillator Using Transformer Feedback ................................................. 17
10. Crystal-Controlled Oscillator with Transformer Feedback and Drift Compensation ................................................. 18
11. Crystal-Controlled Oscillator with Tuned-Base Circuit ................................................. 19
12. Circuit of Figure 11 Modified for Single Battery Operation ................................................. 19
13. 1-Mc Crystal-Controlled Oscillator ................................................. 20
14. 11-Mc Free-Running Oscillator ................................................. 22

LIST OF TABLES

1. SUMMARY OF CIRCUIT PERFORMANCE DATA ................................................. 23
I. PURPOSE

The purpose of this project is to evolve data and information regarding transistor-oscillator circuitry and the degree of reliability and reproducibility attained with transistor-parameter variations. It includes both analytical and experimental studies of crystal-controlled and free-running oscillators employing transistors exclusively as active elements.

The work is divided into three principal tasks as follows:

(1) The evaluation of crystal-controlled and free-running transistor oscillators from the point of view of reliability, size, weight, circuit requirements, frequency limitations and other similar considerations;

(2) The development of new circuits and/or techniques specifically to take advantage of the unique characteristics of transistors regarding size, weight, power requirements, etc; and

(3) The development of practical engineering data and information to serve as a basis for designing free-running and crystal-controlled junction- and point-contact transistor oscillators in the range 0.1 to 15 mc.
II. ABSTRACT

The results of a series of experiments performed for the purpose of finding out whether Groszkowski's equation gives sufficiently accurate results to be useful in computing the frequency pulling due to distortion in transistor oscillators are described. These tests were made on series-mode oscillators at 1,000 cps. When the pulling is of the order of 0.5 per cent, the theory accounts for about four-fifths of it. The cause of the remaining discrepancy has not been determined.

Evaluation tests on several transistor-oscillator circuits have been made. These circuits include both free-running and crystal-controlled oscillators at frequencies from 100 kc to 10 mc. Frequency measurements were made by means of a Berkeley model 5570S frequency meter. The results of the tests are summarized in a table which shows the relative frequency stability together with an estimate of the relative temperature stability. Based on a one per cent change of power-supply voltage, frequency stabilities as good as 0.3 part per million were observed.
III. CONFERENCES

Mr. E. Gonzales of SCEL visited Georgia Tech December 15 and 16, 1953. The objectives of the project and the current status of the work were discussed with him.

Dr. B. J. Dasher and Dr. W. B. Jones, Jr., attended the AIEE-IRE Conference on Transistor Circuits at the University of Pennsylvania in Philadelphia, February 18 and 19, 1954.
IV. FACTUAL DATA

A. Introduction

The elementary theory of two-terminal, negative-resistance oscillators was discussed in Quarterly Report No. 2. The mechanism of amplitude limiting in this type of circuit was considered in detail, and it was pointed out that limiting which depends entirely on amplitude distortion is not very satisfactory. This is the type of limiting which has been used in most of the transistor circuits studied so far. The main objections to this limiting are these:

1. The distortion causes the negative resistance to have a reactive effect associated with it, and since the magnitude of this effect varies with distortion, the oscillator frequency likewise varies with distortion.

2. Proper operation depends on relatively critical adjustments of circuit parameters. For this reason changing transistors ordinarily requires resetting these adjustments.

The relation between distortion and frequency is considered further in the following pages. Tests have been made on several circuits for the purpose of evaluating their relative frequency stability and other operating characteristics. The results of these tests are also presented.

Oscillator analysis in terms of linear-circuit theory has also been considered in previous reports. It has been concluded, at least tentatively, that this method of analysis is not likely to be helpful in these studies, and so this work has been discontinued.

B. Effect of Distortion on Frequency

Groszkowski studied the relation between harmonic distortion and frequency in a two-terminal, negative-resistance oscillator. He derived an expression
which can be used to calculate the frequency of oscillations, provided the volt-
ampere characteristic of the nonlinear resistance is a single-valued function
of either current or voltage. This condition is satisfied if the characteristic
does not show any hysteresis. Physically it means that the negative resistance
cannot store any electrical energy. Groszkowski's equation may be written as
follows:

\[
x_1 = - \sum_{n=2}^{\infty} \left( \frac{v_n}{v_1} \right)^2 \left| \frac{z_n}{z_1} \right|^2 x_n.
\]

In this expression \(v_n\) is the magnitude of the tuned-circuit voltage at the
nth harmonic, \(z_n\) is the impedance of the tuned circuit at the nth harmonic
and \(x_n\) is its reactance. It was shown in the last report that when the pas-
sive part of the circuit is a simple series-resonant impedance equation 1
yields:

\[
\Delta \omega = - \frac{\omega}{2Q_0} \sum_{n=2}^{\infty} \left( \frac{v_n}{v_1} \right)^2 \frac{n^2}{n^2 - 1} \tag{2}
\]

in which \(Q_0\) is the Q of the passive circuit at its resonant frequency \(\omega_0\) and
\(\Delta \omega = \omega - \omega_0\). The operating frequency is \(\omega\).

A series of experiments have been performed for the purpose of finding
out whether Groszkowski's equation gives sufficiently accurate results to be
useful in connection with transistor oscillators. The question is not whether
the equation is correct, but rather whether it applies to transistor circuits
which are not as simple as the theory assumes. These experiments were per-
formed at low frequencies (approximately 1,000 cps) in order to reduce the
effects of parasitic capacitances, transistor phase shifts, etc. Also, low-Q circuits were used to facilitate frequency measurements. Frequency was measured by means of a Berkeley model 554 Eput meter. Since this instrument has a relatively low input impedance, a cathode-follower circuit employing a type 6C4 tube was used as an isolation amplifier. Voltages were measured with a General Radio type 736A wave analyzer.

It is not feasible to calculate \( \omega_0 \) directly because the element values cannot be determined with sufficient accuracy. It probably would be possible to measure the resonant frequency of the tuned circuit. However, instead of comparing measured and computed operating frequencies, it was decided to use equation 2 to calculate \( f_o \), the natural resonant frequency of the tank circuit, in terms of the actual frequency and the distortion voltages. Solving equation 2 for \( f_o \) gives

\[
f_o \approx f \left[ 1 + \frac{1}{2Q_0^2} \sum_{n=2}^{\infty} \left( \frac{V_n}{V_1} \right)^2 \frac{n^2}{n^2 - 1} \right]. \tag{3}
\]

Equations 2 and 3 involve the assumption that \( \Delta \omega/\omega_0 \ll 1 \), an assumption which is entirely justified.

Figure 1 shows the results obtained by varying the collector-supply voltage in the circuit of Figure 2.

When studying the curves in Figure 1, it is important to remember that the primary objective of these experiments is to determine whether the Groszkowski equation is applicable to the circuit. The significance of the curves relative to oscillator performance is of secondary interest.
Figure 1. Harmonic Distortion and Frequency as Functions of Collector Supply Voltage.
At low voltages the emitter current is positive throughout the cycle, and limiting is due mainly to collector cut-off. As the collector-supply voltage increases, the amplitude of oscillations increases until the voltage reaches 27.5 volts. At this point, the emitter current is negative during a small part of the cycle, and limiting is due almost entirely to emitter cut-off. Insofar as amplitude stability is concerned, emitter limiting appears to be more effective than collector limiting. This is to be expected because the circuit has a large positive resistance when the emitter diode is reversed—much larger than when the collector diode is reversed. This large positive resistance readily offsets the negative resistance, which is effective during most of the cycle. For voltages below 27.5, the distortion is largely second harmonic. Above this voltage, the higher order harmonics become increasingly important. It may also be observed that both above and below the voltage at
which emitter limiting begins the frequency stability is relatively good, but the frequency decreases rapidly during the transition from collector limiting to emitter limiting.

On either side of this transition, the natural frequency of the tank circuit, as computed from equation 3, is nearly constant, but it tends to follow the actual frequency during the transition. Of course, $f_0$ should be quite independent of the collector-supply voltage. Since, at the higher voltages, the fifth harmonic is larger than the second and almost as large as the third, it appears that better results would be obtained if harmonics of higher order were included in the calculations.

Figure 3 shows another comparison of actual frequency and computed natural frequency. The circuit used for this data was the same as in Figure 1 except that a different transistor was used. Harmonics through the fifth were included in the calculations. The variation of the harmonic voltages, although not shown, was similar to that in Figure 1. The curves in Figure 4 were obtained from the same transistor as those in Figure 1. Harmonics through the eighth were included in these calculations. In all three of these tests, the emitter-bias current was held constant at 2 mA. The differences in the results are believed to be largely caused by differences in ambient temperature. This conclusion is supported by the fact that the collector current varied from day to day. Two successive runs made with the same setup on the same day gave substantially the same results.

One striking discrepancy among these results is that the computed natural frequency of the tank circuit varies from run to run, even at low voltages where the distortion is small. It was suspected that this variation might be
Figure 3. Comparison of the Computed Natural Frequency and the Actual Frequency.

Figure 4. Frequency Comparison Curves Including the Effects of the First Eight Harmonics.
Figure 5. Results Obtained with Battery Operation.

carced by reactive impedances of the electronic-regulated power supplies that were used as bias-current sources. When these power supplies were replaced by batteries, the results shown in Figure 5 were obtained, harmonics through the eighth being included as before. Here the computed natural frequency at low voltages is about 1,002 cps. This value represents the best agreement with the measured natural frequency which was 1,000.5 cps. Otherwise, the curves in Figure 5 are not much different from those in Figure 1. The "wiggles" in
the curves representing computed natural frequency are probably caused by experimental errors, but the persistent difference of about 2 cps between the highest and lowest voltages is evidently caused by something that has been neglected. This may be caused by phase shifts within the transistor, by ineffective by-pass capacitors, or something else as yet unknown. In spite of these discrepancies, the theory predicts the general character of the performance reasonably well and further study along these lines appears to be justified. In particular, it should be possible to compare various kinds of amplitude limiters and to make useful estimates of their relative influence on frequency stability. For example, a limiting process that depends on a large number of harmonic components of small amplitude would be superior to one that depends on a small number of components having large amplitudes because the effect on frequency varies as the square of the harmonic-voltage ratios. Again, if the above-mentioned 2-cps-frequency change is assumed to be caused by changes in some circuit parameter, the Groszkowski equation shows that most of the actual frequency variation is caused by the distortion.

C. Parasitic Oscillations

During the experiments described in the preceding section a good deal of trouble was experienced because of parasitic oscillations. Although the circuit arrangement is basically a series-mode oscillator, it is possible for parallel-mode oscillations to exist for certain combinations of parameters. The presence of as much as 20 μf of capacitance in parallel with the tuned circuit would cause spurious high-frequency oscillations to appear—presumably parallel-mode oscillations. These oscillations were present during only part of a cycle of the principal mode, namely, while the emitter current was very close
to zero. This difficulty was almost entirely eliminated through the use of the
cathode follower mentioned earlier.

D. Evaluation of Practical Circuits

During the period covered by this report evaluation tests on several tran-
sistor-oscillator circuits have been made. These circuits included both LC- and
crystal-controlled oscillators. Frequency measurements were made by means of
a Berkeley model 5570S direct-reading frequency meter. This instrument has a
short-time accuracy of at least one part in $10^6 \pm 1$ count. It is provided with
a "scanning" switch which makes possible accurate counting for any time inter-
val that is a multiple of two seconds.

The simplest of the circuits studied is shown in Figure 6. By the proper
selection of $R_1$ and $E_c$, negative-resistance values as high as 50,000 ohms could
be obtained at point a-a. This resistance was measured at zero frequency. When
a series L-C circuit was connected at points a-a as shown in the figure, os-
cillations were obtained. When series resistance up to 40,000 ohms was added
to the L-C circuit, oscillations were still obtained at frequencies up to
20,000 cps.

It was found that the L/C ratio was very important for this circuit.
There appeared to be a critical value of L/C for any particular frequency and
series-resistance value. For L/C ratios larger than this critical value, fair
stability resulted and the frequency was not greatly affected by small changes
in the resistance. For L/C ratios below this critical value (at the same res-
onant frequency), the frequency of oscillation was controlled almost entirely
by C. Short-circuiting the inductance changed the frequency only slightly in
this case. The frequency of oscillation was also very dependent on the
amount of series resistance.

-13-
The maximum frequency obtained with the circuit of Figure 6 was 120,000 cps. This was possible only when the series resistance was reduced to 200 ohms, the resistance of the inductance used at L. This fact indicates that the dynamic negative resistance is a function of frequency, being as high as 50,000 ohms at low frequencies and as low as a few hundred ohms at 100,000 cps.

The same circuit was tried using a W. E. type 1698 transistor, and in this case oscillations were obtained at frequencies up to 320,000 cps.

The tuned-emitter circuit shown in Figure 7 was also tried. The values of negative resistance obtained were much lower than for the previous circuit. However, frequency stability was considerably better. Approximate frequency variations were as follows:

Figure 6. Free-Running Oscillator Using a Single Bias Source.
Figure 7. Tuned-Emitter Oscillator.

8 cycles per second per volt change in $E_e$ at $E_e = 50$ v, $f = 50$ kc and $E_c = 16$ v.

2.2 cycles per second per volt change in $E_c$ at $E_e = 50$ v, $f = 50$ kc and $E_c = 16$ v.

In each case, $C = 1,000$ μf and $L = 10$ mh.

When the tuned circuit was changed to $L = 60$ mh and $C = 220$ μf, the frequency variation with change in emitter voltage was reduced to approximately 0.5 cycle per volt change in $E_e$ at $E_e = 50$ v, $E_c = 16$ v and $f = 50$ kc. The distortion of the waveform for the larger $L/C$ ratio was also much less.

Figure 8 shows a crystal-controlled, negative-resistance oscillator.

In order to obtain oscillations, it was necessary to adjust $R_c$ very carefully. This critical adjustment was required in order to obtain a negative, dynamic collector resistance greater than the positive, equivalent-series resistance of the crystal at resonance. The equivalent-series resistance of the
available 50-kc crystal was of the order of 40,000 ohms. At this frequency, a negative resistance greater than the required 40,000 ohms could be obtained only by placing a parallel-tuned circuit in series with the base as shown. The frequency variation for this circuit was 0.4 cycle per volt at $E_c = 25$ v and $f = 50$ kc. This large variation is attributed partly to the relatively poor crystal available.

Attempts were made to obtain oscillations at higher frequencies (1 mc) where crystals were available with series resistances as low as 150 ohms. However, sufficiently high negative dynamic resistances could not be obtained at these frequencies because of the low cut-off frequency of the transistors.

Each of the circuits in Figures 9, 10, 11, 12 and 13 exhibits a dynamic negative resistance at the operating frequency which is not present at any of the transistor electrodes at frequencies approaching direct current. For this reason, by-passing with large capacitances is usually possible at points within
the circuit where it is desired to obtain a low impedance to ground. Use will be made of this fact in some of the circuits to make possible proper operation with only one source of supply voltage.

Figure 9 shows a circuit making use of a W. E. type 1768 transistor. The frequency of oscillation varies with supply voltage as follows:

0.03 cycle per volt change in $E_C$ for $I_E = 0.5\, \text{ma}$ (held constant), $E_C = 50\, \text{v}$ and $f = 50\, \text{kc}$.

7.0 cycles per ma change in $I_E$ for $I_C = 2\, \text{ma}$, $I_E = 0.5\, \text{ma}$ and $f = 50\, \text{kc}$.

Figure 9. Crystal-Controlled Oscillator Using Transformer Feedback.

The circuit was modified as shown in Figure 10 in an attempt to obtain better frequency stability. The new frequency variations were as follows:

0.02 cycle per volt change in $E_C$ for $E_C = 80\, \text{v}$ and $f = 50\, \text{kc}$.

This value is the average obtained for several samples of the W. E. type 1768 transistor. Some particular transistors showed variations of 0.03 cycle per volt while others showed only 0.01 cycle per volt. In this circuit, oscillations also occurred when the crystal was replaced by a 13 $\mu\text{f}$ or larger capacitor.
Figure 10. Crystal-Controlled Oscillator with Transformer Feedback and Drift Compensation.

The circuit shown in Figure 11 makes possible crystal-controlled oscillations at 1 mc and higher frequencies. Frequency variations were determined for a range of supply-voltage variations, and then the circuit of Figure 12 was designed in an attempt to improve the stability. The resulting frequency variations were as follows:

1.0 cycles per volt change in $E_c$ at $E_c = 30$ v and $f = 1$ mc.

The variation was practically linear over a range of $E_c$ from 20 v to 50 v, the frequency varying 29 cycles over this range.

In analyzing this circuit it was suspected that, in addition to the positive feedback path resulting from the collector coil, other feedback paths internal to the transistor were also present. This was proved to be true when it was found that the circuit would oscillate with the crystal returned to ground instead of to the tap on the collector coil. This modification is shown in Figure 13. The new circuit resulted in an improved stability as follows:
Figure 11. Crystal-Controlled Oscillator with Tuned-Base Circuit.

Figure 12. Circuit of Figure 11 Modified for Single Battery Operation.
Figure 13. 1-Mc Crystal-Controlled Oscillator.

0.5 cycle per volt change in $E_c$ over a range of $E_c$ from 20 v to 50 v for $f = 1$ mc.

0.2 cycle per volt change in $E_c$ over a range of $E_c$ from 35 v to 40 v for $f = 1$ mc.

It was also observed that changing the value of $L_1$ from 5.5 mh to 16 mh had very little effect on the frequency of oscillation.

The frequency variation was approximately the same for seven out of ten transistor samples tried. The other three failed to oscillate because of very low activity. The exact frequency of oscillation was also directly dependent on the activity of the transistor sample. From those well below the manufacturer's minimum specifications to those well above the maximum specifications the frequency variation was about 50 cps. This frequency variation was correlated with crystal-voltage variation for different transistor activity ($E_c = 30$ v in all cases). Frequency variation for the same crystal was then determined as a function of crystal drive by means of a crystal-impedance meter.
Indications were that about half of the frequency variation with changing 
transistor activity was caused by the crystal's characteristics and not by the 
transistor.

This circuit did not oscillate when the crystal was removed. However, 
changing the tuned circuit in the base from that shown to \( L = 0.25 \, \text{mh} \) and \( C = 
100 \, \mu\text{f} \) produced oscillations with or without the crystal. With the crystal 
in place, the stability was approximately the same as before. Without the 
crystal, the frequency was approximately 20 per cent higher and stability was 
very poor.

With the circuit shown in Figure 14, oscillations were obtained at fre-
quencies as high as 11 mc. The stability with changing supply voltage was re-
latively poor. Also only one transistor out of 15 tried would oscillate at 
this high frequency. This one transistor was damaged before accurate measure-
ments of frequency stability could be made. Another transistor was found which 
would oscillate at 9 mc but it was likewise damaged. Only transistors with 
activities far above normal would operate at the higher frequencies. It was 
unfortunate that these transistors could be damaged at dissipations of only a 
fraction of their ratings. The same voltages and currents did not harm samples 
of lower activity.

The maximum frequency at which average transistors will oscillate in the 
circuit of Figure 13 (using W. E. type 1698 transistors) is 5 mc. Frequency-
stability measurements were not available in time for this report.

Since accurate temperature-control methods are not yet available, no exact 
data on temperature effects can be given. It was observed, however, that tem-
perature variations have fairly great influence on the negative static resist-
ance type of circuit. In contrast, the circuit of Figure 12 was little affected
by temperature. The circuit of Figure 13 showed less than 2 cycles per second change in frequency for 10°C change in ambient temperature. This observed variation was for operating conditions on two different days and, therefore, could be in error in either direction because of mechanical disturbances during the elapsed time.

Table I summarizes the performance of the various circuits discussed in this section. The quantity $D_f$ is a measure of the frequency stability. It is defined as the frequency change in parts per million corresponding to a one per cent change in supply voltage. Thus,

$$D_f = \frac{10^6 \Delta f / f}{\Delta E / E},$$

with $f$ in cps and $E$ in volts.
<table>
<thead>
<tr>
<th>Circuit</th>
<th>D_f</th>
<th>Variable</th>
<th>Relative Temperature Stability</th>
</tr>
</thead>
<tbody>
<tr>
<td>Figure 7</td>
<td>5</td>
<td>E_e</td>
<td>poor</td>
</tr>
<tr>
<td>Figure 8</td>
<td>2</td>
<td>E_c</td>
<td>fair</td>
</tr>
<tr>
<td>Figure 9</td>
<td>1.2*</td>
<td>E_c</td>
<td>fair</td>
</tr>
<tr>
<td>Figure 9</td>
<td>0.7**</td>
<td>E_e</td>
<td>fair</td>
</tr>
<tr>
<td>Figure 10</td>
<td>0.25-0.80</td>
<td>E_c</td>
<td>fair to good</td>
</tr>
<tr>
<td>Figure 11</td>
<td>0.135</td>
<td>E_e</td>
<td>fair</td>
</tr>
<tr>
<td>Figure 11</td>
<td>0.5</td>
<td>E_c</td>
<td>fair</td>
</tr>
<tr>
<td>Figure 12</td>
<td>0.3</td>
<td>E_c</td>
<td>good</td>
</tr>
<tr>
<td>Figure 13</td>
<td>0.15</td>
<td>E_c</td>
<td>good</td>
</tr>
</tbody>
</table>

*I_e was held constant at 0.5 milliampere.

**I_c was held constant at 2.0 milliamperes.
V. CONCLUSIONS

Oscillator analysis in terms of linear-circuit theory, considered in previous reports, does not appear likely to contribute very much new information for these studies and this work has, therefore, been discontinued.

The frequency pulling due to distortion has been computed from Groszkowski's equation for several circuits operating at low frequencies. When the pulling is of the order of 0.5 per cent, the theory accounts for about four-fifths of it. The cause of the remaining discrepancy has not been determined.

Tests have been made on both free-running and crystal-controlled oscillator circuits at frequencies from 100 kc to 10 mc. Based on a one per cent change of power-supply voltage, frequency stabilities as good as 0.3 part per million were observed.
VI. PROGRAM FOR THE NEXT INTERVAL

1. The investigation of the relation between amplitude limiting and frequency stability will be continued.

2. Evaluation tests of practical oscillators will be continued.
VII. PERSONNEL

The key personnel working on this project during the period covered by this report are:

<table>
<thead>
<tr>
<th>Name</th>
<th>Position</th>
<th>Approximate Hours Worked</th>
</tr>
</thead>
<tbody>
<tr>
<td>B. J. Dasher, Sc.D.</td>
<td>Project Director</td>
<td>260</td>
</tr>
<tr>
<td>D. L. Finn, Ph.D.</td>
<td>Research Associate</td>
<td>100</td>
</tr>
<tr>
<td>W. B. Jones, Jr., Ph.D.</td>
<td>Research Associate</td>
<td>200</td>
</tr>
<tr>
<td>S. N. Witt, Jr., M.S.</td>
<td>Research Engineer</td>
<td>180</td>
</tr>
</tbody>
</table>

Mr. S. N. Witt, Jr., joined the project as a Research Engineer in January, 1954. He received his B.S. in E.E. from Tennessee Polytechnic Institute in June, 1950, and completed the requirements for the M.S. in E.E. in September, 1953, at the Georgia Institute of Technology where he is currently continuing his graduate studies. Mr. Witt was an Instructor in Electrical Engineering at Tennessee Polytechnic Institute from June, 1950, to June, 1951, at which time he became associated with the Engineering Experiment Station.

Respectfully submitted:

Approved:

B. J. Dasher
Project Director

W. B. Jones, Jr.
Research Associate

S. N. Witt, Jr.
Research Engineer

Paul K. Calaway, Acting Director
Engineering Experiment Station
VIII. BIBLIOGRAPHY


QUARTERLY REPORT NO. 4
PROJECT NO. 236-206

TRANSISTOR OSCILLATORS OF
EXTENDED FREQUENCY RANGE

By

B. J. DASHER and S. N. WITT, Jr.

CONTRACT NO. DA-36-039-sc-42712

DEPARTMENT OF THE ARMY PROJECT: 3-99-11-022
SIGNAL CORPS PROJECT: 142B

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### TABLE OF CONTENTS

<table>
<thead>
<tr>
<th>Section</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>I. PURPOSE</td>
<td>1</td>
</tr>
<tr>
<td>II. ABSTRACT</td>
<td>2</td>
</tr>
<tr>
<td>III. CONFERENCES</td>
<td>3</td>
</tr>
<tr>
<td>IV. FACTUAL DATA</td>
<td>4</td>
</tr>
<tr>
<td>A. Introduction</td>
<td>4</td>
</tr>
<tr>
<td>B. Amplitude Limiting</td>
<td>4</td>
</tr>
<tr>
<td>C. Evaluation of Oscillator Circuits</td>
<td>17</td>
</tr>
<tr>
<td>D. Free-Running Oscillators</td>
<td>23</td>
</tr>
<tr>
<td>E. Junction Transistor Oscillators</td>
<td>25</td>
</tr>
<tr>
<td>V. CONCLUSIONS</td>
<td>26</td>
</tr>
<tr>
<td>VI. PROGRAM FOR THE NEXT INTERVAL</td>
<td>27</td>
</tr>
<tr>
<td>VII. PERSONNEL</td>
<td>28</td>
</tr>
<tr>
<td>VIII. BIBLIOGRAPHY</td>
<td>29</td>
</tr>
</tbody>
</table>

This Report Contains 29 Pages
<table>
<thead>
<tr>
<th>Figure</th>
<th>Description</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.</td>
<td>Idealized Emitter Characteristic</td>
<td>8</td>
</tr>
<tr>
<td>2.</td>
<td>(a) Volt-ampere Characteristics</td>
<td>10</td>
</tr>
<tr>
<td></td>
<td>(b) Voltage Waveforms</td>
<td>10</td>
</tr>
<tr>
<td>3.</td>
<td>Variations of Type I Characteristic</td>
<td>12</td>
</tr>
<tr>
<td>4.</td>
<td>Normalized Performance Curves for Type I and Type II Oscillators</td>
<td>14</td>
</tr>
<tr>
<td>5.</td>
<td>400-kc Oscillator</td>
<td>17</td>
</tr>
<tr>
<td>6.</td>
<td>Change in Frequency for a Given Change in Voltage as a Function of Crystal Resistance for the Circuit of Figure 4.</td>
<td>19</td>
</tr>
<tr>
<td>7.</td>
<td>Equivalent Circuit of 400-kc Oscillator of Figure 5.</td>
<td>20</td>
</tr>
<tr>
<td>8.</td>
<td>Amplitude-and Distortion-Stabilized Version of Figure 5</td>
<td>21</td>
</tr>
<tr>
<td>9.</td>
<td>Transformer-Coupled 400-kc Oscillator</td>
<td>21</td>
</tr>
<tr>
<td>10.</td>
<td>Circuit for Detailed Low-Frequency Study of the Effect of Q, Z_o, and Other Factors on Stability.</td>
<td>23</td>
</tr>
</tbody>
</table>
I. PURPOSE

The purpose of this project is to evolve data and information regarding transistor oscillator circuitry and the degree of reliability and reproducibility attained with transistor parameter variations. It includes both analytical and experimental studies of crystal-controlled and free-running oscillators employing transistors exclusively as active elements.

The work is divided into three principal tasks as follows:

(1) the evaluation of crystal-controlled and free-running transistor oscillators from the point of view of reliability, size, weight, circuit requirements, frequency limitations, and other similar considerations;

(2) the development of new circuits and/or techniques specifically to take advantage of the unique characteristics of transistors regarding size, weight, power requirements, etc.; and

(3) the development of practical engineering data and information to serve as a basis for designing free-running and crystal-controlled junction- and point-contact transistor oscillators in the range 0.1 to 15 mc.
II. ABSTRACT

An analytical treatment of two-terminal oscillators is begun in this report. The analysis is based on a Fourier series representation in which the nonlinear negative resistance characteristic is approximated by means of straight-line segments. Instead of assuming the characteristics to be nearly linear the oscillation is assumed to be nearly sinusoidal. In the case of a series-mode oscillator this assumption implies a sinusoidal current which permits the voltage to be expressed as a Fourier series. The ratio of the fundamental component of voltage to the fundamental component of current is the effective input resistance which, when considered as a function of the current amplitude, permits the amplitude of oscillation to be determined for a specified external circuit. Applying Groszkowski's equation to the corresponding voltage wave yields the frequency of the oscillation. In this way the relation between amplitude stability and frequency stability can be studied. Universal curves are given for two elementary volt-ampere characteristics, one symmetrical and the other asymmetrical. Use of these curves is illustrated by means of numerical examples.

Evaluation tests described in Quarterly Report No. 3 have been continued, and the results of these tests are presented. In general, it is possible to adjust the circuit of a given crystal-controlled oscillator so that a voltage stability of a few parts in $10^8$ is obtained for a one percent change in supply voltage. The use of a diode clipping arrangement gives a circuit that is less sensitive to transistor parameters.
III. CONFERENCES

Mr. B. J. Dasher attended the 12th annual Frequency Control Symposium at Asbury Park, N. J., on April 12-14. Conferences were held with Mr. E. Gonzales and other SCEL representatives concerning the work of the project.
IV. FACTUAL DATA

A. Introduction

The results of a series of experiments performed for the purpose of finding out whether Groszkowski's theory of two-terminal oscillators may be advantageously applied to transistors have been described in previous reports. These experiments were carried out at low frequencies (1,000 cps) in order to eliminate as nearly as possible reactive effects associated with the transistor. The data obtained so far indicate that the Groszkowski equation relating frequency pulling to harmonic distortion does not account for all the observed frequency variation. (The term frequency pulling refers to the tendency of an oscillator to operate at a frequency different from the unity-power-factor frequency of its tank circuit.) Since the Groszkowski equation is based on the assumption that the negative-resistance element of the oscillator cannot store energy, it may be assumed that the transistor exhibits appreciable energy-storage effects even at 1,000 cps. It is planned to pursue this point further experimentally during the next quarter. In the meantime, some analytical studies of amplitude limiting have been started. These are presented in the next section.

The evaluation tests described in Quarterly Report No. 3 have been continued during this quarter and some circuits employing junction transistors have also been studied.

B. Amplitude Limiting

It has been pointed out frequently in previous reports that amplitude-limiting phenomena and frequency pulling caused by distortion are closely related in two-terminal oscillators. It is desirable, therefore, to
investigate some of the ways in which limiting takes place with particular reference to the associated distortion. It is very important to obtain the results in engineering terms if possible. One of the objections that can be made to classical treatments of nonlinear oscillators is that the analysis is not versatile from the engineering point of view. As an example, one might consider the isocline technique which was described briefly in Quarterly Report No. 1. It is possible, in principle at least, to determine the exact frequency of oscillation of a system by this method. Practically, however, the limitations of graphical analysis render it almost useless for studying frequency variations of the magnitude that must be considered here. What is more important is that the isocline diagram can not be easily related to quantities that can be observed or measured in the laboratory. When multi-loop circuits and high-order differential equations are involved, this method appears formidable indeed.

Groszkowski's equation, on the other hand, yields results in such terms as circuit Q, impedance, and harmonic distortion. These are down-to-earth quantities that really mean something. When it comes to practical matters, an approximate solution showing the relative importance of the significant variables is a great deal more useful than an exact solution submerged in mathematical details.

Another difficulty that arises in connection with nonlinear systems is the problem of classifying the nonlinear characteristics. One way of doing this is through power-series representation. A system is then characterized as having second-order curvature, third-order curvature, and so on. This procedure is very satisfactory for some purposes, but when the
order of curvature is high the number of variables to be considered is so large that it becomes unwieldy. Power-series representation is most useful for a nearly linear device. It has been seen, however, that a transistor oscillator operating on the "linear" part of a negative-resistance characteristic tends to depend on critical adjustments for satisfactory operation. When the operating range includes sharp bends in the characteristic, circuit adjustments are much less critical, amplitude stability is improved, and frequency stability is approximately the same as in the case of nearly linear operation. Consequently, there is a need for an analysis suited to this type operation.

One method of analysis that appears promising is based on a Fourier series representation in which the negative resistance characteristic is approximated by means of straight-line segments. Before describing this method it is appropriate to review the approximations and assumptions involved.

In the first place, instead of assuming the oscillation to be nearly linear it is assumed to be nearly sinusoidal. In the case of a series-mode oscillator this assumption implies a nearly sinusoidal current, and in the case of a parallel-mode oscillator it implies a nearly sinusoidal voltage. Second, energy storage in the nonlinear part of the circuit is neglected. (Some possible means for including energy storage have been considered but a discussion of these will be postponed because at this point it would only complicate matters.) Third, it is assumed that oscillations of constant amplitude occur when (a) the real part of the impedance of the nonlinear circuit equals the negative of the real part of the impedance of the resonator, both calculated at the fundamental frequency,
and (b) Groszkowski's equation is satisfied. Finally, it is convenient but not essential to assume the resonator to consist of a simple tuned circuit in which case it must have a reasonably high Q.

It has not been pointed out previously that the condition on the real parts of the impedances is an approximation. Actually, the fundamental requirement is that the energy dissipated per cycle in the resonator must equal the energy supplied by the active circuit, that is, by the negative resistance. If the current were perfectly sinusoidal, harmonic voltages would supply no power to the resonator and no approximation would be involved. Otherwise, it is probable that power supplied at the harmonic frequencies must be taken into account in the criterion for constant amplitude. This writer does not know of the existence of a complete treatment of this particular phase of the oscillator problem. For the present purpose the issue will be avoided by taking refuge in the assumption of nearly sinusoidal currents. Under these circumstances the effect of the harmonics must be small, whatever its nature.

The procedure for calculating the amplitude and frequency of an oscillator can now be described. The volt-ampere characteristics of the negative resistance and the constants of the tuned circuit are assumed to be known. For convenience, the discussion will be based on series-mode operation.

The volt-ampere characteristic to be considered is shown in Figure 1. This may be thought of as an idealized emitter characteristic for a point-contact transistor. The coordinates are the incremental current and voltage and the origin, therefore, represents the quiescent operating point.
The dotted line is the load line which represents the alternating-current resistance of the resonator, $R_0$. For the purpose of this illustration the direct currents are disregarded.

As the first step, assume a sinusoidal current having an amplitude corresponding to the intersection of the load line with the emitter characteristic. Call this amplitude $I_1$. Next, plot the voltage wave and determine the amplitude of the fundamental component, $V_1$, and the magnitude of the input resistance, $|R_1| = \frac{V_1}{I_1}$. If the load resistance is less than $R_1$, make a second calculation using a slightly larger value for $I_1$. Continue on in this way until a value of $I_1$ is found such that $|R_1| = R_0$, and then make a complete harmonic analysis of the corresponding voltage wave. Since the frequency of the current is assumed to be the resonant frequency of the tank, the impedance can be determined for each harmonic.
voltage and, consequently, an approximate harmonic analysis of the current wave can be found.

Next, the whole procedure is repeated using the distorted current wave instead of the pure sinusoid. If the current distortion is small, as has been assumed, the second voltage wave will differ very little from the first, and it is evident that this process will converge rapidly to a solution. Having found the voltage wave, Groszkowski's equation can be used to find the frequency and the problem is completely solved.

Theoretically, the above procedure can be refined so as to give as accurate a solution as is desired. Also, the actual volt-ampere characteristic can be used, since the straight-line approximation in no way affects the principles involved.

Actually, it is not proposed to solve oscillator problems by the method just described but the principles it illustrates together with the assumptions stated earlier form the basis for an analysis which is expected to yield important information about the relation between amplitude limiting and frequency stability. The success of this analysis hinges on two important points which should be restated at this time in somewhat more general terms. First, only an approximate solution is expected but the approximation will be good enough to predict pulling effects with adequate accuracy. Otherwise, it would be necessary to consider too many small details which would defeat the purpose of the method. Second, it is assumed that the characteristics of the tank circuit, whether or not it is a simple series circuit, are such that the current is very nearly sinusoidal. Under these circumstances the actual voltage wave is essentially
the same as the one obtained from a pure sinusoid. Finally, since the parallel-mode oscillator is the dual of the series-mode oscillator the results apply to both types when the appropriate exchange of variables is made.

The notation to be used is illustrated in Figure 2. Two volt-ampere characteristics to be considered are shown in part (a) and the corresponding voltage waveforms are shown in part (b). It is convenient to plot \(-v\) instead of \(v\) and to let \(\omega_0 t = \Theta\) since the algebra is thereby simplified. The solid lines apply to an asymmetrical characteristic, such as the one in Figure 1, and the dotted lines show the modifications for a symmetrical characteristic which represents certain types of push-pull oscillators. The asymmetrical characteristic is designated type I and the symmetrical characteristic is designated type II. \(R_o\) is the
resistance corresponding to the linear part of the characteristic. Let
\[ i = K I_o \cos \theta = I_1 \cos \theta. \]

Then, for the type I characteristic,
\[ -v = \begin{cases} R_o I_o, & 0 \leq \alpha \leq \theta \\ K R_o I_o, & \alpha \leq \theta \leq \pi \end{cases} \quad (2) \]

and
\[ -v(-\theta) = -v(\theta). \quad (3) \]

Note that \( K \cos \alpha = 1. \)

In terms of its Fourier series,
\[ -v = \frac{a_0}{2} + \sum_{n=1}^{\infty} a_n \cos n\theta \quad (5) \]

with
\[ a_n = \frac{1}{\pi} \int_{0}^{2\pi} (-v) \cos n\theta \, d\theta. \quad (6) \]

Substituting for \(-v\) from equation 2 and taking advantage of the fact that \(-v\) is an even function of \( \theta \) gives
\[ a_n = \frac{2R_o I_o}{\pi} \left[ \int_{0}^{\alpha} \cos n\theta \, d\theta + \int_{\alpha}^{\pi} K \cos \theta \cos n\theta \, d\theta \right]. \quad (7) \]

Equation 7 yields:
\[ a_1 = K R_o I_o \left[ 1 + \frac{\sin 2\alpha - 2\alpha}{2\pi} \right] = -v_1 \quad (8) \]
\[ a_n = \frac{K R_o I_o}{\pi n} \left[ \frac{\sin (n + 1)\alpha}{n + 1} - \frac{\sin (n - 1)\alpha}{n - 1} \right] = -v_n \quad n = 2, 3, 4, \ldots (9) \]

For the type II characteristic a similar analysis gives:
\[ a_1 = K R_o I_o \left[ 1 + \frac{\sin 2\alpha - 2\alpha}{\pi} \right] \quad (10) \]
A useful interpretation of these results requires a knowledge of the manner in which the volt-ampere characteristics change under the influence of oscillator disturbances. The ways in which a type I characteristic may vary are illustrated in Figure 3. A change from a to b changes $I_0$ but $K$ will not change if $R_1$ remains fixed. In other words, this change will affect only the amplitude of the oscillations. A change from a to c changes $R_0$ and so, in general, would affect both amplitude and distortion. Reference to equations 8 and 9 shows that the relative distortion depends on $\alpha$ and hence on $K$ but not on $I_0$. Therefore, a change from a to c is equivalent to changing the external resistance except for a difference in amplitude. Likewise, a change from a to d is equivalent to a change of the external resistance combined with a change of amplitude. Therefore, the
stability of an oscillator with respect to changes in its negative-resistance characteristic is essentially the same as its stability with respect to a change of the external (linear) circuit resistance.

Equation 8 yields

$$\frac{-R_1}{R_o} = \left[1 + \frac{\sin 2\alpha - 2\alpha}{2} \right]$$

(12)

for type I and equation 10 gives

$$\frac{-R_1}{R_o} = \left[1 + \frac{\sin 2\alpha - 2\alpha}{\pi} \right]$$

(13)

for type II. These results are plotted in Figure 4 as functions of \(K\).

The frequency shift caused by distortion depends also on the properties of the resonator. In the case of a single tuned circuit the frequency is given by

$$f = f_0 \left[1 - \frac{1}{2\omega_0^2} \sum_{n=1}^{\infty} \left(\frac{n}{\pi}\right)^2 \frac{n^2}{n^2 - 1}\right].$$

(14)

Hence, the quantity inside the summation sign is seen to be a kind of "pulling" factor for the oscillator. Call this quantity \(P^2\). Then equation 14 becomes

$$f = f_0 \left[1 - \frac{P^2}{2\omega_0^2}\right].$$

(15)

The pulling factors for the type I and type II oscillators are also plotted in Figure 4. Harmonics only through the eighth have been included in calculating \(P^2\). Neglecting the higher harmonics introduces appreciable errors for \(K > 6\).

The use of these performance curves may be illustrated by means of a few numerical examples.

1. A series-mode oscillator having a type I characteristic has a negative resistance of 1,000 ohms over its linear part. The "break-point"
Figure 4. Normalized Performance Curves for Type I and Type II Oscillators.
I₀ corresponds to an incremental current of 0.5 ma. The tank circuit has a resistance of 800 ohms, a Q of 10, and is resonant at 1,000 cps. Find the amplitude and frequency of the oscillations.

Solution:

\[-R_0/R_0 = 0.8\] and Figure 4 gives \( K = 2, \ P^2 = 0.05 \).

Therefore,

\[ V_1 = 2 \times 0.5 \times 10^{-3} \times 800 = 0.8 \text{ volt (peak)} \]

and

\[ f = 1,000 \left( 1 - \frac{0.05}{2 \times 100} \right) = 1,000 \left( 1 - 0.00025 \right) = 999.75 \text{ cps} \]

2. A ten per cent increase in the supply voltage of the oscillator in example 1 changes its parameters to \(-R_0 = 1200 \text{ ohms}, I_0 = 0.5 \text{ ma}.\) Find the new amplitude and frequency.

Solution:

\[-R_0/R_0 = \frac{800}{1200} = 0.666.\] Figure 4 gives \( K = 3.8 \) and \( P^2 = 0.12 \).

Therefore,

\[ V_1 = 3.8 \times 0.5 \times 10^{-3} \times 800 = 1.52 \text{ volts (peak)} \]

and

\[ f = 1,000 \left( 1 - \frac{0.12}{2 \times 100} \right) = 1,000 \left( 1 - 0.0006 \right) = 999.4 \text{ cps} \]

3. If the oscillator in example 1 has no type II characteristics find the amplitude and frequency.

Solution:

As before,

\[-R_0/R_0 = 0.8, \] but now \( K = 1.45 \) and \( P^2 = 0.023 \). Therefore,

\[ V_1 = 1.45 \times 0.5 \times 10^{-3} \times 800 = 0.58 \text{ volts (peak)} \]
and

\[ f = 1,000 \left(1 - \frac{0.023}{2 \times 100}\right) = 1,000 \left(1 - 0.00012\right) = 999.88 \text{ cps.} \]

4. Changing the supply voltage for this oscillator as in example 2 gives the following results.

\[ -\frac{R_1}{R_0} = 0.666, \quad K = 1.8 \quad \text{and} \quad P^2 = 0.05. \]

Therefore,

\[ V_1 = 1.8 \times 0.5 \times 10^{-3} \times 800 = 0.72 \text{ volts (peak)} \]

\[ f = 1,000 \left(1 - \frac{0.05}{2 \times 100}\right) = 1,000 \left(1 - 0.00025\right) = 999.75 \text{ cps}. \]

These calculations indicate that the symmetrical oscillator has an amplitude stability about 40 per cent better than the asymmetrical circuit and a frequency stability nearly three times as good.
C. Evaluation of Oscillator Circuits

The work during this report period was continued along the same lines as previously reported. As noted in Quarterly Report No. 3, crystal-controlled oscillations have been obtained at frequencies as high as 1 mc only by using LC tuned circuits in addition to the crystal. It was considered desirable in the interest of stability to further simplify the circuits by eliminating as many components as possible. To do this it was necessary to operate at a lower frequency. The frequency chosen for circuit designs was 400 kc at which frequency an average W.E. type 1698 transistor has an alpha greater than one.

After investigating the effects of changing the collector and emitter voltages, the circuit and the parameter values shown in Figure 5 were chosen for a particular transistor and a particular crystal. This circuit showed less than one cycle per second variation in frequency at 400 kc when E was changed from 30 to 50 volts. Although temperature sensitivity was not measured quantitatively it seemed to be relatively small.

![Figure 5. 400-kc Oscillator.](image-url)
With transistors and crystals other than the particular ones used for the circuit design, frequency variations as great as 50 cycles per second were observed when E was changed from 30 to 50 volts. Variations in crystal characteristics had more effect on frequency stability than did variations in transistor parameters. It was found, however, that by changing the resistance values, the stability could be made very good for almost any particular transistor and crystal combination.

For the circuit of Figure 5 and the particular transistor for which it was designed, Figure 6 shows frequency variations for a 20-volt change in E plotted as a function of equivalent series resistance, \( R_e \), of the crystal. The points indicate that some functional relation does exist between the variables.

A study of variations in frequency (with changes in applied voltage, E) as a function of collector resistance, \( R_c \), was also made. In some cases it was found that the change in frequency with voltage could be reduced to zero by the proper selection of \( R_c \). However, this was true only for certain transistor samples. Thus, this procedure does not appear to be a complete solution to the stability (with respect to voltage) problem. The base resistance, \( R_b \), and emitter resistance, \( R_e \), also have some effect on the stability since they determine the emitter-base current ratio. The improvement that could be obtained by varying these constants was too small to warrant serious consideration.

It is believed that an analytical study of this circuit may lead to a better design basis. For this purpose it is desirable to have more accurate information on the transistor parameters than is now available. At present, an analytical study is being undertaken on the basis of the equivalent circuit shown in Figure 7.
Figure 6. Change in Frequency for a Given Change in Voltage as a Function of Crystal Resistance for the Circuit of Figure 4.
Another approach to stabilizing the circuit of Figure 5 has been investigated. The resulting circuit, shown in Figure 8, is based on the evidence that better stability is obtained if the amplitude and waveform distortion of the feedback voltage are maintained constant. This is the purpose of the diode clipping arrangement. This modification resulted in stabilities not quite as good as the best obtained for Figure 5, but the over-all stability for different crystals and transistors was greatly improved without adjusting the circuit constants. Out of the 10 crystals used with a particular transistor the best stability obtained was 0.52 parts in $10^7$ for one per cent voltage change. The poorest stability was 0.98 parts in $10^7$ for one per cent voltage change. It was found that only the more active transistors (alpha greater than or equal to 2 at 400 kc) would oscillate in this circuit. Those that did oscillate all showed about the same relative frequency stability.
One of the effects which was considered undesirable in the circuit of Figure 5 was the shunt capacitance of the crystal. The circuit of Figure 9 was designed to effectively eliminate this capacitance by the proper adjustment of $C$.

Figure 9. Transformer Coupled 400-kc Oscillator.
Because of increased feedback due to the use of transformer coupling, this circuit produced oscillations at much lower supply voltages than did the previous circuits. By varying $R_k$, frequency variation with changes in supply voltage could be minimized for particular crystals and transistors. In one case the frequency variation was only 0.5 cycles per second at 400 kc for changes in supply voltage from 10 volts to 25 volts. Based on a nominal supply voltage of 15 volts, this amounted to $1.3 \times 10^{-8}$ for one per cent change in supply voltage. However, more drift was caused by temperature change than with the circuit of Figure 5. The use of diode clipping with this circuit tended to produce free-running rather than better stability. Thus, the desirable features of Figure 9 (less change in frequency with applied voltage and ability to oscillate with less active transistors) could not be conveniently combined with those of Figure 8.

Again an attempt was made to obtain accurate measurements of the transistor parameters in order that a more complete analysis of the foregoing circuits could be made. The test set described in Quarterly Report No. 1 was found inadequate because of the high frequencies involved. The work on these circuits was temporarily discontinued. After a better test set has been constructed, it is hoped that accurate design criterion can be formulated which will enable the design of oscillators using these basic circuits such that best stability will be obtained. This may result in some combination of the desirable features of the circuits discussed above.

Some two-transistor oscillator circuits using negative feedback for stabilization were tried. The results were so inconclusive that an account of the circuits will not be given here.


B. Free-Running Oscillators

A review of all of the previously constructed circuits indicated that the circuit shown in Figure 14 of Quarterly Report No. 3 shows the greatest promise for high-frequency oscillations. It is also better suited to the free-running type of control. The use of a free-running oscillator makes possible the control of $Q$ and the resonant impedance of the frequency-determining element. The circuit has the additional advantage that very few components are required. Therefore, this circuit was chosen for a detailed study at frequencies where both the magnitude and phase of alpha are relatively constant. The simplified version of this circuit appears in Figure 10.

![Circuit Diagram](image)

Figure 10. Circuit for Detailed Low-Frequency Study of the Effect of $Q$, $Z_0$, and Other Factors on Stability.
This circuit was first operated at approximately 1 kc with various values of $R_e$ and $R_b$. Best stability with an LC parallel resonant circuit having a $Q$ of 6 was obtained with $R_e$ equal to 4700 ohms and $R_b$ equal to 1700 ohms. The resonant impedance, $R_o$, of the LC circuit was about 20,000 ohms. Under these conditions, the frequency varied from 900 cycles per second to 891 cycles per second as $E$ was changed from 5 volts to 20 volts. Decreasing the $Q$ to 3 and the resonant impedance to 10,000 ohms caused a greater variation in frequency. However, the data indicated that possibly an optimum value of $Q$ or an optimum value of resonant impedance exists for any particular set of circuit components. Due to the fact that a value of $Q$ greater than 6 could not be obtained at this low frequency, the decision was made to conduct further tests at frequencies near 25 kc. This frequency is still sufficiently low that the phase and amplitude characteristics of alpha can be neglected for the W.E. type 1698 transistor.

Initial data at 25 kc indicated that for a resonant impedance of 15,000 ohms the value of $Q$ for best stability is near 15. Values of $Q$ near 20 and near 10 both resulted in greater frequency variation for a given change in $E$.

Since sufficient data has not yet been obtained, the above conclusion concerning the existence of an optimum $Q$ can by no means be stated as a fact. This is especially true since conventional theory indicates that best stability should accompany the maximum obtainable $Q$. 
E. Junction Transistor Oscillators

Some circuits employing junction transistors have been designed and tested. Since this work has been in progress for only a few weeks a discussion of it will not be given in this report.
V. CONCLUSIONS

The oscillator analysis presented in this report promises to help materially in understanding the basic mechanisms of two-terminal oscillators. Although it is theoretically possible to obtain exact solutions by this method through the use of successive approximations, this is not its main purpose. Its usefulness arises from the fact that it helps to guide oscillator design by showing what are the important factors governing performance.

Most of the tests that have been reported so far have been made using one or two types of point contact transistors in a variety of circuits. Under these circumstances it was not practical to make complete tests. It is planned now to make thorough studies on the most promising of these circuits. These tests will include such factors as temperature effects, power output, maximum practical frequencies, and others that have been largely overlooked up to this time.
VI. PROGRAM FOR THE NEXT INTERVAL

1. The analytical work will be continued. Curves similar to those of Figure 4 will be prepared for several other types of volt-ampere characteristics.

2. Evaluation tests will be continued. It is expected that these tests will include the effect of varying ambient temperature and other factors not heretofore considered quantitatively.

3. Circuits employing junction transistors will be studied.
The key personnel working on this project during the period covered by this report are:

<table>
<thead>
<tr>
<th>Name</th>
<th>Position</th>
<th>Approximate Hours Worked</th>
</tr>
</thead>
<tbody>
<tr>
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</tbody>
</table>

Mr. W. B. Warren, Jr., joined the project as a research assistant in April 1954. He received his B. S. in E. E. from the Georgia Institute of Technology in March 1953 and is currently continuing in the Graduate Division. Mr. Warren was an Electronic Technician in the U. S. Navy for three years. He has been on the staff of the Engineering Experiment Station since January 1953.

Respectfully submitted:

Approved:

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VIII. BIBLIOGRAPHY


QUARTERLY REPORT NO. 5
PROJECT NO. 236-206

TRANSISTOR OSCILLATORS OF
EXTENDED FREQUENCY RANGE

By

B. J. DASHER, D. L. FINN, S. N. WITT, JR.,
W. B. WARREN, JR., and T. N. LOWRY

CONTRACT NO. DA-36-039-sc-42712

DEPARTMENT OF THE ARMY PROJECT: 3-99-11-022
SIGNAL CORPS PROJECT: 142B

SEPTEMBER 30, 1954
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# TABLE OF CONTENTS

<table>
<thead>
<tr>
<th>Section</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>I. PURPOSE</td>
<td>1</td>
</tr>
<tr>
<td>II. ABSTRACT</td>
<td>2</td>
</tr>
<tr>
<td>III. CONFERENCES</td>
<td>3</td>
</tr>
<tr>
<td>IV. FACTUAL DATA</td>
<td>4</td>
</tr>
<tr>
<td>A. Introduction</td>
<td>4</td>
</tr>
<tr>
<td>B. Design Curves for Negative-Resistance Oscillators</td>
<td>5</td>
</tr>
<tr>
<td>C. Oscillator Design Problem</td>
<td>16</td>
</tr>
<tr>
<td>D. Free-Running Oscillators</td>
<td>28</td>
</tr>
<tr>
<td>E. Junction Transistor Oscillators</td>
<td>36</td>
</tr>
<tr>
<td>V. CONCLUSIONS</td>
<td>44</td>
</tr>
<tr>
<td>VI. PROGRAM FOR THE NEXT INTERVAL</td>
<td>45</td>
</tr>
<tr>
<td>VII. PERSONNEL</td>
<td>46</td>
</tr>
</tbody>
</table>

This Report Contains 46 Pages
LIST OF FIGURES

<table>
<thead>
<tr>
<th>Figure Description</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>1. Volt-Ampere Characteristics and Voltage Waveforms</td>
<td>5</td>
</tr>
<tr>
<td>2. Volt-Ampere Characteristics and Voltage Waveforms (M ≠ 0)</td>
<td>6</td>
</tr>
<tr>
<td>3. Normalized Performance Curves for a Type I Oscillator</td>
<td>8</td>
</tr>
<tr>
<td>4. Normalized Performance Curves for a Type II Oscillator</td>
<td>9</td>
</tr>
<tr>
<td>5. Normalized Performance Curves for a Type I Oscillator</td>
<td>11</td>
</tr>
<tr>
<td>6. Normalized Performance Curves for a Type II Oscillator</td>
<td>12</td>
</tr>
<tr>
<td>7. Normalized Performance Curves for a Type I Oscillator (Expanded)</td>
<td>13</td>
</tr>
<tr>
<td>8. Normalized Performance Curves for a Type II Oscillator (Expanded)</td>
<td>14</td>
</tr>
<tr>
<td>9. Circuit for Obtaining Input Characteristics</td>
<td>17</td>
</tr>
<tr>
<td>10. Effect of Collector Voltage on Input Characteristic</td>
<td>18</td>
</tr>
<tr>
<td>11. Effect of Emitter Resistance on Input Characteristic</td>
<td>19</td>
</tr>
<tr>
<td>12. Effect of Collector Resistance on Input Characteristic</td>
<td>20</td>
</tr>
<tr>
<td>13. Characteristics Employed for Oscillator Design</td>
<td>21</td>
</tr>
<tr>
<td>14. 1-Kc Oscillator</td>
<td>22</td>
</tr>
<tr>
<td>15. 25-Kc Free-Running Oscillator</td>
<td>29</td>
</tr>
<tr>
<td>16. 25-Kc Oscillator Using Diode Amplitude Limiting</td>
<td>31</td>
</tr>
<tr>
<td>17. Frequency Variation as a Function of Supply Voltage for Different Values of $R_e$</td>
<td>33</td>
</tr>
<tr>
<td>18. Frequency Variation as a Function of Supply Voltage for Different Values of $R_e$ (Normalized at 25 Volts)</td>
<td>34</td>
</tr>
<tr>
<td>19. Circuit Showing Conditions on Loop Phase Shift</td>
<td>37</td>
</tr>
<tr>
<td>20. LC Oscillator</td>
<td>37</td>
</tr>
<tr>
<td>21. Alpha Characteristics of a Typical Type CK 720 Junction Transistor</td>
<td>38</td>
</tr>
<tr>
<td>22. Crystal-Controlled Oscillator</td>
<td>39</td>
</tr>
</tbody>
</table>

(Continued)
LIST OF FIGURES (Continued)

Page

23. Variation of Frequency with Collector Supply Voltage for Various Values of Emitter Supply Voltage .......................... 40

24. Compensated Crystal Oscillator With Single Battery Supply .......... 41

25. Variation of Frequency with Supply Voltage for Two Values of R .. 43
I. PURPOSE

The purpose of this project is to evolve data and information regarding transistor oscillator circuitry and the degree of reliability and reproducibility attained with transistor parameter variations. It includes both analytical and experimental studies of crystal-controlled and free-running oscillators employing transistors exclusively as active elements.

The work is divided into three principal tasks as follows:

(1) the evaluation of crystal-controlled and free-running transistor oscillators from the point of view of reliability, size, weight, circuit requirements, frequency limitations, and other similar considerations;

(2) the development of new circuits and/or techniques specifically to take advantage of the unique characteristics of transistors regarding size, weight, power requirements, etc.; and

(3) the development of practical engineering data and information to serve as a basis for designing free-running and crystal-controlled junction- and point-contact transistor oscillators in the range 0.1 to 15 mc.
II. ABSTRACT

The analysis of two-terminal oscillators begun in Quarterly Report No. 4 is continued. Curves relating the pulling factor, $P^2$, and the resistance ratio, $-R_1/R_0$, have been computed for more general types of negative-resistance characteristics. The theory has been applied to the design of a parallel-mode oscillator at 1000 cps. By actual performance tests the theory predicted the frequency stability with respect to supply voltage variations to within a factor of about three and the amplitude stability to within a factor of less than two. Since the measured frequency stability was approximately 15 parts in $10^7$ for one per cent change in supply voltage and the analysis involves several approximations, these results are considered to be very good.

Experimental studies of a basic parallel-mode grounded-collector oscillator circuit are described, with emphasis on the effects of changing transistors. Three related oscillator circuits employing junction transistors are described, one of which operates at five times the alpha cut-off frequency of the transistor employed. A method of bias compensation is described and applied to a crystal-controlled oscillator, resulting in frequency stabilities of less than one part in $10^7$ for one per cent change in supply voltage.
III. CONFERENCES

Mr. E. Gonzales and Mr. Hellstrom of the SCEL visited Georgia Tech August 10, 11 and 12, 1954. The objectives of the project and the present status of the work were reviewed. The desirability of obtaining transistors of recent design suitable for higher frequencies was discussed. It was recommended that Georgia Tech be enabled to purchase such transistors when they cannot be supplied by SCEL. A procedure for implementing this recommendation has subsequently been established by an amendment to the contract.
IV. FACTUAL DATA

A. Introduction

An analytical treatment of two-terminal oscillators was begun in Quarterly Report No. 4. This analysis is based on a Fourier-series representation in which the nonlinear negative-resistance characteristic is approximated by straight-line segments. Instead of assuming the characteristic to be nearly linear, the oscillation is assumed to be nearly sinusoidal. Through the use of successive approximations, it is theoretically possible to obtain exact solutions to oscillator problems by this method, but that is not its main purpose. Its main purpose is to provide a deeper insight into the principles of negative-resistance oscillators. At the same time, it provides results of sufficient accuracy for most engineering purposes with a minimum of effort.

The work described in the last report has been extended by compiling additional data for the pulling factors and resistance ratio for various idealized volt-ampere characteristics. These data do not yet include all cases that may be of interest. However, the present plan is to try out the results already obtained at higher frequencies before any further extension of the work neglecting internal phase-shifts in the transistor. The possibility of modifying the method so that it can be applied to junction transistors must also be investigated.

Heretofore the discussion of two-terminal oscillators has been presented in terms of series-mode oscillators. Unfortunately this type circuit has some serious practical limitations. The most important limitation so far observed is a result of the unavoidable distributed capacitance of inductors. Increasing the Q of a series resonant circuit while keeping its resistance constant requires increasing the inductance. The resulting small value of the series
capacitor and the relatively large coil capacitance makes the circuit susceptible to a variety of parasitic oscillations. In addition, the output voltages tend to contain a large proportion of harmonics, a situation which is usually undesirable. For these reasons, more and more attention has been given to parallel-mode circuits. The theory so far developed has been applied to the design of a parallel-mode oscillator. The details of this design and the actual performance data are presented in this report.

Experimental work at high frequencies also has been continued during this quarter and several circuits employing junction transistors have been tested.

B. Design Curves for Negative-Resistance Oscillators

Figure 4 of Quarterly Report No. 4 shows normalized performance curves for two basic types of negative-resistance oscillators. These oscillators are distinguished by the nature of their volt-ampere characteristics which are shown in Figure 1a. The type-I characteristic consists of two straight-line

Figure 1. (a) Volt-Ampere Characteristics. (b) Voltage Waveforms.
segments, and amplitude limiting is assumed to be produced by asymmetrical clipping of the voltage wave. (Note that $-v$ is plotted instead of $v$ and the negative-resistance part of the characteristic is, therefore, represented by the positive-slope segment of the curve.) The type-II characteristic consists of three straight-line segments, and in this case limiting is obtained by symmetrical clipping of the voltage wave.

Figure 2a shows how these characteristics are modified to obtain more general results, and Figure 2b shows the corresponding voltage wave. The asymmetrical case is still called a type-I characteristic, and the symmetrical case is called type II.

![Figure 2. (a) Volt-Ampere Characteristics. (b) Voltage Waveforms $(M=0)$](image)

In Figure 2a the part of the volt-ampere characteristic that lies outside the negative-resistance region represents a linear positive resistance of value $MR_o$, $R_o$ being the magnitude of the negative resistance. A negative value of $M$ corresponds to a resistance that is negative over the whole range but has
different values in the two regions. The characteristic in Figure 1a is seen to correspond to \( M = 0 \).

A family of performance curves for a type-I oscillator is shown in Figure 3, and a family for a type-II oscillator is shown in Figure 4. It should be remembered that the pulling-factor curves depend on the tank-circuit configuration as well as on the distortion and that the curves shown here apply to a simple tuned circuit.

There are so many variables to consider in the design and operation of an oscillator that it is difficult to determine the real significance of these curves. They do not point clearly to any particular collection of "do's" and "don'ts" for oscillator designers. On the other hand, they certainly do show the way in which amplitude stability and frequency stability are related in negative-resistance oscillators and a few observations in this connection are appropriate.

As explained in the previous report, the curves of Figures 3 and 4 may be used in the following way. Variations of transistor characteristics are equivalent to variations of \(-R_1/R_0\) and \(I_o\). If both these quantities vary appreciably (as the result of a variation of temperature, supply voltage, or other factors) the variation of \(I_o\) must be taken into account in determining amplitude variations. On the other hand, the pulling factor, \(P^2\), depends on \(K\) but not \(I_o\). Hence, frequency stability can be investigated without regard to variations of \(I_o\). These factors are considered in detail in the next section in which a specific design problem is discussed. For the present purpose it will be convenient to assume that variations of \(I_o\) are comparatively small and, hence, negligible.
Figure 3. Normalized Performance Curves for a Type I Oscillator.
Figure 4. Normalized Performance Curves for a Type II Oscillator.
Suppose, now, that various values of \(-\frac{R_1}{R_0}\) are encountered. For any selected value the corresponding value of \(K\) is found from the curves, and this \(K\) value yields the corresponding \(P^2\) value. In this way it is found that for small \(M\), \(P^2\) varies slowly for values of \(-\frac{R_1}{R_0}\) near 0.7, but \(K\) varies rapidly. For large \(M\), the converse is true. Thus, the large \(M\)-values lead to relatively good amplitude stability and relatively poor frequency stability whereas the small \(M\)-values lead to relatively good frequency stability and relatively poor amplitude stability.

Another important conclusion that can be drawn is that \(-\frac{R_1}{R_0}\) should be as close to unity as possible. This requirement is off-set by the fact that small variations of \(-\frac{R_0}{R_0}\) may cause the oscillations to die if a value too close to unity is used for design center. In this case, poor amplitude stability may result from excessive variations of \(-\frac{R_1}{R_0}\). A value between 0.7 and 0.8 appears to be a good compromise for general-purpose design. In every case the best value can be found only by examining the actual range of variation of parameters caused by the disturbing environment.

These conclusions are more clearly evident from the curves in Figures 5 and 6. In these curves, \(P^2\) is plotted as a function of \(K\). Two families of curves are given in each case: one family with \(M\) as a parameter and the other for \(-\frac{R_1}{R_0}\) as a parameter. Variations of \(-\frac{R_1}{R_0}\) correspond to motion along the \(M = \text{constant}\) lines. Since this form of presentation of the data is probably more useful than that of Figures 3 and 4, the parts of the curves for \(K\) values between one and two are shown to an expanded scale in Figures 7 and 8. All of these curves may be used for both series-mode and parallel-mode oscillators. For the latter, \(-\frac{R_1}{R_0}\) is replaced by \(-\frac{G_1}{G_0}\) and \(K\) represents the voltage amplitude ratio instead of the current amplitude ratio.
Figure 5. Normalized Performance Curves for a Type I Oscillator.
Figure 6. Normalized Performance Curves for a Type II Oscillator.
Figure 7. Normalized Performance Curves for A Type I Oscillator (Expanded).
Figure 8. Normalized Performance Curves for a Type II Oscillator (Expanded).
The next question that must be considered is that of the actual shapes of volt-ampere characteristics of transistors and their relation to circuit design. Actual characteristics depend not only on the transistors themselves but also on the external circuits other than the tank circuit. A survey of the possibilities has been made for a particular type of transistor, and it is probably permissible to assume these results to be typical. At first, it was hoped that this information could be obtained from the published curves. Unfortunately, the data in the critical regions are given with very poor accuracy and so it was necessary to make point-by-point measurements. Some possibilities for determining the most important characteristics from the published curves have been considered, but their usefulness has not been determined so they will not be presented in this report. Figures 10, 11, and 12 show the general effects of various circuit parameters. It was desired to find characteristics that could be approximated with reasonable accuracy by the straight-line segments so that the theory already developed could be tested on a design problem. This problem is presented in the next section. The results are considered to be very satisfactory. In spite of the many approximations that must be made, the amplitude of the oscillations was correctly predicted to within a factor of two, and the frequency stability with respect to supply voltage was predicted to within a factor of about three. It is probable that a more exact analysis of the available data would yield even better results. This has not been attempted for two reasons: First, the order of magnitude of the frequency variation (50 parts per million) is such that it would be necessary to verify the linearity of the tuning capacitor; and second, it is desirable to explore the problem at higher frequencies before considering any further refinements.
Since the objectives of the project are directed toward high frequencies, it is necessary to conclude the study of nonlinear effects, at least temporarily, with this report. The results so far obtained have already proved useful in gaining a clearer understanding of the mechanism of oscillation, and further study along these lines should prove fruitful.

C. Oscillator Design Problem

The analysis presented previously has been applied to the design of a two-terminal, parallel-mode oscillator. Information obtained from analysis of idealized characteristics served as a guide in the selection of parameter values and prediction of resulting performance. The common-emitter configuration is best suited to parallel-mode operation with point-contact transistors. Therefore, to make use of the method described, some information is needed on the negative-resistance characteristic of a point-contact transistor operating with common emitter.

Useful input characteristics for the common-emitter configuration cannot readily be obtained from the typical characteristics published by manufacturers. Such a technique requires a great deal of graphical manipulation, and the results obtained are unreliable because of discrepancies between the published data and the behavior of any particular transistor. For this design problem, input characteristics were obtained from point-by-point measurements of a W.E. Type 1768 point-contact transistor. Of course, such measurements may not represent a practical method for routine design problems, but the information obtained in this way may indicate the appropriate methods for later use.

The circuit for obtaining input characteristics is shown in Figure 9.
The resulting $I_b - V_1$ characteristic exhibits a well-defined negative conductance region suitable for use in a parallel-mode oscillator. Variations in shape and orientation of the characteristic with changes in collector voltage, emitter resistance and collector resistance are shown in Figures 10, 11 and 12. The operating region selected corresponds to a collector voltage of 8 volts, an external emitter resistance of 1000 ohms and an external collector resistance of zero. This combination of parameters afforded the best compromise between linearity of the negative-resistance region and stability of characteristics with changes in collector voltage among the various combinations investigated.

Figure 13 presents an expanded portion of the characteristics for collector voltages of six, eight and ten volts. Actual dynamic curves for the transistor would be displaced vertically from those of Figure 11 by a small amount, because the measurements were taken with the transistor "cold" (by rapid transition from low to high currents, as a means of reducing discont...
Figure 10. Effect of Collector Voltage on Input Characteristic.
Figure 11. Effect of Emitter Resistance on Input Characteristic.
Figure 12. Effect of Collector Resistance on Input Characteristic.
Figure 13. Characteristics Employed for Oscillator Design.
nuities due to heating). However, the shape of each characteristic employed is sufficiently accurate for present purposes.

A comparison of Figure 1a with Figure 13 indicates a similarity between the idealized type-I characteristic and the actual characteristics for $V_1$ values between plus one and minus two volts. For values of $V_1$ more negative than two volts, the actual curves depart considerably from the idealized curve, so the operating point was chosen to maintain oscillation in the more linear region. The point $P_1$ shown on Figure 13 corresponds to a base bias voltage of -1.0 volt and a bias current of 1.35 milliamperes. These bias conditions were obtained by the use of a voltage divider across the collector supply, as shown in Figure 14.

![1-Kc Oscillator](image)

Figure 14. 1-Kc Oscillator.
The variation in bias voltage due to changes in supply voltage depends jointly on the values of $R_1$ and $R_2$ and on the displacement of transistor characteristics. The dashed lines of Figure 13 represent bias characteristics, for which the slope is given by:

$$G_{\text{source}} = -\frac{R_1 + R_2}{R_1 R_2}$$

and the voltage intercepts for zero current are given by:

$$V_{\text{intercept}} = \frac{R_1 E_{cc}}{R_1 + R_2}.$$  

For reasons discussed below, a value of 100 ohms was selected for $R_1$. $R_2$ is correspondingly determined by:

$$R_2 = \frac{|E_{cc} - V_1| R_1}{|V_1| - I_b R_1}$$

which for the selected operating point becomes:

$$R_2 = \frac{(8 - 1) 100}{1 - (1.35 \times 10^{-3}) 100} = 809 \text{ ohms}.$$  

The slope of the bias characteristic is, therefore:

$$G_{\text{source}} = -\frac{100 + 809}{100 \times 809} = -11.24 \text{ millimhos (89 ohms).}$$

The intercepts corresponding to the three collector voltages are:

- $V_{\text{intercept}}$ for 6 volts = $\frac{100}{909} \times -6 = -0.66v$
- $V_{\text{intercept}}$ for 8 volts = $\frac{100}{909} \times -8 = -0.88v$
- $V_{\text{intercept}}$ for 10 volts = $\frac{100}{909} \times -10 = -1.10v$. 

-23-
Intersections of the bias and transistor characteristics define points $P_2$ and $P_3$ of Figure 13, which represent the quiescent operating points for supply-voltage variations of ±25 per cent from the normal eight volts.

Graphical measurements from Figure 13 of the transistor characteristic slopes at points $P_1$, $P_2$ and $P_3$ indicate incremental negative conductances as follows:

- $G_{o1} = -1.29$ millimhos (775 ohms)
- $G_{o2} = -1.38$ millimhos (725 ohms)
- $G_{o3} = -1.21$ millimhos (826 ohms)

The slope of the positive conductance region of each characteristic (for positive values of $V_1$) is approximately 0.05 millimhos, corresponding to a positive resistance of 20 kilohms. The ratio of positive and negative slopes is therefore 0.04, and $M$ may be considered zero for this case.

Predicted behavior of the oscillator for different values of load conductance may be observed in Figure 5, by regarding the ratios $-R_1/R_o$ as equivalent to $-G_1/G_o$ in the parallel-mode case. Each intersection of a $-R_1/R_o$ curve with the $M = 0$ curve represents a possible operating condition, with corresponding values of $P^2$ and $K$. Since the extreme values for $G_o$ are already known ($-1.38$ and $-1.21$ millimhos), the selection of any conductance will establish extreme values of $P^2$ and $K$, thereby providing information about frequency and amplitude stability. A load conductance of 1.0 millimho (1000 ohms) was selected, and the following values of $P^2$ and $K$ were obtained from Figure 5:
The frequency of operation was set at 1 kc to minimize reactive effects of the transistor. The resonant circuit was constructed with a large air-core inductor of approximately 3.6 millihenries, tuned to 1000 cycles, and shunted by a resistance to yield a 1000-ohm impedance at resonance. The effective Q of the completed resonator was determined by bandwidth measurements to be 37. Several performance tests have been run on the oscillator, in which frequency was measured with a Berkeley Model 5570 Frequency Meter and amplitude was measured with a calibrated oscilloscope. Results of a typical test are shown below:

<table>
<thead>
<tr>
<th>Supply Voltage (volts)</th>
<th>$G_0$ (mmho)</th>
<th>$-G_1/G_0$</th>
<th>$P^2$ (Figure 5)</th>
<th>$K$ (Figure 5)</th>
</tr>
</thead>
<tbody>
<tr>
<td>6</td>
<td>-1.38</td>
<td>0.725</td>
<td>0.09</td>
<td>2.80</td>
</tr>
<tr>
<td>8</td>
<td>-1.29</td>
<td>0.775</td>
<td>0.06</td>
<td>2.15</td>
</tr>
<tr>
<td>10</td>
<td>-1.21</td>
<td>0.826</td>
<td>0.04</td>
<td>1.85</td>
</tr>
</tbody>
</table>

For the range of supply voltages chosen, measurements show an amplitude variation of 2 to 1 and a frequency variation of 0.07 cycle or 70 ppm.
The relation of frequency to \( P^2 \) and \( Q \) was given in Quarterly Report No. 4 as:

\[
f = f_0 \left[ 1 - \frac{P^2}{2Q_0^2} \right].
\]

From this relation, the stability corresponding to any pair of values for \( P^2 \) is:

\[
\frac{\Delta f}{f} = \frac{\Delta P^2}{2Q_0^2}.
\]

Using the values for \( P^2 \) obtained from the analysis, predicted frequency stability is:

\[
\frac{\Delta f}{f} = \frac{(0.09 - 0.04)^2}{2 \times 37^2} = 18 \text{ ppm}
\]

To predict amplitude stability, graphical measurements of \( V_0 \) (analogous to \( I_0 \) previously discussed) must be obtained from Figure 13. By approximating a "break point" for each characteristic, the following values are obtained:

\[
V_0 \text{ for 6 volts} = 0.3 \text{ volts}
\]

\[
V_0 \text{ for 10 volts} = 0.9 \text{ volts}.
\]

Predicted amplitudes corresponding to supply voltages of 6 volts and 10 volts are therefore:

\[
\text{Voltage amplitude for 6 volts supply} = KV_0 = 2.80 \times 0.3 = 0.84 \text{ volts}
\]

\[
\text{Voltage amplitude for 10 volts supply} = KV_0 = 1.85 \times 0.9 = 1.67 \text{ volts}
\]
The predicted amplitude variation is therefore 1.99 (compared with a measured variation of 2.0), and the ratio of predicted amplitudes to measured amplitudes is approximately 1.4.

The reason for choosing a voltage divider as a bias source will be apparent from a study of Figure 13. If bias were obtained from a single resistor in series with the tank circuit, such a resistor would have a value of 741 ohms. The slope of the resulting bias characteristic would be almost the same as the slope of the transistor characteristic, thus affording very poor definition of the operating point and correspondingly poor temperature stability. To provide bias characteristics which intersect the transistor characteristics in well-defined points, the bias source must have an internal conductance which is large compared with the effective negative conductances of the transistor. This consideration was the basis for selecting 100 ohms as the value for $R_1$.

The chief disadvantage of such an arrangement is the large power dissipated in the voltage divider, which in this case averaged 76 per cent of the total power consumed.

The value of load conductance described above was selected to provide a particular performance. The original goal was 10 ppm, which would have required $\Delta P^2$ of 0.05 with a Q of 50. From Figure 5, a value of $P^2 = 0.05$ corresponds to a $-R_1/R_o$ ratio of 0.8. Assuming that $\Delta P^2$ would be approximately equal to $P^2$ for the nominal supply voltage (8 volts), this ratio of 0.8 indicated a desired load conductance of 1.03 millimho, which was taken as 1.0 millimho for convenience. The Q of 50 was obtained with an inductance employing a powdered-iron toroidal core, which presented the desired load conductance at resonance. However, nonlinearity of the core caused variations
of inductance with amplitude, resulting in frequency variations much larger than those due to the transistor alone. Within the accuracy of the method used to measure the inductance variation, the core variation accounted for all of the frequency variation. This result led to the construction of an air-core inductance as a means of obtaining linearity.

The results described above represent an accuracy of prediction by analysis which approaches the limits of available information on the components of the resonant circuit. To refine the results at this frequency would require an investigation of the linearity of the capacitance employed, which present objectives do not justify. It can reasonably be concluded that the analysis is effective at audio frequencies.

D. Free-Running Oscillators

The work on the free-running oscillator circuit which was described briefly in Quarterly Report No. 4 was continued during this quarter. The circuit shown in Figure 15 is the same as that which appeared in the earlier report. As noted previously, 25 kc was the frequency chosen for the experimental analysis since this frequency is sufficiently low to assure relative freedom from transistor phase characteristic variations, while at the same time it is sufficiently high to permit a wide selection of Q and resonant impedance for the tuned circuit. This frequency permitted the use of air-core inductors which eliminated core saturations as another possible source of nonlinearity.

For this investigation, several parallel resonant LC circuits were designed and constructed. The following table indicates the characteristics of some of these circuits.
Figure 15. 25-Kc Free-Running Oscillator.

It may be noted from the table that an attempt was made to obtain one set of LC circuits all having the same resonant impedance and frequency but having
a range of \( Q \) values and also another set having approximately equal \( Q \)'s and resonant frequencies but having a range of resonant impedance values.

The experimental results obtained with the circuit of Figure 15 led to the following tentative conclusions:

1. Better stability is obtained with the higher values of \( Q \) although a \( Q \) greater than 15 or 20 does not produce appreciable improvements. An optimum \( Q \) does not exist.

2. Best stability is obtained with low values of resonant impedance (5,000 to 10,000 ohms). It should be noted that the resonant impedance of the tuned circuit must be greater than the negative resistance presented by the base of the transistor, otherwise, parallel-mode oscillations will not be obtained.

3. Higher values of \( R_e \) usually produce the best stability while at the same time reducing the tendency to oscillate. Typical values for \( R_e \) range from 2000 ohms to 8,000 ohms. An optimum value of \( R_e \) does exist in most instances.

4. The choice of \( R_b \) did not greatly affect the stability although some improvement was observed as \( R_b \) was decreased in value. In many cases \( R_b' \) could be made zero, the only direct-current resistance remaining in the circuit being \( R_1 \), the resistance of the inductor.

The best stability that was obtained by observing the above principles was a frequency change of 2.5 parts per million for 1 per cent change in supply voltage. This result could not always be duplicated due to the dependance of stability on temperature. No attempts were made to determine the effect of temperature on stability due to the lack of a means of temperature control.

At this low frequency it was assumed that the effect of transistor capacitance and phase could be neglected. Thus, practically all frequency shift could be attributed to the effects of harmonic distortion. Very little
frequency drift with time was observed as long as the temperature remained constant. However, changing transistors in this circuit produced frequency changes as great as five per cent. Also, the value of $R_e$ required for best stability was not the same for various transistors of the same type.

Several circuit modifications were tried in an attempt to overcome the above disadvantages. Amplitude limiting was tried using type 1N34 silicon germanium diodes. The relatively low back resistance prevented oscillations from occurring. Later, some Texas Instrument Company, type-600, silicon, junction diodes were tried in the circuit of Figure 16.

Figure 16. 25-Kc Oscillator Using Diode Amplitude Limiting.

A total of 16 transistors were tested in this circuit. Five of them failed to oscillate. With eight of the transistors it was possible by the proper selection of $R_e$ to reduce the frequency variation to less than one part per million for one per cent voltage change (tuned circuit number 4 was used). The other three transistors oscillated properly but produced greater frequency
variation. No method was found to obtain best stability for all transistors without changing the value of $R_e$.

Curves showing the effect of $R_e$ on stability are presented in Figure 17. The proper choice of $R_e$ for the particular transistor and tuned circuit would be near 4700 ohms.

An inductor was mounted in a temperature-controlled oven to determine whether the effect due to temperature change was partially caused by changes in the inductance and $Q$ of the tuned circuit. Because of space limitations, the maximum inductance obtainable was 0.6 millihenry. This required the circuit to be operated at 100 kc in order to obtain a satisfactory $Q$ and resonant impedance. The stability was considerably poorer than at 25 kc. This was caused primarily by transistor phase and capacitance characteristics which begin to become important at 100 kc. It was learned, however, that changing the temperature of the inductance has an appreciable effect on the frequency of oscillation. The important factor seemed to be the variation in $Q$ rather than the variation in inductance, since the change in inductance was quite small. Any change in $Q$ will cause a change in the amount of harmonic distortion which will in turn cause a shift in the frequency of oscillation.

At 100 kc, using the oven-mounted inductance, a value of $R_e$ which would give stabilities comparable to that obtained at 25 kc could not be found.

The same inductance was resonated at 275 kc. At this frequency, $R_e$ had more control over stability than at 100 kc. However, the frequency variation-versus-voltage relation (shown in Figure 18) was no longer a straight line. The best stability over a range of 5 volts variation in $E$ was less than 4 parts per million for 1 per cent change in $E$. Random frequency drift with time was
Figure 17. Frequency Variation as a Function of Supply Voltage for Different Values of $R_e$. 
Figure 18. Frequency Variation as a Function of Supply Voltage for Different Values of $R_e$ (Normalized at 25 volts).
still present. This was probably due to temperature variations of other circuit components.

These results indicate that this circuit is of greatest use only as long as the frequency of operation is not greater than about 10 per cent of the alpha cut-off frequency of the transistor. This may seem to be a high penalty to pay for a reasonably stable free-running oscillator, but where circuit simplicity is important, consideration should be given to this circuit. It is expected that with high-frequency, point-contact transistors (such as the 2N33), oscillators using the circuit of Figure 16 can be designed having stabilities of the order of 1 part per million at frequencies as high as 5 megacycles and perhaps higher. The diode limiting would permit the interchangeability of transistors; however, $R_e$ would have to be retained as a variable to compensate for differences of individual transistor characteristics. In addition, the diodes used must have a high back resistance to provide proper limiting at the higher frequencies. The ideal diode characteristic would be such that no current would flow for reverse voltages and no current for forward voltages up to some definite forward voltage at which point the diode should present zero dynamic impedance to the source. This would eliminate the need for batteries to establish the clipping level. Some of the silicon junction diodes presently available approach this characteristic fairly well. For example, in the case of the Texas Type 600, very little forward current flows for forward voltages less than 0.5 volt. For voltages greater than this value, the forward current increases very rapidly. Thus, in the circuit of Figure 16, the clipped amplitude at the base is slightly greater than one volt peak-to-peak. At higher frequencies, however, this diode is not as useful because of the variation in junction capacitance with voltage and current.
The experimental work on these circuits was performed without benefit of the theoretical analysis and design data described in sections B and C of this report, since these were concurrent activities. A careful examination of the experimental results reveals them to be in substantial agreement with the theory, at least in a qualitative way. Thus, the discussion on page 27 leads to the conclusion that decreasing $R_o'$, Figure 15, should improve the performance. In addition, it moves the operating point to a less favorable position, so that the improved performance must be obtained at the expense of reduced amplitude of oscillations. There are some indications that despite the wide range of parameters tested, the optimum values were not found.

E. Junction Transistor Oscillators

Junction transistors oscillate quite readily in most of the standard circuit configurations such as the Hartley and the Colpitts. However, in constructing junction transistor oscillators to operate at frequencies where the angle of $\alpha$ is large and the magnitude of $\alpha$ is much less than unity it was found that the Hartley and Colpitts circuits do not oscillate readily. Consider the possibility of using a network such as shown in Figure 19 to produce a current in $R_1$ such that $I_2\angle\theta = I\angle0$. In order to meet this condition, analysis shows that:

1. The deviation of the frequency of $I$ from the series resonant frequency of the network is required to be large in order to maintain $\theta = 0$ for small values of $\theta$.

2. $R_1$ and $R_2$ should be as small as possible in order to meet the condition $I_2\angle\theta = I\angle0$ in the immediate vicinity of resonance.

3. $L$ should be large.
With these conditions in mind, a junction transistor oscillator was constructed, using the network of Figure 19 in the feedback path as shown in the circuit of Figure 20.

Figure 19. Circuit Showing Conditions on Loop Phase Shift.

Figure 20. LC Oscillator.
Figure 21. Alpha Characteristics of a Typical Type CK 720 Junction Transistor.
In this circuit, $R_1$ and $R_2$ of the previously described network are the emitter and coil resistances respectively. Using a transistor whose alpha characteristics are as shown in Figure 21, it was possible to obtain oscillations at frequencies as high as $2\, mc$, which is about 5 times alpha cutoff, and in a region of large phase shift in alpha. It was found that $L$ must be large and $R_1 + R_2$ must be small to maintain oscillations, as was expected from analysis.

The same type of network was used to construct a crystal-controlled oscillator, as shown in Figure 22.

![Crystal-Controlled Oscillator](image)

Referring to Figure 23, it is noticed that the frequency of oscillation is greatly dependent on the values of emitter and collector supply voltage. However, an increase in collector voltage produces a decrease in frequency, while an increase in emitter voltage causes an increase in the frequency of oscillation. This suggests that if the proper ratio of emitter to collector voltage could be maintained, the frequency of oscillation might be made independent of
Figure 23. Variation of Frequency With Collector Supply Voltage for Various Values of Emitter Supply Voltage.
power-supply fluctuations. This can be accomplished by the use of a voltage divider so that the emitter bias voltage is obtained from the collector supply voltage. However, perfect compensation can be obtained only over that region of collector voltage variation for which the curves of the various emitter voltages are equally spaced. Very good compensation can be obtained over this region. Also, it is possible to obtain better average stability over a wider range of supply-voltage variations by sacrificing some of the stability against voltage variations in the region where the emitter curves are equally spaced. This can be accomplished by computing the values in the voltage divider on the basis of the average ratio of changes in emitter voltage to changes in collector voltage required to maintain a constant frequency.

A circuit embodying this type of compensation is shown in Figure 24.

Figure 24. Compensated Crystal Oscillator With Single Battery Supply.
The variation of frequency with voltage of this circuit is shown in Figure 25. These curves indicate a substantial improvement over the uncompensated circuit.

With the compensated circuit, stability of better than 1 part in $10^7$ was obtained for a 1 per cent change in supply voltage.
Figure 25. Variation of Frequency with Supply Voltage for Two Values of R.
V. CONCLUSIONS

The approximate theory which has been developed for sinusoidal oscillators yields results of acceptable engineering accuracy at low frequencies. The results so far obtained indicate that this theory can be used as the basis for practical oscillator design. The novel features of the theory are the use of design curves derived from idealized transistor characteristics and the direct calculation of frequency stability.

The circuit shown in Figure 15, which is one of the simplest circuits possible using point-contact transistors, does not give optimum performance. Desirable modifications include methods of compensating for internal phase shift at high frequencies and circuits for stabilizing the base bias voltage at the operating point.

Junction transistors can be made to operate at frequencies well above their alpha cut-off frequencies through the use of the current multiplication obtained in an antiresonant circuit.

The performance of both free-running and crystal-controlled oscillators is greatly influenced by the method of obtaining base bias. The best frequency stability is achieved when this bias is obtained from the collector supply voltage through the use of a properly designed voltage divider.
VI. PROGRAM FOR THE NEXT INTERVAL

1. Because of the difficulty of obtaining a suitable test chamber, data on the effect of temperature have been delayed. A test chamber which will permit controlled variation of temperatures over a wide range has been constructed and temperature characteristics of circuits already described will be investigated during this quarter.

2. Circuits of the high-frequency oscillators previously described will be re-examined with a view of improving them in the light of what has been learned from the low-frequency experiments.

3. Work on junction transistors will be continued.
VII. PERSONNEL

The key personnel working on this project during the period covered by this report are:

<table>
<thead>
<tr>
<th>Name</th>
<th>Position</th>
<th>Approximate Hours Worked</th>
</tr>
</thead>
<tbody>
<tr>
<td>B. J. Dasher, Sc. D.</td>
<td>Project Director</td>
<td>130</td>
</tr>
<tr>
<td>D. L. Finn, Ph. D.</td>
<td>Research Associate</td>
<td>60</td>
</tr>
<tr>
<td>S. N. Witt, Jr., M. S.</td>
<td>Research Engineer</td>
<td>260</td>
</tr>
<tr>
<td>W. B. Warren</td>
<td>Research Assistant</td>
<td>260</td>
</tr>
<tr>
<td>T. N. Lowry</td>
<td>Research Assistant</td>
<td>260</td>
</tr>
</tbody>
</table>

Mr. T. N. Lowry joined the project as a research assistant in July 1954. He received his B.E.E. from the Georgia Institute of Technology in June 1952 and is currently continuing in the Graduate Division. Mr. Lowry was a radio relay officer in the Signal Corps for two years and an application engineer (student) with Westinghouse Electric Corporation for fifteen months.

Respectfully submitted:

Approved:

B. J. Dasher
Project Director

D. L. Finn
Research Associate

S. N. Witt, Jr.
Research Engineer

W. B. Warren, Jr.
Research Assistant

J. E. Boyd, Head
Physics Division

Paul K. Calaway, Acting Director
Engineering Experiment Station

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Research Assistant
QUARTERLY REPORT NO. 6
PROJECT NO. 236-206

TRANSISTOR OSCILLATORS OF EXTENDED FREQUENCY RANGE

By

B. J. DASHER, D. L. FINN, S. N. WITT, JR.,
W. B. WARREN, JR., and T. N. LOWRY

- o - o - o - o -

CONTRACT NO. DA-36-039-sc-42712

DEPARTMENT OF THE ARMY PROJECT: 3-99-11-022
SIGNAL CORPS PROJECT: 142B

- o - o - o - o -

DECEMBER 30, 1954
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DECEMBER 30, 1954
TABLE OF CONTENTS

I. PURPOSE .................................................. 1
II. ABSTRACT .................................................. 2
III. CONFERENCES. ............................................ 3
IV. FACTUAL DATA ............................................ 4
   A. Introduction ........................................... 4
   B. Temperature Test Chamber .............................. 5
   C. Temperature Stability Data ............................ 6
   D. Negative-Resistance Oscillators ...................... 10
   E. Junction-Transistor Oscillators ..................... 13
   F. Graphical Determination of Input Characteristics .. 16
V. CONCLUSIONS .............................................. 26
VI. PROGRAM FOR THE NEXT INTERVAL ....................... 27
VII. PERSONNEL ............................................. 28
VIII. BIBLIOGRAPHY .......................................... 29

This Report Contains 29 Pages
## LIST OF FIGURES

<table>
<thead>
<tr>
<th>Figure Description</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>1. 400-Kc Amplitude-Stabilized Oscillator</td>
<td>6</td>
</tr>
<tr>
<td>2. Free-Running Low-Frequency Oscillator</td>
<td>8</td>
</tr>
<tr>
<td>3. 1-Kc Oscillator</td>
<td>9</td>
</tr>
<tr>
<td>4. Compensated Crystal Oscillator With Single Battery Supply</td>
<td>10</td>
</tr>
<tr>
<td>5. Frequency-Versus-Temperature Characteristics for Circuit of Figure 4</td>
<td>11</td>
</tr>
<tr>
<td>6. Push-Pull Negative-Resistance Oscillator</td>
<td>13</td>
</tr>
<tr>
<td>7. Crystal-Controlled Oscillator With Resonant Collector Circuit</td>
<td>14</td>
</tr>
<tr>
<td>8. Simplified Crystal Oscillator</td>
<td>14</td>
</tr>
<tr>
<td>9. Frequency-Versus-Voltage Characteristics for Several Transistors</td>
<td>15</td>
</tr>
<tr>
<td>10. Emitter-Controlled Crystal Oscillator</td>
<td>16</td>
</tr>
<tr>
<td>11. Grounded-Collector Transistor Circuit</td>
<td>17</td>
</tr>
<tr>
<td>12. Characteristic Curves for a Typical Point-Contact Transistor</td>
<td>20</td>
</tr>
<tr>
<td>13. Effect of Collector Voltage on Input Character</td>
<td>21</td>
</tr>
<tr>
<td>14. Idealized Forward Characteristics of a Point-Contact Transistor</td>
<td>23</td>
</tr>
<tr>
<td>15. Idealized Input Volt-Ampere Character for Circuit of Figure 11</td>
<td>25</td>
</tr>
</tbody>
</table>
I. PURPOSE

The purpose of this project is to evolve data and information regarding transistor oscillator circuitry and the degree of reliability and reproducibility attained with transistor parameter variations. It includes both analytical and experimental studies of crystal-controlled and free-running oscillators employing transistors exclusively as active elements.

The work is divided into three principal tasks as follows:

(1) the evaluation of crystal-controlled and free-running transistor oscillators from the point of view of reliability, size, weight, circuit requirements, frequency limitations and other similar considerations;

(2) the development of new circuits and or techniques specifically to take advantage of the unique characteristics of transistors regarding size, weight, power requirements, etc.; and

(3) the development of practical engineering data and information to serve as a basis for designing free-running and crystal-controlled junction- and point-contact transistor oscillators in the range 0.1 to 75 mc.
II. ABSTRACT

A laboratory-constructed temperature test chamber is described, and some data on temperature stability of transistor oscillators, obtained by making use of the test chamber, are presented herein. The best temperature stability obtained is 4 parts in $10^8$ for 1°C temperature change, with an oscillator having a voltage stability of 1 part in $10^8$ for a 1 per cent voltage change. This oscillator makes use of a point-contact transistor. Circuit schematics are presented for junction-transistor oscillators having similar orders of stability.

Data on a negative-resistance oscillator operating both at 1 kc and 32 kc are given. It is concluded that this type oscillator has best stability only when the frequency of operation is low compared to the alpha cut-off frequency of the transistor.

Additional junction-transistor oscillators are described, the circuits of which provide a novel method of using crystal control.

A graphical method of determining the input characteristic of grounded-collector circuits is explained. This method is applied to the point-contact transistor.
III. CONFERENCES

No conferences were held during this quarter.
IV. FACTUAL DATA

A. Introduction

During this quarter, the construction of a temperature test chamber has been completed. Several of the circuits previously described have undergone tests in this chamber.

The 1-kc oscillator discussed in Quarterly Report No. 5 has been modified to operate at 32 kc to determine the effect of transistor phase shift on stability calculations based on distortion.

Several attempts have been made to convert the two-terminal oscillator shown in Figure 15, Quarterly Report No. 5, to crystal control. Two major difficulties arise from efforts to substitute the crystal directly for the parallel resonant LC circuit. First, the crystal has no path for direct current. Second, the parallel resonant impedance of a crystal is extremely high. At first thought, it might appear feasible to connect a large inductance in parallel with the crystal, but, when this is done, oscillations occur at the parallel resonant frequency of the inductance and the shunt capacitance of the crystal. Even if this resonance is made to occur at the crystal frequency, very little control is maintained by the crystal.

Another approach to this problem is to operate the crystal in its series mode and invert the impedance by means of a transmission line. Since a transmission line is an open circuit at low frequencies, it is still necessary to provide a direct-current path. This has been successfully accomplished by the use of diodes paralleling the crystal. One such circuit making use of a quarter-wave lumped-constant transmission line and a 1-mc crystal gave a voltage stability of better than 1 part in 10^7 for a 1 per cent change in supply voltage. Since the data on this circuit is incomplete, it will not be presented in this report.
Some experimental work involving push-pull oscillators has also been conducted during this quarter and the work on junction-transistor oscillators has been continued.

A method is presented for the graphical determination of the input characteristics of a point-contact transistor connected in the grounded-collector configuration.

B. Temperature Test Chamber

The temperature test chamber completed during this quarter has performed very satisfactorily in initial tests over the temperature range from -10 to +50°C.

The chamber is housed in an outer shell made of 1/4-inch plywood. The inner shell is constructed of aluminum and is separated from the outer shell by wooden spacers and two inches of insulating material. The door to the 450-cubic-inch test compartment is attached by 4 wing nuts and is provided with 12 banana jacks for making electrical connections.

Heat for the chamber is provided by two 50-watt heating elements and one 60-watt lamp. Any number or combination of the heaters may be used as required. Cooling is effected by a 450-cubic-inch dry-ice chamber. Air circulation is accomplished by a centrifugal blower with the motor mounted outside of the insulated chamber. A Fenwal Thermoswitch is used to control the temperature by opening or closing the electrical circuit to the heaters.

A dial-type thermometer is provided for reading the temperature. It has been found that the thermometer reading does not always precisely agree with the actual transistor surface temperatures as measured by other means. Where precise results are necessary, thermocouples in contact with the transistors
may be required. Measurement techniques involving this and other points will be improved as experience is gained in the use of the test chamber.

Five pounds of dry ice are sufficient for approximately two hours of cyclic operation of the chamber between -10° and +50° C.

C. Temperature Stability Data

Some of the previously discussed oscillator circuits have been tested for temperature stability. One of the first circuits tested in this way is a modification of Figure 8 of Quarterly Report No. 4. This circuit was chosen because of its good frequency stability with respect to voltage variations. As may be seen in Figure 1, the only changes from the original circuit are in the value of $R_k$, the type of diodes used and the type of transistor used.

![400-KC CRYSTAL](image)

**400-KC CRYSTAL**

**W.E. 1729**

**TEXAS 601**

$10K$ $3.9K$ $0.01\mu f$

$E$

$R_k$ $C_k$

$0.01\mu f$

Figure 1. 400-Kc Amplitude-Stabilized Oscillator.

With $R_k = 2.2K$, typical frequency stability with respect to supply-voltage variation was 5 parts in $10^8$ for 1 per cent voltage change at a frequency of 400 kc. Best stability obtained was 1 part in $10^8$ for 1 per cent voltage change. This occurred only over a limited voltage range.
Typical stability with respect to temperature was found to be $4 \times 10^{-8}$ for $1^\circ \text{C}$ temperature change. Typical change in frequency when transistors were changed was $2 \times 10^{-6}$. The above-mentioned data applies over a temperature range from $+25^\circ \text{C}$ to $-12^\circ \text{C}$ and a voltage range from 20 to 25 volts. With $R_k = 1\, \text{k}$, better results can be obtained with a wider range of transistor parameters without sacrificing stability. Under all conditions, the output voltage across $R_c$ remained essentially constant at 1.3 volts peak-to-peak. The direct current supplied to the oscillator did not vary with temperature. The available output power was about 0.1 milliwatt into a 4000-ohm resistive load. The substitution of a suitable transformer in the place of $R_c$ will permit greater output.

Another circuit from which temperature data has been obtained is shown in Figure 2. This circuit was first presented as Figure 10 of Quarterly Report No. 4. As noted previously, this circuit has very poor frequency-versus-voltage stability. With a W. E. Type 1698 transistor, the temperature stability was $6 \times 10^{-4}$ for $1^\circ \text{C}$ change. With a W. E. Type 1729 transistor, the temperature stability was even worse. However, the circuit was not adjusted for optimum performance with the 1729. For a description of the resonant circuit of Figure 2, see LC Circuit No. 4, page 29, Quarterly Report No. 5.
Figure 2. Free-Running Low-Frequency Oscillator.

The 1-kc oscillator of Figure 3, which was described in Quarterly Report No. 5, has been tested for temperature stability in the temperature test chamber. With a W. E. Type 1768 transistor, the oscillator maintained oscillations over only a 10°C temperature range for any single adjustment of the bias network. By readjustment of the bias network, the 10-degree range could be established anywhere between -10°C and +50°C, but no single adjustment could be found which would provide oscillations over more than the 10-degree range. Frequency stability of the oscillator within each 10-degree range averaged about 2 parts in $10^5$ per degree centigrade, for constant supply voltage. As the bias network was readjusted for continuously changing temperatures, oscillation was restored each time at a frequency within 1 part in $10^4$ of the preceding frequency. The critical behavior of this circuit for variations in temperature is largely due to the fact that the amplitude of oscillation is determined by the shape of the transistor characteristic. The selection of a load resistance nearly equal to the negative resistance was made
to facilitate analysis of the oscillator, but this condition causes relatively poor definition of amplitude.

![Figure 3. 1-Kc Oscillator.](image)

Temperature runs have also been made on the junction-transistor crystal-controlled oscillator first shown as Figure 24 of Quarterly Report No. 5. This circuit is repeated here as Figure 4. Typical results are shown by the curve of Figure 5, which indicates frequency as a function of temperature with a constant supply voltage. This curve indicates a stability of $1.4 \times 10^{-7}$ for $1^\circ C$ temperature change over the range from $-10^\circ$ to $+40^\circ$ C. Similar values of stability were obtained over this temperature range for other values of supply voltage.

All of the circuits discussed in this section were operated with only the transistor in the temperature test chamber in order to avoid the necessity of investigating the temperature characteristics of other components. The possibility of compensating for the transistor changes with temperature-sensitive passive elements is also recognized. This principle normally will be applied in the final circuit designs.
D. Negative-Resistance Oscillators

The oscillator described in Part IV-C of Quarterly Report No. 5 was designed and constructed to test the accuracy of the analysis set forth in that report. During this quarter, the same circuit (see Figure 3) was operated at higher frequencies in order to examine the effect of transistor reactance on predicted frequency stability. A parallel resonant circuit was constructed for 32 kc which exhibited the same resonant impedance (1000 ohms) and Q (40) as the 1-kc circuit originally employed. From the analysis based on the assumption of a resistive negative impedance, the oscillator performance expected would be the same at 32 kc as it was at 1 kc.

Performance of the circuit at 32 kc did not compare favorably with that at 1 kc when the original W. E. Type 1768 transistor was employed. Frequency
Figure 5. Frequency - Versus - Temperature Characteristics for Circuit of Figure 4.
variation for 1 per cent voltage change increased from 1.5 parts in $10^6$ at 1 kc
to 45 parts in $10^6$ at 32 kc, and oscillation could be maintained only over a
voltage range of ±10 per cent of the nominal supply voltage. Substitution of
W. E. Type 1698 and W. E. Type 1729 transistors in the oscillator of Figure 3
with the 32-kc resonant circuit improved the frequency stability by a factor of
10 and permitted voltage variations of ±25 per cent of nominal supply voltage.
With no alteration of the oscillator other than adjustment of the bias network,
stabilities between 4 and 5 parts in $10^6$ for 1 per cent voltage change were
obtained. These results are largely what would be expected from a consider-
ation of the alpha cut-off frequencies of the transistor types employed, as
indicated below:

<table>
<thead>
<tr>
<th>Transistor Type</th>
<th>Alpha Cut-off Frequency (mc)</th>
<th>Frequency Variation†</th>
</tr>
</thead>
<tbody>
<tr>
<td>W. E. 1768</td>
<td>0.35</td>
<td>45</td>
</tr>
<tr>
<td>W. E. 1698</td>
<td>2.0</td>
<td>4.5</td>
</tr>
<tr>
<td>W. E. 1729</td>
<td>3.7</td>
<td>4.2</td>
</tr>
</tbody>
</table>

†Parts in $10^6$ for one per cent supply voltage change at 32 kc.

Oscillographic studies of the volt-ampere characteristic at the transistor-
input terminals provide a means of determining the conditions which produce multi-
valued relations between voltage and current. At 32 kc, for example, the input
characteristic of a W. E. Type 1768 transistor produces an unmistakably open
figure, whose apparent reactance varies considerably with operating point and
amplitude of applied signal. Under the same conditions, the input character-
istics of W. E. Type 1698 and W. E. Type 1729 transistors are single valued and
produce single lines rather than open figures.
Another negative-resistance circuit briefly investigated during this quarter is shown in Figure 6. Using the 1-kc resonant circuit previously employed, this configuration oscillates with wide variations in supply voltage and individual transistors. No oscillation occurs when either transistor is removed. The resonant circuit must provide a direct-current path to maintain oscillation, so the oscillator must be free-running or must employ an impedance inverter such as that described in Part IV-A.

![Figure 6. Push-Pull Negative-Resistance Oscillator.](image)

E. Junction-Transistor Oscillators

It has been found that the circuit of Figure 4 oscillates quite well when the average base current is zero. Although this operating point offers no particular advantage from the standpoint of stability, it does offer the possibility of considerable circuit simplification. The circuit of Figure 7 makes use of this operating point in obtaining single-battery operation without the use of a bleeder. Over the range of supply voltages from 30 to 70 volts, it exhibits a stability of about 5 parts in $10^8$ for a one per cent voltage change. Oscillation was obtained over the range from 30 to 80 volts with a good sinusoidal waveform at the collector.
Figure 7. Crystal-Controlled Oscillator With Resonant Collector Circuit.

The circuit of Figure 8 represents an additional simplification but does not have as good stability as the circuit of Figure 7. The frequency of oscillation varies considerably as various transistors are placed in the circuit. This fact is demonstrated by the curves of Figure 9 which show the variation of frequency with supply voltage for five different CK-720 transistors.

Figure 8. Simplified Crystal Oscillator.
Figure 9. Frequency - Versus - Voltage Characteristics for Several Transistors.
Another configuration is obtained if the crystal is placed in the emitter circuit as shown in Figure 10. The circuits of Figures 8 and 10 are essentially the same as that of Figure 4 with the ground point moved to the collector instead of to the base and with the operating point at zero base current.

![Emitter-Controlled Crystal Oscillator](image)

**Figure 10. Emitter-Controlled Crystal Oscillator.**

**F. Graphical Determination of Input Characteristics**

In section IV of Quarterly Progress Report No. 5, especially prepared curves were used in the design of a two-terminal, parallel mode oscillator. It was necessary to consider the actual shape of the volt-ampere characteristics of the transistor circuits in order to make use of these design curves. These characteristics depend not only on the transistors themselves but also on the external circuits other than the tank circuit.

There are at least three ways that information about the shape of the circuit volt-ampere characteristic may be obtained: by actual measurement, by graphical analysis of static transistor characteristic curves or by linearization of the transistor characteristic curves. In this section a
simple method of graphical analysis and a method of linearization for determination of the input volt-ampere characteristic of a grounded-collector point-contact transistor circuit will be presented.

The grounded-collector transistor circuit under consideration is shown in Figure 11.

![Grounded-Collector Transistor Circuit](image)

Figure 11. Grounded-Collector Transistor Circuit.

Equations describing the operation of this circuit are:

\[ V_c - V_e = I_e R_e - E_{cc} \]  
\[ V_l = V_c - E_{cc} \]  
\[ I_b = -I_e - I_c \]

The desired input characteristic of this circuit shows the variation of \( I_b \) with \( V_l \). The functional relationship between these two variables can be obtained by solution of equations 1, 2 and 3.

One obvious method of solving these equations graphically is as follows: The input characteristic (\( V_e \) versus \( I_e \) for constant \( I_c \)) and the forward characteristic (\( V_c \) versus \( I_e \) for constant values of \( I_c \)) can be used to plot the variation of \( V_c - V_e \) with \( I_e \) for constant values of \( I_c \). Load lines can then be

-17-
constructed on this new set of curves for different values of $R_e$ and $E_{cc}$ by the use of equation 1.

Any point on one of these load lines corresponds to a particular value of $I_e$ and $I_c$ for the circuit of Figure 11. When these two currents are determined, the value of $I_b$ for this particular operating condition is specified by equation 3.

The voltage $V_c$ can be found by referring the particular value of $I_e$ selected on the load line back to the forward characteristic. The voltage $V_1$ may then be obtained by use of equation 2. Calculation of the points for the entire input volt-ampere characteristic of the circuit may be completed by continued repetition of this procedure.

This graphical method is time consuming because the forward characteristic and the input characteristic of the transistor must be combined to provide a new set of characteristic curves. Fortunately, a study of typical point-contact transistor characteristics shows that $V_e$ is small compared with $V_c$ for most operating conditions. Furthermore, a study of the behavior of the input volt-ampere characteristics for various load lines shows that the negative-resistance portion of the characteristic does not ordinarily occur in the region where $V_e$ is comparable in magnitude to $V_c$. This means that the voltage $V_e$ can ordinarily be neglected in equation 1, and, consequently, the load lines can be drawn on the forward characteristic according to the approximation

$$V_c = I_e R_e - E_{cc}. \hspace{1cm} (4)$$

Any point on the load line drawn on the forward characteristic by use of equation 4 specifies a particular value of $I_e$ and $I_c$ for the circuit of Figure
11. The current $I_b$ for this operating condition may then be found by use of equation 3. The voltage $V_1$ may be determined by combining equations 4 and 2.

$$V_1 = -V_c - E_{cc} = -I_e R_e \quad (5)$$

The value of collector current $I_c$ for $V_1 = -I_e R_e = 0$ is most easily obtained by referring the value of $E_{cc}$ used to the output characteristic of the transistor. In fact, the number of points calculated for the input characteristic can be increased considerably by use of the output characteristic as well as the forward characteristic.

The forward characteristic and output characteristic curves for a typical point-contact transistor are shown in Figure 12. A load line drawn by the use of equation 4 for $R_e = 2000$ ohms and $E_{cc} = 14$ volts is shown on the forward characteristic. The transistor characteristics shown in Figure 12 were used to calculate the input volt-ampere characteristics of the grounded-collector transistor circuit as shown in Figure 13.

Inspection of the load line shown in Figure 12 shows that for large values of $V_1 = -I_e R_e$ the emitter current $I_e$ changes slowly while the collector current $I_c$ changes rapidly. The collector current $I_c$ first increases in magnitude and then decreases as the emitter current $I_e$ decreases. This causes a peak in the input characteristic of the circuit. The rapidly changing collector current has the greatest effect on the formation of this peak. Consequently, the peak occurs near the point where the load line crosses the "knee" of the forward characteristic curves. Actually, the peak occurs for a somewhat smaller value of $I_e$ than the value at the knee. This small shift is caused by the effect of the variation of $I_e$ on the base current $I_b$. 
Figure 12. Characteristic Curves for a Typical Point-Contact Transistor.
Figure 13. Effect of Collector Voltage on Input Characteristic (Computed by Graphical Method).
A reasonably accurate input characteristic for the grounded-collector circuit can usually be calculated very rapidly by the method just described provided the transistor characteristic curves are available. In case the transistor curves are not available, a crude idea of the shape of the input characteristic may be obtained by consideration of small-signal parameters for the transistor. The method given here for accomplishing this requires the linearization of the forward characteristic of the transistor and is subject to the approximations inherent in the graphical method discussed above as well as the rather serious errors inherent in the linearization process.

The idealized forward characteristic of the transistor is shown in Figure 14. The idealized characteristic is divided into three regions:

Region 1: \( I_e < 0 \).

Region 2: \( 0 < I_e < -I_c/\alpha \).

Region 3: \( I_e > -I_c/\alpha \).

The slopes of these straight lines are approximated by the small-signal parameter \( r_{21} \) measured in the approximate region.

The intercept of the lines for \( I_e = 0 \) may be obtained from the emitter-diode characteristic which is assumed to be described by the relation

\[ V_c = I_c r_{22} \]  \hspace{1cm} (7)

The small-signal parameter \( r_{22} \) ideally is measured for \( I_e = 0 \). An idea of the extent of the approximation involved in using equation 7 may be obtained by inspecting the \( I_e = 0 \) curve on the input characteristic shown in Figure 12.
Figure 14. Idealized Forward Characteristics of a Point-Contact Transistor.
The equations of the three straight lines used in the idealized forward characteristic are:

\[ V_c = I_e (r_{21})_1 + I_c r_{22}. \]  
(8)  

Region 2:

\[ V_c = I_e (r_{21})_2 + I_c r_{22}. \]  
(9)  

Region 3:

\[ V_c = I_c (r_{21})_3 + I_c \left[ r_{22} + \frac{(r_{21})_3}{\alpha} - \frac{(r_{21})_2}{\alpha} \right]. \]  
(10)  

The equation of the knee of the curves is

\[ V_c = I_e \left[ \alpha r_{22} - (r_{21})_2 \right]. \]  
(11)  

Thus, the knee of the curve occurs at a value of \( V_c = 0 \) when

\[ \alpha r_{22} - (r_{21})_2 = 0. \]  
(12)  

Equations 8, 9 and 10 may be combined with equations 2, 3 and 4 to determine the idealized input volt-ampere characteristic of the grounded-collector point-contact transistor circuit shown in Figure 11. This idealized input volt-ampere characteristic is shown in Figure 15.

It should be emphasized, again, that the assumptions used in obtaining this idealized characteristic are so severe that they can not ordinarily be used to obtain accurate quantitative information.
Figure 15. Idealized Input Volt-Ampere Characteristics for Circuit of Figure 11.
V. CONCLUSIONS

The temperature-stability data that have been obtained thus far indicate that when the frequency-versus-voltage stability is good the frequency-versus-temperature stability is also good. This might be expected from the theory since both temperature and voltage have similar effects on the transistor characteristics. In the future, temperature runs will be made only on the circuits which have good frequency-versus-voltage stabilities since less time is required to make voltage variation tests.

Most point-contact oscillator circuits investigated so far indicate that good frequency stability can be obtained only if the frequency of oscillation is low compared to the alpha-cut-off frequency of the transistor. This limitation was observed when both the 1-kc and the 25-kc oscillators were operated at higher frequencies. Means are now being sought to overcome this difficulty by artificially compensating for the phase-shift characteristic of the transistors.

About the same order of voltage and temperature stabilities have been obtained for the point-contact and junction-transistor oscillators. Likewise, about the same maximum frequency of operation for reasonable stability has been obtained. Thus, nothing at present indicates that the major emphasis should be placed on either the point-contact or the junction circuits. However, the greater production uniformity of the junction transistor would make it the preferable type where other factors are equal.
VI. PROGRAM FOR THE NEXT INTERVAL

1. Temperature tests on those circuits which show good frequency-versus-voltage stability will continue.

2. Both junction and point-contact oscillator circuit investigations will continue. Oscillators will be designed for higher frequencies making use of some of the newer transistors which have been recently obtained. More emphasis will be placed on junction-transistor oscillator circuits than in the past.
VII. PERSONNEL

The key personnel working on this project during the period covered by this report are:

<table>
<thead>
<tr>
<th>Name</th>
<th>Position</th>
<th>Approximate Hours Worked</th>
</tr>
</thead>
<tbody>
<tr>
<td>B. J. Dasher, Sc. D.</td>
<td>Project Director</td>
<td>130</td>
</tr>
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</tr>
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VIII. BIBLIOGRAPHY


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QUARTERLY REPORT NO. 7
PROJECT NO. 236-206

TRANSISTOR OSCILLATORS OF
EXTENDED FREQUENCY RANGE

By
B. J. DASHER, S. N. WITT, JR.
and W. B. WARREN, JR.

CONTRACT NO. DA-36-039-sc-42712

DEPARTMENT OF THE ARMY PROJECT: 3-99-11-022
SIGNAL CORPS PROJECT: 142B

MARCH 31, 1955
QUARTERLY REPORT NO. 7

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- o - o - o - o -

MARCH 31, 1955
# TABLE OF CONTENTS

<table>
<thead>
<tr>
<th>Section</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>I. PURPOSE</td>
<td>1</td>
</tr>
<tr>
<td>II. ABSTRACT</td>
<td>2</td>
</tr>
<tr>
<td>III. CONFERENCES</td>
<td>3</td>
</tr>
<tr>
<td>IV. FACTUAL DATA</td>
<td>4</td>
</tr>
<tr>
<td>A. Introduction</td>
<td>4</td>
</tr>
<tr>
<td>B. Definitions of Stability</td>
<td>5</td>
</tr>
<tr>
<td>C. Impedance Inversion Oscillator</td>
<td>7</td>
</tr>
<tr>
<td>D. Bridge Oscillator Circuits</td>
<td>8</td>
</tr>
<tr>
<td>E. General Conditions for Oscillator Stability</td>
<td>12</td>
</tr>
<tr>
<td>F. Stabilization by Distortion</td>
<td>25</td>
</tr>
<tr>
<td>G. Measurement of Stability</td>
<td>30</td>
</tr>
<tr>
<td>V. CONCLUSIONS</td>
<td>35</td>
</tr>
<tr>
<td>VI. PROGRAM FOR THE NEXT INTERVAL</td>
<td>37</td>
</tr>
<tr>
<td>VII. PERSONNEL</td>
<td>38</td>
</tr>
<tr>
<td>VIII. BIBLIOGRAPHY</td>
<td>39</td>
</tr>
</tbody>
</table>

This Report Contains 39 Pages
<table>
<thead>
<tr>
<th>Figure</th>
<th>Description</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>Impedance Inversion Oscillator</td>
<td>7</td>
</tr>
<tr>
<td>2</td>
<td>1000-kc Meacham Bridge Transistor Oscillator</td>
<td>9</td>
</tr>
<tr>
<td>3</td>
<td>General Feedback Oscillator</td>
<td>13</td>
</tr>
<tr>
<td>4</td>
<td>Modification of the Thevenin Generator</td>
<td>15</td>
</tr>
<tr>
<td>5</td>
<td>Simple Frequency-Controlling Network</td>
<td>17</td>
</tr>
<tr>
<td>6</td>
<td>Design Curves for Oscillator of Figure 5</td>
<td>18</td>
</tr>
<tr>
<td>7</td>
<td>Standard 1000-kc Oscillator</td>
<td>21</td>
</tr>
<tr>
<td>8</td>
<td>1000-kc Oscillator Designed for Impedance Match</td>
<td>21</td>
</tr>
<tr>
<td>9</td>
<td>1000-kc Oscillator Designed From Curves of Figure 6</td>
<td>22</td>
</tr>
<tr>
<td>10</td>
<td>1000-kc Negative-Feedback Oscillator</td>
<td>24</td>
</tr>
<tr>
<td>11</td>
<td>Zero-Base-Current Oscillator With External Distortion</td>
<td>26</td>
</tr>
<tr>
<td>12</td>
<td>Performance Curves for Circuit of Figure 11</td>
<td>27</td>
</tr>
<tr>
<td>13</td>
<td>Zero-Base-Current Oscillator With Controllable Distortion</td>
<td>28</td>
</tr>
<tr>
<td>14</td>
<td>Performance Curves for Circuit of Figure 13</td>
<td>29</td>
</tr>
<tr>
<td>15</td>
<td>Hartley Oscillator With Distortion Compensation</td>
<td>30</td>
</tr>
<tr>
<td>16</td>
<td>Performance Curves for Circuit of Figure 15</td>
<td>31</td>
</tr>
<tr>
<td>17</td>
<td>Compensated Hartley Oscillator With Zero Base Current</td>
<td>32</td>
</tr>
<tr>
<td>18</td>
<td>Performance Curves for Circuit of Figure 17</td>
<td>33</td>
</tr>
</tbody>
</table>
I. PURPOSE

The purpose of this project is to evolve data and information regarding transistor oscillator circuitry and the degree of reliability and reproducibility attained with transistor parameter variations. It includes both analytical and experimental studies of crystal-controlled and free-running oscillators employing transistors exclusively as active elements.

The work is divided into three principal tasks as follows:

(1) the evaluation of crystal-controlled and free-running transistor oscillators from the point of view of reliability, size, weight, circuit requirements, frequency limitations and other similar considerations;

(2) the development of new circuits and/or techniques specifically to take advantage of the unique characteristics of transistors regarding size, weight, power requirements, etc.; and

(3) the development of practical engineering data and information to serve as a basis for designing free-running and crystal-controlled junction- and point-contact transistor oscillators in the range 0.1 to 75 mc.
II. ABSTRACT

In order to facilitate comparisons among oscillator circuits, the term stability has been defined as a **numerical quantity** which **increases** as a circuit is made more stable. Stability may be expressed as a function of any variable parameter, such as voltage or temperature.

The use of impedance inversion to obtain crystal control of point-contact oscillators has been found to be impractical in most cases.

A transistorized Meacham bridge oscillator has been developed which has performance comparable to a similar oscillator using vacuum tubes. The best stability obtained was approximately 2 parts in $10^9$ frequency variation for one per cent voltage change.

A design procedure is described which provides a method for obtaining best stability for most feedback-type oscillators.

Data are given on a novel method of stabilization of junction-transistor oscillators using distortion control.

The problems involved in measuring frequency stability are discussed.
III. CONFERENCES

Dr. B. J. Dasher and Mr. S. N. Witt, Jr., visited the Signal Corps Engineering Laboratories, Fort Monmouth, New Jersey, January 13, 1955. The objectives of the project and the present status of the work were discussed with Dr. G. K. Guttwein, Lt. M. B. Herscher and Mr. Eelstrom. A suggestion was made by SCEL personnel that greater emphasis be placed on junction-transistor research.

Dr. B. J. Dasher attended the AIEE-IRE Conference on Transistor Circuits at the University of Pennsylvania in Philadelphia, February 17 and 18, 1955.
A. Introduction

During this quarter, the emphasis has been shifted from point-contact to junction transistors for the single transistor type of oscillator. The use of impedance inversion to obtain crystal control for two-terminal oscillators which require a parallel resonant circuit was further investigated. It was concluded that such oscillators become too complicated to be practical.

The application of transistors to the Meacham bridge type of oscillator has been investigated. It was found that most of the junction transistors presently available are not suitable for this application at a frequency of 1000 kc. However, many of these transistors should perform properly in this circuit at lower frequencies. A successful Meacham bridge oscillator was constructed using two Western Electric Type 1729 transistors. The performance of this oscillator was comparable to that of a vacuum-tube oscillator of the same type which consumed 130 times as much power. Because of the high stability of the transistor oscillator, a major problem in measuring stability resulted. This required accurate comparison of frequency-measuring equipment with radio station WWV.

An oscillator design procedure using the Philco Surface Barrier Transistor Type L-5100 was developed and illustrated. The resulting stabilities compared favorably with the Meacham bridge oscillators. Again, the stabilities quoted for these oscillators are in some cases limited by the frequency-measuring technique rather than by the oscillators.

Methods of stabilizing the Hartley-type junction-transistor oscillator have been investigated. It was found that such oscillators could often be stabilized by producing distortion external to the transistor. This was
accomplished by using diode limiting. Almost any degree of voltage stability can be obtained by this method; however, the short-time or drift stability is not necessarily improved by the same amount.

Some experimental junction oscillators have been operated at frequencies between 1 and 15 mc. One crystal-controlled 12-mc oscillator was constructed in a total volume of less than 0.8 cubic inch, including the transistor, crystal and chassis. The oscillator operated on a single dry-cell battery. However, the stability was only about 5 parts in $10^7$ for one per cent voltage change. This relatively poor stability was perhaps caused by the characteristics of the crystals used. The work on junction-transistor oscillators at higher frequencies has not progressed sufficiently for the details to be included in this report.

Investigations of accurate methods of measuring stability have been conducted.

B. Definitions of Stability

The following definitions will be used throughout this report to express the stability of an oscillator.

\[
S_v = \frac{(\Delta v)(f)}{(\Delta f)(v)10^4}
\]

(1)

where: $\Delta v$ is the total change in the supply voltage in volts,

$\Delta f$ is the frequency change in cycles per second caused by $\Delta v$,

$v$ is the nominal operating voltage in volts and

$f$ is the nominal frequency of oscillation in cycles per second.

Voltage stability as here defined is the reciprocal of ppm for one per cent voltage change as used in previous reports. $S_v$ increases as the stability is
improved. Typical values for this quantity will range from 0.1 for average free-
running oscillators to 1000 for exceptionally good crystal oscillators. The
unit of $S_v$ is percentage of voltage change/ppm frequency change.

$$S_t = \text{temperature stability} = \frac{(\Delta t) (f)}{(\Delta f) 10^6}$$

(2)

where: $\Delta t$ is the change in temperature in degrees centigrate,

$\Delta f$ is the frequency change in cycles per second caused by $\Delta t$ and

$f$ is the nominal frequency of oscillation in cycles per second.

$S_t$ is the reciprocal of ppm for one-degree-centigrade temperature change. The
unit of $S_t$ is degrees/ppm frequency change.

$$S_s = \frac{(\Delta s) (f)}{(\Delta f) (s) 10^4}$$

(3)

where: $\Delta s$ is the change in a specific parameter other than temperature or

voltage (examples: resistance, capacitance, inductance, etc.),

$\Delta f$ is the frequency change in cycles per second caused by $\Delta s$,

$s$ is the nominal value of the specific parameter expressed in the same

unit of measure as $\Delta s$ and

$f$ is the nominal frequency of oscillation in cycles per second.

The name applied to this unit would be determined by the parameter which is varied.

$$S = \text{short-time stability} = \frac{f}{(\Delta f) 10^6}$$

(4)
where: \( \Delta f \) is the maximum change in frequency in cycles per second over a period of four hours under specified conditions of voltage, temperature and other variables and

\( f \) is the nominal frequency of oscillation in cycles per second.

The unit of \( S \) is megacycles/cycle. Voltage stability, temperature stability and short-time stability have been arbitrarily adjusted so as to have convenient magnitudes for most crystal-controlled and free-running oscillators.

C. Impedance Inversion Oscillator

As mentioned in Quarterly Report No. 6, several oscillators have been constructed in which crystal control was employed in a grounded-collector circuit. One such circuit which operated satisfactorily is shown in Figure 1. The dotted enclosure represents a lumped-constant transmission line with an electrical length of one-quarter wavelength at 1000 kc. The characteristic impedance is 700 ohms, which is the value required to transform the 100-ohm resistance of the crystal to the desired 5000-ohm resistance needed at the

---

**Figure 1. Impedance Inversion Oscillator.**
transistor base. The quarter wavelength was chosen to transform the series resonance of the crystal to a parallel resonance. The capacitor, C, was made variable so that the electrical length of the line could be adjusted precisely to one-quarter wavelength.

Since a transmission line has no closed d-c path, the diodes were added in parallel with the crystal to provide one. The diodes also provide some control of amplitude and distortion and, thus, tend to improve the stability. The diodes do not materially reduce the Q of the crystal since the crystal is operating at series resonance.

The best voltage stability obtained for this circuit was $S_v = 20$. The use of a low-loss distributed-constant transmission line should improve the voltage stability because fewer spurious resonances would result. However, such a line would be more difficult to construct and would not be conveniently variable in length.

The primary objectives in designing this circuit were (1) to take advantage of the simplicity of the grounded-collector oscillator and (2) to permit the use of the design data that have been previously developed for this type of circuit. The first objective was not realized because of the complexity of the artificial line. The second objective has not been realized because the low-frequency analysis does not include all the factors that are important at 1000 kc. Therefore, further investigation of this circuit has not been conducted.

D. Bridge Oscillator Circuits

The Meacham bridge oscillator shown in Figure 2 was constructed and tested. The amplifier portion uses two W. E. type 1729 point-contact transistors. Based on the assumption that the input impedance of the amplifier is 100 ohms
Figure 2. 1000-kc Meacham Bridge Transistor Oscillator.
or greater, the over-all power gain is greater than 34 db with a 100-ohm-load resistor across the output terminals. The negative-feedback factor of the bridge is 0.5, which gives an over-all phase-reduction factor of about 25.

The amplifier has an undesirable phase frequency characteristic in that the phase shift becomes several radians before the gain is reduced to unity. If the bridge output is connected directly to the amplifier input, oscillations occur that are not controlled by the crystal. If an additional crystal, X2, is placed in series with the feedback path, oscillations can occur only near the series resonance of this crystal. In this region, positive feedback occurs only at the series resonant frequency of X1. Thus, X2 acts as a bandpass filter while X1 controls the frequency of the oscillator.

This undesirable phase-shift characteristic is partly caused by the alpha characteristic of the point-contact transistor since, at high frequencies, the phase angle of alpha can exceed 180 degrees. The use of junction transistors should help to reduce this phase shift.

A vacuum-tube Meacham bridge oscillator was constructed for comparison with the transistor oscillator. The characteristics of the two circuits are presented in the table on the following page.

The operating conditions for the two oscillators were made as nearly the same as possible. The same crystal and oven were used. Room-temperature variations were approximately the same while the oscillators were under test.

The volume of space required for the vacuum-tube oscillator was many times that required for the transistor oscillator. The power required for the vacuum-tube oscillator was 130 times that required for the transistor oscillator. The voltage stability for the two oscillators was approximately the same. The short-
time stability for the vacuum-tube oscillator was slightly better than for the transistor oscillator.

<table>
<thead>
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<th>Item to be Compared</th>
<th>Transistor Oscillator</th>
<th>Vacuum-Tube Oscillator</th>
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</thead>
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<tr>
<td>Number and type of amplifying element used</td>
<td>two W.E.1729</td>
<td>one 12BY7</td>
</tr>
<tr>
<td></td>
<td>one 6S4</td>
<td></td>
</tr>
<tr>
<td>Type of frequency control</td>
<td>Crystal in JKO3 oven</td>
<td>Crystal in JKO3 oven</td>
</tr>
<tr>
<td>Frequency</td>
<td>1000 kc</td>
<td>1000 kc</td>
</tr>
<tr>
<td>Output</td>
<td>3 V P. to P. to high impedance</td>
<td>3 V P. to P. to high impedance</td>
</tr>
<tr>
<td>Power supply</td>
<td>nine 1.5-v batteries</td>
<td>200-v, d-c regulated supply</td>
</tr>
<tr>
<td></td>
<td>6.3-v, a-c transformer</td>
<td></td>
</tr>
<tr>
<td>Power requirements</td>
<td>0.12 watt exclusive of oven</td>
<td>16 watts exclusive of oven</td>
</tr>
<tr>
<td>Permissible voltage variation</td>
<td>± 25%</td>
<td>± 25%</td>
</tr>
<tr>
<td>Frequency-versus-voltage stability</td>
<td>S_v = 500</td>
<td>S_v = 500</td>
</tr>
<tr>
<td>Short-time stability</td>
<td>S = 2.5</td>
<td>S = 3.3</td>
</tr>
</tbody>
</table>

Several junction-transistor amplifiers have been designed for tentative use in this circuit. The transistors which have been tried are the RD-2525A, the CK 720, the RCA 2N34 and the GE ZJ1. The maximum gain obtained at 1000 kc for a two-transistor amplifier was 26 db. The gain for this amplifier, which used the RD-2525A transistors, was 8 db lower than the gain of the amplifier shown in Figure 2. The gain also varied greatly with the supply voltage. The greatest power output was obtained from a circuit designed for use with the GE ZJ1. However, the gain for two GE ZJ1 transistors was only 12 db. The circuits designed for use with the CK 720 and the RCA 2N34 gave less gain than the RD-2525A circuit and less power output than the GE ZJ1 circuit. Thus, none of the junction amplifier circuits using the transistors mentioned above had characteristics desirable for use in the Meacham bridge 1000-kc oscillator.
The Philco L-5100, surface-barrier transistor, was found to have a very high gain as a 1000-kc amplifier but gave very low power output. An amplifier is presently being designed to make use of the high gain of the L-5100 and the higher power output of other transistor types.

For the Meacham bridge circuit, the power output required from the amplifier may range from a minimum of 4 mw to a maximum of about 40 mw, depending on the type of lamp chosen for use as the amplitude regulator.

E. General Conditions for Oscillator Stability

For a low-distortion oscillator, two primary factors affect the ultimate stability. The first is the phase shift in the amplifier portion of the oscillator, particularly the change in phase shift when external parameters such as voltage and temperature are changed. The second factor is the change in frequency required to cause an equal and opposite change in phase shift in the frequency-controlling feedback loop. The second factor must be considered because the total loop phase shift must be zero.

These two factors are easily separated in oscillators which have only external feedback loops. This separation cannot be made as easily for the negative-resistance oscillators. In some circuits, such as the Meacham bridge oscillator, which use both positive and negative feedback, the negative feedback can be considered as either a part of the amplifier or as a part of the frequency-controlling feedback loop. If the negative feedback is considered to be part of the amplifier, it serves to stabilize the phase shift of the amplifier. If it is considered to be a part of the frequency-controlling loop, it serves to reduce the required phase shift in the frequency-controlling element for a given phase shift in the amplifier. Either analysis leads to the same result.
The block diagram for a feedback oscillator is shown in Figure 3. The amplifier may be converted to an equivalent Thevenin voltage generator with a voltage, $K_e L \theta$, in series with a resistance, $r_o$, where $r_o$ is to be considered as a part of the feedback network. The input impedance of the amplifier is considered to be infinite since any reactive component may be canceled at any one frequency and voltage, and the resistive component may be included as a part of $R$. The phase angle, $\theta$, is assumed to be a function of all external parameters such as temperature and voltage. It will be assumed, however, that $\theta$ has been made zero for the normal static condition of the external parameters. $K$ will be assumed to be a real positive constant so that no distortion of the waveform can occur. Since the frequency-controlling network is passive, the oscillator output will be a pure sinusoidal waveform. In an actual oscillator some distortion must occur in order that amplitude limiting will result. However, by carefully selecting $k$, this distortion can be kept small so that the above-mentioned assumptions will be approximately correct.

The primary factors affecting stability now become $\phi(f)$ and $\theta(s)$ where $s$ is a general external parameter affecting the amplifier phase shift. The object is to make $d\theta/ds$ as small as possible and $d\phi/df$ as large as possible.
Several methods are available for reducing the changes in transistor amplifier phase shift for changes in voltage and temperature. One method is to operate the transistor well below its alpha cut-off frequency. Internal positive and/or negative feedback may also be considered as a possible means of improving the amplifier characteristic. However, for this analysis, the use of over-all negative feedback will be considered as a part of the frequency-controlling feedback loop rather than a part of the amplifier. Clipping or other amplitude control means may in some cases be used to improve the amplifier phase-shift characteristics, provided the over-all distortion is kept low in order that the analysis which is to follow may apply in its simplest form.

The factors influencing $\phi(f)$ are the geometry and element values of the frequency-controlling network. The choice of network geometry is limited only by practical values of $K$ and $r_o$. For most transistors, $r_o$ is much too large for use with practical networks. Therefore, the use of a transformer is usually necessary. If the values of $K$ and $r_o$ for the transistor alone are compared with the values for the transistor with a transformer, the quantity $K^2/r_o$ will be found to remain constant. This is shown in Figure 4 where $K'$ and $r'_o$ are the characteristics of the transistor without the transformer. Either $K$ or $r_o$ may be selected for best performance with any particular feedback network, however, $K^2/r_o$ must be held constant. For a practical transformer-coupled amplifier, it is necessary that any transformer loss and phase shift be included in the characteristics, $K/\theta$ and $r_o$. This may be accomplished by measuring the amplifier characteristics with a representative transformer in use. The amplifier and transformer combination may then be completely
characterized in terms of $\theta$ and $K^2/r_o = A = \text{four times the available power output for one-volt input.}$

![TURNS RATIO 1:1]

$$K^2 + (\omega K)^2 = (\omega K)^2 \frac{r_o}{a^2 r_o'} = A$$

Figure 4. Modification of the Thevenin Generator.

To illustrate the method by which a practical oscillator may be designed, the network shown in Figure 5 is analyzed as follows:

Let $jX = jX_1 - jX_2$. Then,

$$e_1 = e_2 \frac{R}{R + R_1 + jX + r_o} \tag{5}$$

or

$$\frac{e_1}{e_2} = k \frac{\tan^{-1} \left( \frac{-X}{R + r_o + R_1} \right)}{k} \tag{6}$$

Near series resonance, $d\phi/df$ increases as $(R + r_o)$ decreases. The minimum value of $(R + r_o)$ that will permit oscillations may be found as follows:

For $x = 0$,

$$e_1 = \frac{Re_2}{r_o + R_1 + R} = \frac{Re_{\perp}}{r_o + R_1 + R}$$

-15-
but

\[ \frac{K^2}{r_o} = A \text{ or } K = r_o^{1/2}A^{1/2}. \]

Therefore,

\[ Rr_o^{1/2}A^{1/2} = r_o + R_1 + R \]

or

\[ R = \frac{r_o + R_1}{r_o^{1/2}A^{1/2} - 1} \quad (7) \]

from which

\[ R + r_o = \frac{r_o + R_1}{r_o^{1/2}A^{1/2} - 1} + r_o. \]

\[ (R + r_o) \text{ is minimum when } \frac{\text{d}(R + r_o)}{\text{d}r_o} = 0 \]

or when

\[ 2r_o^{3/2}A^{1/2} - 3r_o - R_1 = 0. \quad (8) \]

By substituting \( R_1 \) as determined from the resonant element chosen for the oscillator and \( A \) as determined by the amplifier, \( r_o \) can be found from equation 8. Figure 6 shows the variation of \( r_o \) with \( A \) as a parameter with \( R_1 \) constant at 100 ohms. For each value of \( r_o \) and \( A \), \( K \) is determined from \( K^2/r_o = A \). \( R \) is found from equation 7. The variations of \( K \) and \( R \) for \( R_1 = 100 \) ohms are also shown in Figure 6.
If $r_0$ and $R$ could both be made zero and still maintain oscillations, the change in frequency for a small amplifier phase shift, $\Delta \theta$, would be:

$$\Delta f = \frac{(\Delta \theta) (f_0)}{2Q}, \text{ for } \Delta \theta \text{ small}$$ \hspace{1cm} (9)

where: $Q = \frac{X_1}{R_1}$ and $f_0 = \frac{1}{\sqrt{LC}}$ for the resonant element. For values of $r_0$ and $R$ other than zero, equation 9 must be corrected by making $Q = \frac{X_1}{(R_1 + R + r_0)}$.

The relative change in frequency for a practical circuit compared to the change for an unloaded resonant circuit for the same $\Delta \theta$ would then be:

$$\rho_\phi = \frac{R_1}{R + r_0 + R_1}.$$ \hspace{1cm} (10)

Thus, $\rho_\phi$ is a figure of merit for the network. If the values of $\rho_\phi$ are compared for two different networks, then the relative stabilities of oscillators using the two networks can be determined since:

$$\frac{(\rho_\phi)_1}{(\rho_\phi)_2} = \frac{(\rho_\phi)_2}{(\rho_\phi)_2}.$$ \hspace{1cm} (11)

for the same amplifier and resonant element.
Figure 6. Design Curves for Oscillator of Figure 5.
Equation 10 expresses $\rho_\phi$ only for the network of Figure 5. For other networks, an appropriate expression for $\rho_\phi$ would have to be found.

The $\rho_\phi$ of the circuit of Figure 5 is plotted in Figure 6 ($R_1 = 100$ ohms).

To design a practical circuit to use the network of Figure 5, a suitable amplifier is first chosen. $A$ is found experimentally. A suitable resonant element is chosen (crystal, series LC circuit, or other). $R_1$ is determined from the resonant element. The correct value of $r_o$ is then found from equation 8. $R$ can then be found from equation 7. $X$ as a function of frequency can be calculated for the resonant element. Then, $\phi$ as a function of frequency can be found from equation 6. If the amplifier phase shift, $\theta$, is known as a function of voltage (or temperature, etc.), then the frequency of oscillation as a function of voltage can be found by making $\phi$ equal to $\theta$ for all voltages.

This method yields a theoretical stability slightly better than can be obtained in a practical circuit since distortion has been neglected and since $R$ cannot be made as small as the value calculated in equation 7. If $R$ is made exactly equal to the minimum value, when a phase shift, $\theta$, occurs, $R$ will no longer be large enough to permit sustained oscillations.

For some transistors, making $R$ smaller than the value calculated in equation 7 will still produce oscillations but at a reduced amplitude. This is because the nonlinear characteristic of the transistor is often such that the gain, $K'$, is much greater for low amplitudes. From one standpoint, this is undesirable since it implies that the transistor transfer characteristic will produce appreciable distortion. However, it does provide better amplitude limiting.

Closer examination of equation 6 indicates that the amount of improvement to be expected from the proper selection of $R$ may be small for a practical
oscillator using this network configuration. This is particularly true when
the input impedance of the amplifier is low since the maximum value of R that
can physically exist is the actual input resistance of the amplifier. Thus,
the improvement in stability obtained by externally unloading the crystal by
R would be found by first substituting the amplifier input resistance into
equation 6 and calculating the stability and then substituting R (minimum) in-
to equation 6 and again calculating the stability. The ratio of the two stabi-
lities will be the improvement due to the proper selection of R.

More important than the correctness of the magnitude of the stability as
given by this method is the comparison that can be made between networks of
different configurations. This method is also very valuable as a tool in
choosing the constants for any particular feedback network. For example, in
the circuit of Figure 5, one might assume that the value of R was relatively
unimportant. However, as shown previously and substantiated later by exper-
imental results, this assumption is incorrect.

Three circuits have been designed to illustrate the above-mentioned design
procedure. The circuit of Figure 7 was designed without any criterion for ob-
taining the best stability. The circuit of Figure 8 was designed by assuming
that best stability would be obtained when the load on the transistor matches
its internal impedance. The circuit of Figure 9 was designed by making use of
the curves of Figure 6. The following discussion illustrates the step-by-step
design procedure.

Design procedure for Figure 7.

(1) The Philco L-5100 transistor was selected because of its high gain
at 1000 kc.
(2) The grounded emitter configuration was selected because of its high current gain. The amplifier was designed empirically, keeping in mind the dissipation limitations of the transistor. The input impedance was kept high to make use of maximum power gain. The input impedance was found to be 1.6 K by measurement. The output impedance, \( r_o \), for the amplifier and transformer was about 125 ohms. From equation 10, \( \rho_\phi \) was found to be 0.055.

Figure 7. Standard 1000-kc Oscillator.

Figure 8. 1000-kc Oscillator Designed for Impedance Match.
(3) The only requirements placed on the transformer were that it provide the necessary phase inversion and the necessary current gain to obtain oscillations.

(4) A crystal was connected as shown to operate in its series mode.

(5) The stability, $S_v$, for changes in collector voltage, was found experimentally to be 10.

Design procedure for Figure 8.

(1) The same amplifier and transistor were used as in Figure 7 so as to give a fair comparison between the circuits. $K'$ and $r_o'$ for the amplifier were found to be 167 and 15,000 ohms respectively. The turns ratio of the transformer was selected so that the reflected primary impedance would be 15,000 ohms while at the same time maintaining oscillations with a minimum value of $R$. The actual measured value of $r_o$ was 125 ohms. This differed from the theoretical value of 150 ohms because of the transformer losses. From equation 10, $\rho_\phi$ was found to be 0.145.
(2) The crystal was connected as shown with $R$ equal to 17 ohms.

(3) The stability, $S_v$, for changes in collector voltage was found experimentally to be 80.

Design procedure for Figure 9.

(1) The same amplifier and transistor were used as in the previous examples. $K^2/r_o$ was found to be 1.86.

(2) From Figure 6, $r_o$ was found to be 14 ohms and $R$ to be 28 ohms. The required turns ratio of the transformer, $K'/K$, was 33. The value of $p_o$ was 0.71.

(3) The constants actually used in the circuit were as shown in Figure 9.

(4) The voltage stability was found experimentally to be greater than 125. The stability could not be measured more accurately for reasons which will be mentioned later.

From equation 11, the voltage stability for the oscillator of Figure 9 was predicted to be 12.8 times that of Figure 7 and 1.7 times that of Figure 8. The improvement factors actually measured were 12.5 and 1.6 respectively. Thus, the mathematical predictions were substantiated by experimental results.

Figure 10 shows another junction-transistor-oscillator circuit, which was designed by a method similar to that outlined above. The design procedure for this network is not sufficiently complete to be included in this report. However, for the amplifier used in the previous examples, point calculations were made to permit an approximation to the values which should give best stability. The measured voltage stability was greater than 250. For this circuit, $p_o$ was greater than unity. This was possible because of the negative-feedback arrangement.

The circuit of Figure 10 was tested using 11 other type L-5100 transistors. The stability was about the same for each transistor. With no circuit or
voltage changes, the following table shows the actual frequency of oscillation for each of the 12 transistors.

<table>
<thead>
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<th>Transistor No.</th>
<th>Frequency</th>
<th>Transistor No.</th>
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</tr>
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<td>87</td>
<td>999903.8</td>
<td>99</td>
<td>999903.7</td>
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The maximum frequency difference was 0.7 cycle per second. The frequency of oscillation could have been adjusted to be the same in each case by varying the capacitance across the primary of the transformer.

In the above examples, no attempt was made to reduce dθ/ds for the amplifier. Such a reduction would result in improved stabilities for all four circuits. It should also be emphasized that the networks chosen for the above illustrations do not necessarily represent the best ones that can be found. Work is presently being conducted on improving both amplifier designs and network designs for use with this method of analysis.
It should also be observed that the maximum frequency change caused by voltage changes for the circuits of Figures 9 and 10 is much less than the change in resonant frequency of a typical crystal for normal variations in room temperature. Thus, in order to make effective use of the stability of these circuits, it is necessary either to use crystal ovens or to obtain better crystals than the HC-6 type, which are presently being used.

F. Stabilization by Distortion

Recent work with junction transistors has been concerned with a means of frequency stabilizing these oscillators by the introduction of a controllable amount of distortion into the oscillator circuit. This method of compensation is based on the fact that most junction oscillators have a frequency-voltage characteristic such that the frequency of oscillation increases with an increase in collector voltage, while the effect of distortion of the waveform decreases the frequency of oscillation. Consequently, if the natural tendency of the frequency of oscillation to increase with increasing collector voltage can be balanced by a downward shift of the frequency due to increased distortion of the waveform, then the frequency of oscillation will be maintained constant as the collector voltage is varied. Several oscillator circuits have been designed to employ this method of frequency stabilization, with resulting improvements in stability by factors as large as 75. One circuit is shown in Figure 11.

The curves of Figure 12 illustrate the order of improvement in stability obtained by the introduction of the diodes in the emitter circuit. The curve labeled "Without Diodes" is a typical characteristic of the oscillator with several different transistors. The other curves demonstrate the improvement obtained by using the diodes for several different transistors. Although the
frequency of oscillation shifts when a different transistor is placed in the circuit, the compensation gives the same order of improvement in stability. In each case, a small readjustment of L or C in the oscillator will return the frequency of oscillation to its original value, and the curves will then essentially fall on top of one another. This circuit has the disadvantage that the amplitude of oscillation is greatly reduced by the insertion of the diodes in the emitter circuit.

If a diode in series with a resistor is connected in the circuit as shown in Figure 13, the amplitude of oscillation is increased instead of decreased and, in addition, some control of the amount of distortion introduced is obtained. If the value of R is zero, the frequency of oscillation decreases as the supply voltage is increased; if R is infinite, the frequency of oscillation has a rising characteristic with voltage; if R equals 1000 ohms, the frequency of oscillation is almost constant with variations in the supply voltage. This behavior is demonstrated by the curves of Figure 14.
Figure 12. Performance Curves for Circuit of Figure 11.
Figure 13. Zero-Base-Current Oscillator With Controllable Distortion.

The introduction of distortion in an oscillator of the Hartley type is shown in Figure 15. In this circuit, single-battery operation is obtained by the use of a bleeder. Here again, the frequency of oscillation may be made to increase or decrease as the supply voltage is increased by varying the value of R. The frequency of oscillation as a function of supply voltage with R as a parameter is shown in the curves of Figure 16.

The bleeder current of the oscillator of Figure 15 can be avoided if the oscillator is operated with zero base current as shown in Figure 17. The frequency of oscillation of this circuit as a function of supply voltage with R as a parameter is illustrated by the curves of Figure 18.

All the circuits described make use of capacitive feedback, and, consequently, the amount of feedback decreases as the frequency of oscillation is decreased. This reduces the amount of distortion of the waveform, which reduces the amount of frequency pulling due to the compensating diodes. Therefore, some readjustment of the amount of distortion that is intentionally introduced must be made in order that the frequency stability against voltage fluctuations be
Figure 14. Performance Curves for Circuit of Figure 13.
maintained. If the frequency of oscillation is too low, more distortion may be required for proper compensation than can be introduced by the diodes, and the frequency stability is reduced. This disadvantage may be overcome by using a larger capacitor in the feedback path to increase the amount of feedback at the lower frequencies.

![Hartley Oscillator With Distortion Compensation](image)

Figure 15. Hartley Oscillator With Distortion Compensation.

G. Measurement of Stability

The development of more stable oscillators has led to various difficulties in the measurement of stability. The usual method of measuring stability has been to couple the unknown frequency into a Berkeley Model 5570 Frequency Meter and obtain a direct frequency reading. This method depended for its accuracy on the 1000-kc oscillator built into the frequency meter. However, it has been found that the frequency of this 1000-kc oscillator may vary by as much as one part per million over a period of a few minutes. This in turn introduces errors of the same magnitude in measurements. Since the drift of the
Figure 16. Performance Curves for Circuit of Figure 15.
1000-kc oscillator is fairly slow and constant, it has been possible to increase
the accuracy of the measurements greatly by comparing the tenth harmonic of the
oscillator with WWV at 10 mc. The procedure has been to observe continually
the beat frequency with WWV while making a measurement. If the beat frequency
can be kept to less than one cycle per second, the resulting frequency reading
will be correct to within one part in ten million. In order to make use of
this accuracy when measuring the stability of transistor oscillators operating
near 1000 kc, it is necessary to count the unknown frequency for a period of
at least 20 seconds. Since the Berkeley Frequency Meter is not equipped to do
this, a laboratory counter was constructed to permit counting for either 20
or 200 seconds. Thus, with a one per cent voltage change, the best voltage
stability, $S_y$, that can be assigned to an oscillator under test is 10. However,
for a 25 per cent voltage change, the limitation on $S_y$ is 250 providing the re-
sulting frequency change can be assumed to be a linear function of voltage.

Figure 17. Compensated Hartley Oscillator With Zero Base Current.

The stabilities quoted for the Meacham oscillator and the oscillator of
Figure 10 were limited by the frequency-measuring technique. The stability
Figure 18. Performance Curves for Circuit of Figure 17.
quoted for the circuit of Figure 9 was also limited by measurement technique but in this instance was limited because of poor reception of WWV. It is therefore possible that these oscillators may be even more stable than claimed. It is interesting to observe that all three of these oscillators have much better short-time and voltage stabilities than the standard oscillator in the Berkeley Frequency Meter. It is planned in the near future to replace the oscillator in the frequency meter by either a transistor or a vacuum-tube Meacham oscillator.
V. CONCLUSIONS

The stabilities obtained with both free-running and crystal-controlled transistor oscillators have been consistently improved during this quarter. Satisfactory oscillators using junction transistors up to frequencies of 1000 kc have been constructed. However, for higher frequencies most junction transistors, with the exception of the L-5100, have not been found suitable. Present indications are that the L-5100 will provide good stabilities up to about 10 mc and will operate with lesser stability to 60 mc.

Good stability has been obtained when using the W. E. 1729 transistor in the Meacham circuit. However, the circuit has not yet been adapted to control by an LC circuit rather than a crystal. Since a tuned amplifier is used, the circuit will perform properly only at the amplifier center frequency. Thus, to change frequency the entire amplifier must be retuned. Work is being conducted to overcome this disadvantage by using broad-tuned amplifiers. The only transistor which promises to be suitable for this purpose is the L-5100.

Great improvements in the stability of junction-transistor oscillators have been made possible by the use of external distortion provided by crystal diodes. This method is applicable to both free-running and crystal-controlled oscillators. This method of stabilization has not been applied at frequencies above 1000 kc.

The development of more stable transistor oscillators has made necessary the investigation of more accurate methods of measuring frequency. It is presently possible to measure absolute frequency to better than one part in $10^7$ with the equipment available. This means that when a stability value is given, the voltage change must be great enough to produce at least one part in $10^7$. 
change in absolute frequency in order for the results to be limited by the oscillator under test rather than by the measuring technique. To give the stability in terms of one per cent voltage change, it is necessary to assume that the frequency is a linear function of voltage. To substantiate this assumption, a less stable resonant element can be used and the actual frequency-voltage curve plotted for steps of one per cent voltage change. This has been the general procedure followed in arriving at the stability values quoted in section IV.
VI. PROGRAM FOR THE NEXT INTERVAL

1. Junction-oscillator-circuit investigations will continue. These investigations will be broadened to include such transistors as the silicon, surface barrier and tetrode transistors.

2. Work on stabilization by distortion will continue and will be extended to higher frequencies.

3. The methods of stabilizing transistor amplifiers will be investigated for use in connection with the procedure outlined in section IV-E. Analytic studies of networks for use with this procedure will continue. Design curves for other oscillator configurations will be constructed.

4. Methods of measuring frequency will be improved. A stabilized transistor oscillator is presently being developed for use as a frequency standard for the Berkeley Frequency Meter.
VII. PERSONNEL

The key personnel working on this project during the period covered by this report are:

<table>
<thead>
<tr>
<th>Name</th>
<th>Position</th>
<th>Approximate Hours Worked</th>
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<tbody>
<tr>
<td>B. J. Dasher, Sc.D.</td>
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</tr>
<tr>
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</tr>
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<td>W. B. Warren, Jr., B.E.E.</td>
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</tr>
<tr>
<td>T. N. Lowry, B.E.E.</td>
<td>Research Assistant</td>
<td>260</td>
</tr>
</tbody>
</table>

Respectfully submitted:

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W. B. Warren, Jr.  
Research Assistant

Approved:

J. E. Boyd  
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Engineering Experiment Station
VIII. BIBLIOGRAPHY


FINAL REPORT
PROJECT NO. 236-206

TRANSISTOR OSCILLATORS OF
EXTENDED FREQUENCY RANGE

By
B. J. DASHER and S. N. WITT, JR.

CONTRACT NO. DA-36-039-sc-42712
DEPARTMENT OF THE ARMY PROJECT: 3-99-11-022
SIGNAL CORPS PROJECT: 142B

JUNE 30, 1955
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- o - o - o - o -

JUNE 30, 1955
# TABLE OF CONTENTS

I. INTRODUCTION ................................................. 1
   A. History of the Contract .......................... 1
   B. Survey of the Problem ......................... 1

II. OSCILLATOR THEORIES AND DESIGN METHODS ................. 5
   A. General Requirements and Definitions ............ 5
   B. The Linear Circuit Approach ..................... 7
   C. The Isocline Diagram ............................ 26
   D. The Groszkowski Analysis of a Nonlinear Oscillator 31
   E. A General Design Procedure for Linear Oscillators 64
   F. The Empirical Approach .......................... 76
   G. Vacuum-Tube Analogies ............................ 77
   H. Relative Advantages of the Design Procedures .... 79
   1. Negative-Resistance Oscillators .................. 79
   2. Feedback Oscillators .............................. 80

III. PROPERTIES OF RESONATORS .................................. 81
   A. Quartz Crystals, Series and Parallel Modes ....... 81
   B. Lumped Resonant Elements .......................... 85
   C. The Importance of Q and Resonant Impedance .......... 86
   D. Impedance Inversion ............................... 87

IV. TYPICAL TRANSISTOR OSCILLATOR CIRCUITS ................. 91
   A. Negative-Resistance Oscillators .................. 91
   B. Free-Running Feedback Oscillators ............... 92
   C. Crystal-Controlled Feedback Oscillators .......... 94

V. SELECTION OF TRANSISTORS ................................... 97

VI. OSCILLATOR STABILIZATION ................................ 103
   A. Bias Compensation ................................ 103
   B. Component Selection and Application .............. 105
   C. Temperature Compensation ......................... 109
   D. Stabilization by Distortion ...................... 110
   E. Bridge Stabilization .............................. 113
   F. Transistor and Crystal Replaceability .......... 116

(Continued)
TABLE OF CONTENTS (Concluded)

<table>
<thead>
<tr>
<th>VII. RECOMMENDED DESIGNS</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>A. Properties of Oscillators Investigated</td>
<td>119</td>
</tr>
<tr>
<td>B. Selection and Recommendations of Circuits</td>
<td>119</td>
</tr>
<tr>
<td>C. Practical Oscillator Circuits</td>
<td>123</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>VIII. HIGH-FREQUENCY OPERATION OF TRANSISTORS</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>185</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>IX. MISCELLANEOUS DATA</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>A. Transformer Design</td>
<td>187</td>
</tr>
<tr>
<td>B. Frequency Measurements</td>
<td>189</td>
</tr>
<tr>
<td>C. Parameter Measurements and Calculations</td>
<td>191</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>X. CONCLUSIONS AND RECOMMENDATIONS</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>A. Summary of Results Achieved</td>
<td>213</td>
</tr>
<tr>
<td>B. Remaining Problems</td>
<td>214</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>XI. PERSONNEL</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>217</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>XII. BIBLIOGRAPHY</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>221</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>XIII. APPENDIX</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>A. Derivation of Groszkowski’s Equation</td>
<td>225</td>
</tr>
<tr>
<td>B. Evaluation of Groszkowski’s Equation as Applied to Transistor Oscillators</td>
<td>229</td>
</tr>
<tr>
<td>C. Oscillator Design Problem</td>
<td>235</td>
</tr>
<tr>
<td>D. High-Frequency Oscillator Design Problem</td>
<td>246</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>XIV. INDEX</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>247</td>
</tr>
</tbody>
</table>
LIST OF FIGURES

<table>
<thead>
<tr>
<th>Figure</th>
<th>Description</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>2.1</td>
<td>Basic Two-Terminal Oscillator</td>
<td>8</td>
</tr>
<tr>
<td>2.2</td>
<td>Equivalent Circuit of Negative-Resistance Oscillator</td>
<td>13</td>
</tr>
<tr>
<td>2.3</td>
<td>Locus of $R_{2r} + jR_{2i}$ for Variable Frequency</td>
<td>16</td>
</tr>
<tr>
<td>2.4</td>
<td>Equivalent Circuit of Grounded-Base Transistor Oscillator</td>
<td>17</td>
</tr>
<tr>
<td>2.5</td>
<td>Equivalent Circuit of Tuned-Collector Oscillator</td>
<td>21</td>
</tr>
<tr>
<td>2.6</td>
<td>Tuned-Emitter Oscillator Circuit</td>
<td>23</td>
</tr>
<tr>
<td>2.7</td>
<td>Tuned-Collector Oscillator Circuit</td>
<td>24</td>
</tr>
<tr>
<td>2.8</td>
<td>Performance of the Circuit in Figure 2.6</td>
<td>24</td>
</tr>
<tr>
<td>2.9</td>
<td>Performance of the Circuit in Figure 2.7</td>
<td>25</td>
</tr>
<tr>
<td>2.10</td>
<td>Transistor Oscillator Circuit</td>
<td>27</td>
</tr>
<tr>
<td>2.11</td>
<td>Idealized Volt-Ampere Characteristic</td>
<td>28</td>
</tr>
<tr>
<td>2.12</td>
<td>Isocline Diagram for a Small L/C Ratio for the Circuit in</td>
<td>30</td>
</tr>
<tr>
<td></td>
<td>Figure 2.10</td>
<td></td>
</tr>
<tr>
<td>2.13</td>
<td>Isocline Diagram for a Large L/C Ratio for the Circuit in</td>
<td>30</td>
</tr>
<tr>
<td></td>
<td>Figure 2.10</td>
<td></td>
</tr>
<tr>
<td>2.14</td>
<td>Amplitude and Frequency as Functions of Collector Voltage</td>
<td>32</td>
</tr>
<tr>
<td>2.15</td>
<td>The Effects of Changes in Q and Changes in Inductance on</td>
<td>32</td>
</tr>
<tr>
<td></td>
<td>Frequency Variations</td>
<td></td>
</tr>
<tr>
<td>2.16</td>
<td>Typical Negative-Resistance Characteristic</td>
<td>34</td>
</tr>
<tr>
<td>2.17</td>
<td>Circuit for Obtaining Negative Resistance</td>
<td>34</td>
</tr>
<tr>
<td>2.18</td>
<td>Circuit for Studying Input Impedance</td>
<td>35</td>
</tr>
<tr>
<td>2.19</td>
<td>Variation of $R_1$ as a Function of Collector Supply Voltage</td>
<td>36</td>
</tr>
<tr>
<td>2.20</td>
<td>Idealized Emitter Characteristic</td>
<td>43</td>
</tr>
<tr>
<td>2.21</td>
<td>(a) Volt-Ampere Characteristics</td>
<td>45</td>
</tr>
<tr>
<td></td>
<td>(b) Voltage Waveforms</td>
<td></td>
</tr>
</tbody>
</table>

(Continued)
## LIST OF FIGURES (Continued)

<table>
<thead>
<tr>
<th>Figure</th>
<th>Description</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>2.22</td>
<td>Variations of a Type I Characteristic</td>
<td>47</td>
</tr>
<tr>
<td>2.23</td>
<td>Normalized Performance Curves for Type I and Type II Oscillators</td>
<td>49</td>
</tr>
<tr>
<td>2.24</td>
<td>(a) Volt-Ampere Characteristics.</td>
<td>52</td>
</tr>
<tr>
<td></td>
<td>(b) Voltage Waveform. (M = 0)</td>
<td></td>
</tr>
<tr>
<td>2.25</td>
<td>Normalized Performance Curves for a Type I Oscillator</td>
<td>53</td>
</tr>
<tr>
<td>2.26</td>
<td>Normalized Performance Curves for a Type II Oscillator</td>
<td>54</td>
</tr>
<tr>
<td>2.27</td>
<td>Normalized Performance Curves for a Type I Oscillator</td>
<td>56</td>
</tr>
<tr>
<td>2.28</td>
<td>Normalized Performance Curves for a Type II Oscillator</td>
<td>57</td>
</tr>
<tr>
<td>2.29</td>
<td>Normalized Performance Curves for a Type I Oscillator (Expanded)</td>
<td>59</td>
</tr>
<tr>
<td>2.30</td>
<td>Normalized Performance Curves for a Type II Oscillator (Expanded)</td>
<td>60</td>
</tr>
<tr>
<td>2.31</td>
<td>Effect of Collector Voltage on Input Characteristic</td>
<td>61</td>
</tr>
<tr>
<td>2.32</td>
<td>Effect of Emitter Resistance on Input Characteristic</td>
<td>62</td>
</tr>
<tr>
<td>2.33</td>
<td>Effect of Collector Resistance on Input Characteristic</td>
<td>63</td>
</tr>
<tr>
<td>2.34</td>
<td>General Feedback Oscillator</td>
<td>66</td>
</tr>
<tr>
<td>2.35</td>
<td>Modification of the Thevenin Generator</td>
<td>68</td>
</tr>
<tr>
<td>2.36</td>
<td>Simple Frequency-Controlling Network</td>
<td>70</td>
</tr>
<tr>
<td>2.37</td>
<td>Design Curves for Oscillator of Figure 2.36</td>
<td>71</td>
</tr>
<tr>
<td>2.38</td>
<td>Negative-Feedback Network</td>
<td>74</td>
</tr>
<tr>
<td>2.39</td>
<td>Design Curves for Oscillator of Figure 2.38</td>
<td>75</td>
</tr>
<tr>
<td>3.1</td>
<td>Equivalent Circuit of a Quartz Crystal</td>
<td>81</td>
</tr>
<tr>
<td>3.2</td>
<td>A Quarter-Wave Transmission Line as an Impedance Inverter</td>
<td>89</td>
</tr>
<tr>
<td>3.3</td>
<td>Pi-Network Impedance Inverter</td>
<td>89</td>
</tr>
<tr>
<td>4.1</td>
<td>Basic Negative-Resistance Oscillator Configurations</td>
<td>91</td>
</tr>
<tr>
<td>4.2</td>
<td>Typical Free-Running Feedback Oscillator Circuits</td>
<td>93</td>
</tr>
</tbody>
</table>

(Continued)
# LIST OF FIGURES (Continued)

<table>
<thead>
<tr>
<th>Figure</th>
<th>Description</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>4.3</td>
<td>Additional Crystal-Controlled Circuits</td>
<td>95</td>
</tr>
<tr>
<td>6.1</td>
<td>Frequency Versus Voltage and Current for the Example</td>
<td>104</td>
</tr>
<tr>
<td>6.2</td>
<td>Circuit for the Bias Compensation Example</td>
<td>105</td>
</tr>
<tr>
<td>6.3</td>
<td>Diode Stabilized Feedback Oscillator</td>
<td>110</td>
</tr>
<tr>
<td>6.4</td>
<td>Typical Diode Characteristics</td>
<td>111</td>
</tr>
<tr>
<td>6.5</td>
<td>Another Method of Stabilization by Distortion</td>
<td>112</td>
</tr>
<tr>
<td>6.6</td>
<td>Effect of Compensating Diode on Free-Running Hartley Oscillator</td>
<td>114</td>
</tr>
<tr>
<td>6.7</td>
<td>Bridge Circuit for Improving the Frequency-Voltage Characteristic of Oscillators</td>
<td>116</td>
</tr>
<tr>
<td>7.1</td>
<td>Collector-Controlled Free-Running Oscillator</td>
<td>126</td>
</tr>
<tr>
<td>7.2</td>
<td>Crystal-Controlled Oscillator for High-Impedance Crystal</td>
<td>127</td>
</tr>
<tr>
<td>7.3</td>
<td>Tuned-Emitter Oscillator</td>
<td>128</td>
</tr>
<tr>
<td>7.4</td>
<td>Crystal-Controlled Tuned-Emitter Oscillator</td>
<td>129</td>
</tr>
<tr>
<td>7.5</td>
<td>Tuned-Base Oscillator</td>
<td>130</td>
</tr>
<tr>
<td>7.6</td>
<td>25-kc Oscillator Using Diode Amplitude Limiting</td>
<td>131</td>
</tr>
<tr>
<td>7.7</td>
<td>Frequency Variation as a Function of Supply Voltage for Different Values of $R_e$</td>
<td>133</td>
</tr>
<tr>
<td>7.8</td>
<td>Frequency Variation as a Function of Supply Voltage for Different Values of $R_e$</td>
<td>134</td>
</tr>
<tr>
<td>7.9</td>
<td>Impedance-Inversion Oscillator</td>
<td>135</td>
</tr>
<tr>
<td>7.10</td>
<td>11-mc Free-Running Oscillator</td>
<td>136</td>
</tr>
<tr>
<td>7.11</td>
<td>95-mc Oscillator</td>
<td>137</td>
</tr>
<tr>
<td>7.12</td>
<td>Crystal-Controlled Oscillator Using Transformer Feedback</td>
<td>138</td>
</tr>
<tr>
<td>7.13</td>
<td>Crystal-Controlled Oscillator With Transformer Feedback and Bias Compensation</td>
<td>139</td>
</tr>
</tbody>
</table>

(Continued)
<table>
<thead>
<tr>
<th>Figure</th>
<th>Description</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>7.14</td>
<td>Crystal-Controlled Oscillator With Tuned-Base Circuit.</td>
<td>140</td>
</tr>
<tr>
<td>7.15</td>
<td>Circuit of Figure 7.14 Modified for Single-Battery Operation.</td>
<td>141</td>
</tr>
<tr>
<td>7.16</td>
<td>Transformer Coupled 400-kc Oscillator.</td>
<td>142</td>
</tr>
<tr>
<td>7.17</td>
<td>400-kc Oscillator.</td>
<td>143</td>
</tr>
<tr>
<td>7.18</td>
<td>Change in Frequency for a Given Change in Voltage as a Function of Crystal Resistance</td>
<td>144</td>
</tr>
<tr>
<td>7.19</td>
<td>400-kc Amplitude Stabilized Oscillator.</td>
<td>145</td>
</tr>
<tr>
<td>7.20</td>
<td>High-Frequency Junction-Transistor Oscillator.</td>
<td>146</td>
</tr>
<tr>
<td>7.21</td>
<td>Hartley Oscillator With Distortion Compensation.</td>
<td>147</td>
</tr>
<tr>
<td>7.22</td>
<td>Compensated Hartley Oscillator With Zero Base Current.</td>
<td>148</td>
</tr>
<tr>
<td>7.23</td>
<td>Performance Curves for Circuit of Figure 7.22.</td>
<td>149</td>
</tr>
<tr>
<td>7.24</td>
<td>Typical 1000-kc Feedback Oscillator.</td>
<td>150</td>
</tr>
<tr>
<td>7.25</td>
<td>1000-kc Oscillator Designed for Impedance Match.</td>
<td>152</td>
</tr>
<tr>
<td>7.26</td>
<td>1000-kc Oscillator Designed from Curves of Figure 2.37.</td>
<td>154</td>
</tr>
<tr>
<td>7.27</td>
<td>1000-kc Negative-Feedback Oscillator.</td>
<td>156</td>
</tr>
<tr>
<td>7.28</td>
<td>Crystal-Controlled Oscillator With Resonant Collector Circuit.</td>
<td>158</td>
</tr>
<tr>
<td>7.29</td>
<td>Simplified Crystal Oscillator.</td>
<td>159</td>
</tr>
<tr>
<td>7.30</td>
<td>Frequency-Versus-Voltage Characteristics for Several CK 720 Transistors.</td>
<td>160</td>
</tr>
<tr>
<td>7.31</td>
<td>Emitter-Controlled Crystal Oscillator.</td>
<td>161</td>
</tr>
<tr>
<td>7.32</td>
<td>Zero-Base-Current Oscillator With External Distortion.</td>
<td>162</td>
</tr>
<tr>
<td>7.33</td>
<td>Performance Curves for Circuit of Figure 7.32.</td>
<td>163</td>
</tr>
<tr>
<td>7.34</td>
<td>Zero-Base-Current Oscillator With Controllable Distortion.</td>
<td>164</td>
</tr>
<tr>
<td>7.35</td>
<td>Performance Curves for Circuit of Figure 7.34.</td>
<td>165</td>
</tr>
</tbody>
</table>
### LIST OF FIGURES (Continued)

<table>
<thead>
<tr>
<th>Figure</th>
<th>Description</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>7.36</td>
<td>LC Oscillator.</td>
<td>166</td>
</tr>
<tr>
<td>7.37</td>
<td>Alpha Characteristics of a Typical Type CK 720 Junction Transistor</td>
<td>167</td>
</tr>
<tr>
<td>7.38</td>
<td>Crystal-Controlled Oscillator.</td>
<td>168</td>
</tr>
<tr>
<td>7.39</td>
<td>Variation of Frequency With Collector Supply Voltage for Various Values of Emitter Supply Voltage</td>
<td>169</td>
</tr>
<tr>
<td>7.40</td>
<td>Compensated Crystal Oscillator With Single-Battery Supply.</td>
<td>170</td>
</tr>
<tr>
<td>7.41</td>
<td>Variation of Frequency With Supply Voltage for Two Values of R</td>
<td>171</td>
</tr>
<tr>
<td>7.42</td>
<td>Frequency-Versus-Temperature Characteristics for circuit of Figure 7.40.</td>
<td>172</td>
</tr>
<tr>
<td>7.43</td>
<td>Wide-Range Crystal Oscillator.</td>
<td>173</td>
</tr>
<tr>
<td>7.44</td>
<td>Simplified Pierce Oscillator.</td>
<td>174</td>
</tr>
<tr>
<td>7.45</td>
<td>Overtone Oscillator.</td>
<td>175</td>
</tr>
<tr>
<td>7.46</td>
<td>1000-kc Meacham Bridge Transistor Oscillator.</td>
<td>176</td>
</tr>
<tr>
<td>7.47</td>
<td>Junction-Transistor Meacham Bridge Oscillator.</td>
<td>179</td>
</tr>
<tr>
<td>7.48</td>
<td>Lamp-Stabilized Oscillator.</td>
<td>182</td>
</tr>
<tr>
<td>7.49</td>
<td>Pierce Oscillator With Isolation Stage.</td>
<td>183</td>
</tr>
<tr>
<td>7.50</td>
<td>Transistorized Butler Oscillator.</td>
<td>184</td>
</tr>
<tr>
<td>9.1</td>
<td>Equivalent Circuit Involving h-Parameters.</td>
<td>194</td>
</tr>
<tr>
<td>9.2</td>
<td>Equivalent Circuits and Important Relations of a Grounded-Base Stage.</td>
<td>196</td>
</tr>
<tr>
<td>9.3</td>
<td>Basic Circuits for Measuring h-Parameters.</td>
<td>197</td>
</tr>
<tr>
<td>9.4</td>
<td>The Effect of Signal Current on $R_1$ and $H_2$, $R_c = 1k$</td>
<td>200</td>
</tr>
<tr>
<td>9.5</td>
<td>The Effect of Signal Current on $R_1$ and $H_2$, $R_c = 2.7k$</td>
<td>200</td>
</tr>
<tr>
<td>9.6</td>
<td>An External Limiter Reduces Distortion.</td>
<td>201</td>
</tr>
</tbody>
</table>

(Continued)
LIST OF FIGURES (Concluded)

<table>
<thead>
<tr>
<th>Figure</th>
<th>Description</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>9.7</td>
<td>Variation of $R_1$ and $H_2$ With Emitter Bias</td>
<td>203</td>
</tr>
<tr>
<td>9.8</td>
<td>Variation of $R_1$ and $H_2$ With Collector Bias</td>
<td>203</td>
</tr>
<tr>
<td>9.9</td>
<td>Grounded-Collector Transistor Circuit</td>
<td>204</td>
</tr>
<tr>
<td>9.10</td>
<td>Characteristic Curves for a Typical Point-Contact Transistor</td>
<td>207</td>
</tr>
<tr>
<td>9.11</td>
<td>Effect of Collector Voltage on Input Characteristic</td>
<td>208</td>
</tr>
<tr>
<td>9.12</td>
<td>Idealized Forward Characteristics of a Point-Contact Transistor</td>
<td>210</td>
</tr>
<tr>
<td>9.13</td>
<td>Idealized Input Volt-Ampere Characteristics for Circuit of Figure 9.9</td>
<td>212</td>
</tr>
<tr>
<td>13.1</td>
<td>Negative-Resistance Oscillator</td>
<td>228</td>
</tr>
<tr>
<td>13.2</td>
<td>Harmonic Distortion and Frequency as Functions of Collector Supply Voltage</td>
<td>231</td>
</tr>
<tr>
<td>13.3</td>
<td>Circuit Diagram of the Test Oscillator</td>
<td>232</td>
</tr>
<tr>
<td>13.4</td>
<td>Comparison of the Computed Natural Frequency and the Actual Frequency</td>
<td>233</td>
</tr>
<tr>
<td>13.5</td>
<td>Frequency Comparison Curves Including the Effects of the First Eight Harmonics</td>
<td>233</td>
</tr>
<tr>
<td>13.6</td>
<td>Results Obtained With Battery Operation</td>
<td>235</td>
</tr>
<tr>
<td>13.7</td>
<td>Circuit for Obtaining Input Characteristics</td>
<td>236</td>
</tr>
<tr>
<td>13.8</td>
<td>Characteristics Employed for Oscillator Design</td>
<td>238</td>
</tr>
<tr>
<td>13.9</td>
<td>1-kc Oscillator</td>
<td>239</td>
</tr>
</tbody>
</table>
**LIST OF TABLES**

<table>
<thead>
<tr>
<th>Table</th>
<th>Title</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>2.1</td>
<td>RELATIVE FREQUENCY STABILITY OF THE TUNED-EMITTER OSCILLATOR OF FIGURE 2.4 EVALUATED AT THE FREQUENCY ( \omega = 1/\sqrt{LC} )</td>
<td>19</td>
</tr>
<tr>
<td>2.2</td>
<td>RELATIVE FREQUENCY STABILITY OF THE TUNED-COLLECTOR OSCILLATOR OF FIGURE 2.5 EVALUATED AT THE FREQUENCY ( \omega = 1/\sqrt{LC} )</td>
<td>20</td>
</tr>
<tr>
<td>2.3</td>
<td>CALCULATED RELATIVE FREQUENCY STABILITY OF ELEMENTARY NEGATIVE-RESISTANCE TRANSISTOR OSCILLATORS</td>
<td>22</td>
</tr>
<tr>
<td>5.1</td>
<td>DESIRABLE CHARACTERISTICS OF TRANSISTORS FOR OSCILLATOR PROTOTYPES</td>
<td>98</td>
</tr>
<tr>
<td>5.2</td>
<td>TRANSISTOR TYPES AVAILABLE TO THE PROJECT</td>
<td>99</td>
</tr>
<tr>
<td>7.1</td>
<td>CHARACTERISTICS OF OSCILLATOR CIRCUITS INVESTIGATED BY THE PROJECT</td>
<td>120</td>
</tr>
<tr>
<td>7.2</td>
<td>RECOMMENDED CIRCUITS</td>
<td>122</td>
</tr>
<tr>
<td>Symbol</td>
<td>Description</td>
<td>First Appears on Page</td>
</tr>
<tr>
<td>--------</td>
<td>-------------</td>
<td>-----------------------</td>
</tr>
<tr>
<td>A</td>
<td>four times the available amplifier power output for one volt input.</td>
<td>67</td>
</tr>
<tr>
<td>a</td>
<td>coefficients of Fourier series</td>
<td>45</td>
</tr>
<tr>
<td>a</td>
<td>transformer turns ratio</td>
<td>68</td>
</tr>
<tr>
<td>a</td>
<td>current amplification factor</td>
<td>17</td>
</tr>
<tr>
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<td>(as a subscript) base</td>
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</tr>
<tr>
<td>b</td>
<td>susceptance</td>
<td>38</td>
</tr>
<tr>
<td>c</td>
<td>(as a subscript) collector</td>
<td>17</td>
</tr>
<tr>
<td>C</td>
<td>capacitance</td>
<td>13</td>
</tr>
<tr>
<td>D</td>
<td>deviation, inverse of stability</td>
<td>10</td>
</tr>
<tr>
<td>d</td>
<td>mathematical derivative operator</td>
<td>15</td>
</tr>
<tr>
<td>E</td>
<td>oscillator supply voltage</td>
<td>34</td>
</tr>
<tr>
<td>e</td>
<td>(as a subscript) emitter</td>
<td>17</td>
</tr>
<tr>
<td>e</td>
<td>alternating current voltage</td>
<td>35</td>
</tr>
<tr>
<td>f</td>
<td>nominal frequency</td>
<td>5</td>
</tr>
<tr>
<td>f</td>
<td>farad</td>
<td>23</td>
</tr>
<tr>
<td>f_0</td>
<td>resonant frequency</td>
<td>48</td>
</tr>
<tr>
<td>G or g</td>
<td>conductance</td>
<td>58</td>
</tr>
<tr>
<td>h</td>
<td>henry</td>
<td>23</td>
</tr>
<tr>
<td>h_{lk}</td>
<td>hybrid matrix parameter</td>
<td>193</td>
</tr>
<tr>
<td>I or i</td>
<td>current</td>
<td>14</td>
</tr>
<tr>
<td>Im</td>
<td>imaginary part of</td>
<td>9</td>
</tr>
<tr>
<td>i</td>
<td>(as a subscript) imaginary part of</td>
<td>12</td>
</tr>
</tbody>
</table>

(Continued)
## LIST OF SYMBOLS (Continued)

<table>
<thead>
<tr>
<th>Symbol</th>
<th>Description</th>
<th>First Appears on Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>j</td>
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<td>13</td>
</tr>
<tr>
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</tr>
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</tr>
<tr>
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<td>23</td>
</tr>
<tr>
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<td>66</td>
</tr>
<tr>
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<td>inductance</td>
<td>13</td>
</tr>
<tr>
<td>M</td>
<td>resistance slope</td>
<td>52</td>
</tr>
<tr>
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<td>127</td>
</tr>
<tr>
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<td>8</td>
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<tr>
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<td>37</td>
</tr>
<tr>
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<td>pulling factor</td>
<td>48</td>
</tr>
<tr>
<td>PI</td>
<td>performance index</td>
<td>82</td>
</tr>
<tr>
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<td>quality factor of a resonator</td>
<td>16</td>
</tr>
<tr>
<td>R or r</td>
<td>resistance</td>
<td>13</td>
</tr>
<tr>
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<td>(as a subscript) real part of</td>
<td>12</td>
</tr>
<tr>
<td>r_o</td>
<td>amplifier or transistor internal impedance</td>
<td>67</td>
</tr>
<tr>
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<td>real part of</td>
<td>9</td>
</tr>
<tr>
<td>r</td>
<td>(as a subscript) series resonant</td>
<td>82</td>
</tr>
<tr>
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<td>short-time stability</td>
<td>7</td>
</tr>
<tr>
<td>S_s</td>
<td>general stability factor</td>
<td>6</td>
</tr>
<tr>
<td>S_t</td>
<td>temperature stability</td>
<td>6</td>
</tr>
<tr>
<td>S_v</td>
<td>voltage stability</td>
<td>5</td>
</tr>
</tbody>
</table>

(Continued)
<table>
<thead>
<tr>
<th>Symbol</th>
<th>Description</th>
<th>First Appears on Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>s</td>
<td>a specific parameter such as resistance, capacitance, inductance, etc.</td>
<td>6</td>
</tr>
<tr>
<td>T or t</td>
<td>temperature</td>
<td>6</td>
</tr>
<tr>
<td>V or v</td>
<td>voltage</td>
<td>5</td>
</tr>
<tr>
<td>WA</td>
<td>wave analyzer</td>
<td>35</td>
</tr>
<tr>
<td>X or x</td>
<td>reactance</td>
<td>9</td>
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<td>matrix admittance parameter</td>
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<td>very large number; when used with a capacitor it means a perfect bypass, and when used with an inductance or resistance it means an infinite impedance at the operating frequency.</td>
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<td>summation sign.</td>
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PURPOSE

The purpose of this project is to evolve data and information regarding transistor oscillator circuitry and the degree of reliability and reproducibility attained with transistor parameter variations. It includes both analytical and experimental studies of crystal-controlled and free-running oscillators employing transistors exclusively as active elements.

The work is divided into three principal tasks as follows:

1. The evaluation of crystal-controlled and free-running transistor oscillators from the point of view of reliability, size, weight, circuit requirements, frequency limitations and other similar considerations;

2. The development of new circuits and/or techniques specifically to take advantage of the unique characteristics of transistors regarding size, weight, power requirements, etc.; and

3. The development of practical engineering data and information to serve as a basis for designing free-running and crystal-controlled junction and point-contact transistor oscillators in the range 0.1 to 75 mc.
ABSTRACT

A 2-year investigation of oscillator theories and transistor oscillator circuits has been made to study design theories and to develop procedures which will make possible the rapid design of stable transistor oscillators. The results of this investigation, as well as other pertinent information, are presented in the several sections of this report.

Several oscillator theories and methods of analysis have been studied. Among the first of these was the linear-circuit analysis. However, it was learned early in the study that the stability of an oscillator is significantly affected by the nonlinear behavior of the active element. Thus, for most oscillators, the linear analysis was not adequate since the distortion and other nonlinear effects are usually appreciable. Also, the linear analysis becomes very complicated as the number of variables is increased. It is concluded that the linear analysis leaves much to be desired for routine transistor oscillator design.

Various approaches to the nonlinear analysis are discussed. The most promising analysis is based on the work of Groszkowski, which has been extended to provide a procedure for accurately predicting the instability of transistor oscillators when the frequency of operation is low compared to the alpha cut-off frequency of the transistor. Although this theory greatly increased the understanding of the nature of transistor oscillators, it did not always lead to a direct method of improving the stability. In addition, in its present form, it is applicable only to the negative-resistance type of oscillator. When investigating the validity of the Groszkowski equation at higher frequencies, factors other than distortion were found to limit
the stability. Among these factors were changes in element capacitances and changes in transit times as functions of voltage and temperature. The theory has not been sufficiently advanced to include these factors but it can probably be modified to include these factors. If this were done, the theory would be useful at much higher frequencies.

In free-running oscillators, it is possible for the frequency pulling due to distortion to be as large as 50 per cent of the operating frequency. Even in oscillators which are reasonably linear, the pulling may be as large as one per cent. In several of the circuits tested, the Groszkowski equation accounted for as least 80 per cent of the frequency variation.

In addition to its direct usefulness, the Groszkowski analysis has predicted several possible methods of improving the stability of both negative-resistance and feedback oscillators. Notable among these methods is the stabilization by distortion which may improve the stability by an order of magnitude.

Various theories describing the behavior of feedback transistor oscillators are also discussed. The most promising of these theories is one which assumes linear operation of the transistor. Application of this theory, then, results in a design which causes the operation to be sufficiently linear to obtain good accuracy. This procedure also provides a method of optimizing any existing feedback-oscillator design.

The empirical approach is discussed as an aid in designing transistor oscillators. This approach consists of choosing an oscillator and then varying the parameters to obtain the best stability. Correlation of the results obtained in this way with the various theories has usually produced agreement.
However, some improvements which have not been completely explained by any of the theories have been empirically obtained. One such procedure was the stabilization by distortion mentioned above. The empirical approach has also resulted in compensation methods which have been used often to improve the stability of transistor oscillators by as much as an order of magnitude.

The importance of the proper selection of crystals and LC resonant circuits is stressed. One of the conclusions which is discussed is that crystals operate best when used in their series mode, while LC resonant circuits usually perform better when connected for parallel resonance. The first of these conclusions results from the low impedances involved in most transistor circuits. The second is due to the stray capacitances of inductances. Such capacitances result in fewer spurious resonances when the elements are connected for parallel resonance.

The evaluation of transistor oscillator performance with regard to transistor parameters is not stressed because of the difficulty of measuring the parameters at higher frequencies. Any design theory which depends upon critical relations among the transistor parameters will lead to poor transistor replaceability.

Transistor oscillator circuits are evaluated primarily with regard to frequency variations resulting from changes in voltage. One reason for this is that the frequency is easily measured and the voltage is the most readily varied parameter. It was found that the variation in frequency for changes in temperature and other external factors can usually be predicted from the frequency-voltage characteristics.

In designing transistor oscillators, the size and the weight of components were considered. Circuits requiring large or heavy components were immediately rejected.
Emphasis was placed on the design of new circuits utilizing the small size and low power consumption of transistors. Appreciable consideration was not given to the power output of circuits since this factor depends directly on the permissible transistor power dissipation. Also, because of the small size of transistors, a power-output stage can be readily added to any of the oscillators presented.

Many of the oscillators which have been developed have frequency stabilities as good as their vacuum-tube counterparts for the same relative operating frequency. Vacuum tubes are usually operated at frequencies far below their alpha cut-off frequency where good stabilities are to be expected. However, the tendency has been to operate transistors as near to their cut-off frequency as possible. In some cases, transistors have been operated at frequencies as great as ten times the alpha cut-off frequency. This operating condition results in reduced stabilities.

The properties of resonators, transformers and other components are discussed. Suggestions for making the proper selection of such components along with suggestions on general construction procedures are included.
I. INTRODUCTION

A. History of the Contract

This project, entitled "Transistor Oscillators of Extended Frequency Range," was sponsored by the U. S. Army Signal Corps through the Frequency Control Branch of the Squier Signal Laboratories, Fort Monmouth, New Jersey, under Contract No. DA-36-039-sc-42712. This contract became effective March 15, 1953, and was later modified to terminate May 31, 1954. The contract was renewed for one additional year, extending through May 31, 1955. The total man-hours involved was equivalent to approximately four man-years.

B. Survey of the Problem

The principal objective of this work was to develop practical engineering data to aid in the rapid design of stable transistor oscillators capable of high frequency operation. In the beginning, it appeared that only point-contact transistors would be suitable for use at frequencies above a few megacycles. However, during the life of the contract, many advances were made in the development of junction transistors so that at the present time junction-transistor oscillators may be successfully operated at frequencies well above a hundred megacycles, and the limit has not yet been reached. Because of the fact that junction transistors may be produced with more uniform characteristics than point-contact transistors, the former are usually preferred when a choice is permissible. Although many transistor oscillator circuits have been described in the literature, few have been systematically studied. For the most part, these circuits have been derived from analogous vacuum-tube circuits, frequently with little objective other than that they oscillate. Thus, quantitative data on the stability of transistor oscillators with respect to changes in environment
such as supply voltage, temperature, etc., have been almost entirely unavai-
ble. The study described in this report is concerned primarily with oscillators
possessing reasonably good frequency stability.

Techniques for manufacturing transistors have been steadily improved and,
no doubt, will continue to improve for several years. For this reason, certain
limitations present in earlier transistors have become less and less important.
The development of silicon transistors may almost eliminate the problem of drift
associated with high temperatures. Also, the over-all reliability of transis-
tors continues to be improved. These considerations make it desirable not to
overemphasize limitations of specific transistor types. Instead, the work has
been aimed at establishing a background of information that is likely to con-
tinue to be useful for some time. Thus, although design methods are presented,
relatively little attention is paid to the particular values of the parameters
of the transistors used in the various circuits.

For the same reasons, a start has been made toward applying the theory of
nonlinear oscillators to transistor circuits. In particular, the relation be-
tween waveform distortion and frequency stability, first studied by Groszkowski,
has been extended and reduced to engineering terms. However, this work is not
finished. The material is presented in a form which is useful only with oscil-
lators employing point-contact transistors at comparatively low frequencies.
However, this information has added substantially to a general understanding of
oscillator performance and has suggested several circuits that may have impor-
tant practical applications. These circuits are described in detail in this re-
port. Several other methods of analysis are described briefly. Except for con-
ventional linear analysis based on small-signal parameters, these other methods
do not appear to contribute much to the problem. Only a few of the circuits described have been arrived at by straightforward design procedures. Each circuit arrived at in this way suggests numerous variations that are more easily investigated empirically.

Several observations of a general nature have been made concerning the use of resonant circuits and quartz crystals. For example, LC circuits are best adapted for use as parallel circuits. Attempts to design series-resonant LC circuits having high Q's almost invariably lead to spurious resonances and parasitic oscillations. On the other hand, quartz crystals are best adapted for use as series-resonant branches because resonant impedances in the parallel mode are usually so large that they are incompatible with other impedances of transistor circuits. Of course, there are exceptions to these statements.

Transistor oscillators have been evaluated primarily with regard to frequency variations resulting from changes in power-supply voltages. The main reason for this choice is that the measurements are easy to make, whereas measurements involving temperature variations or transistor-parameter variations are comparatively difficult to make. The voltage stability of a circuit is usually a good index of its performance under other conditions. Consequently, the voltage stability is considered to be a satisfactory basis for comparing one circuit with another. The reliability of the circuits over long periods of time has not been investigated.

Several of the developed circuits provide frequency stabilities that compare favorably with those of vacuum-tube oscillators under similar conditions. The best of these is a bridge-stabilized oscillator patterned after the Meacham circuit. Other promising circuits make use of diode rectifiers to compensate for changes in power-supply voltages.
II. OSCILLATOR THEORIES AND DESIGN METHODS

A. General Requirements and Definitions

In advancing a theory to aid in the design of transistor oscillators, several important things must be considered. First, the theory must be accurate and applicable to practice. It must not require excessive computation on the part of the technician. The theory must have parameters which are readily determined from the circuit configuration or readily measured from the transistor and other components used. The approximations must be realistic and all variables must be carefully considered. One theory should be applicable to as many circuit configurations as possible. The theory should not exclude future developments and improvements in transistors and other components. It should indicate the desirable active element characteristics to guide in the choice of transistors. Methods of improving the oscillator stability by the proper selection of components should be indicated. The theory should yield, quantitatively, the expected stability and power output. And last, but not least, it should provide a criterion for future transistor design and development.

This project has not developed a single theory which accomplishes all of these objectives. However, some of the theories proposed do satisfy several of these requirements. The limitations of each of the theories and design procedures presented in this report will be discussed in section II-H.

Several definitions should be presented before elaborating on the theories and design procedures. These definitions are as follows:

\[ S_v = \text{voltage stability} = \frac{(\Delta v)}{(v)} \times 10^{-4} \]  

(2.1)

where: \( \Delta v \) is the total change in the supply voltage in volts,

\( \Delta f \) is the frequency change in cycles per second caused by \( \Delta v \),
v is the nominal operating voltage in volts and

f is the nominal frequency of oscillation in cycles per second.

Voltage stability as here defined is the reciprocal of ppm for one percent voltage change. $S_v$ increases as the stability is improved. Typical values for this quantity will range from 0.1 for average free-running oscillators to 1000 for exceptional crystal oscillators. The unit of $S_v$ is the percentage of voltage change/ppm frequency change.

$$S_v = \text{temperature stability} = \frac{(\Delta t) (f)}{(\Delta f)} \times 10^{-6} \quad (2.2)$$

where: \(\Delta t\) is the change in temperature in degrees centigrade,

\(\Delta f\) is the frequency change in cycles per second caused by \(\Delta t\) and

\(f\) is the nominal frequency of oscillation in cycles per second.

$S_t$ is the reciprocal of ppm for one-degree-centigrade temperature change. The unit of $S_t$ is degrees/ppm frequency change.

$$S_t = \frac{(\Delta s) (f)}{(\Delta f) (s)} \times 10^{-4} \quad (2.3)$$

where: \(\Delta s\) is the change in a specific parameter other than temperature or voltage (examples: resistance, capacitance, inductance, etc.)

\(\Delta f\) is the frequency change in cycles per second caused by \(\Delta s\),

\(s\) is the nominal value of the specific parameter expressed in the same unit of measure as \(\Delta s\) and

\(f\) is the nominal frequency of oscillation in cycles per second.

The name applied to this unit would be determined by the parameter which is varied.
where: $\Delta f$ is the maximum change in frequency in cycles per second over a period of 4 hours under specified conditions of voltage, temperature and other variables and 

$f$ is the nominal frequency of oscillation in cycles per second.

The unit of $S$ is megacycles/cycle.

The units of voltage stability, temperature stability and short-time stability have been arbitrarily adjusted so as to yield convenient magnitudes of these quantities for most crystal-controlled and free-running oscillators.

B. The Linear Circuit Approach

The frequency of an oscillator is usually determined primarily by either a resonant circuit (resonator) or a phase-shifting network. For purposes of analysis, it may be assumed that the parameters of these circuits are sufficiently independent of their environment and that, if they alone determined the oscillator frequency, an adequate stability would be obtained. Of course, this is not necessarily true in practice, but it is certainly reasonable to suppose the resonator to be the most stable part of a transistor oscillator. Hence, failure of an oscillator to maintain satisfactory stability is principally due to either nonlinear effects or changes in the parameters of circuits external to the resonator, or to both. In either case, the difficulties are usually associated with comparatively isolated portions of the circuit, for example, with the amplifier or with the load. Accordingly, an oscillator whose frequency is fixed by a resonator may be supposed to consist of three distinct parts as indicated in Figure 2.1.
Here, $Z_0$ represents the resonator; $Z_2$ represents an element that is supposed variable; $N$ is a coupling network; and $Z_1$ is the impedance of the network terminated in $Z_2$. Since any portion of a complete circuit may be designated $Z_2$, various components may be investigated one at a time and their effects combined to obtain the total effect of all variations.

In the case of a crystal-controlled oscillator for example, $Z_0$ represents the crystal regardless of what the rest of the circuit is like. The advantage of this two-terminal-oscillator point of view is that it separates the crystal, whose characteristics are known and supposed constant, from the part of the circuit which is considered to be inherently unstable. Thus, it is clear that, given the variable element $Z_2$, the essential problem is to design the network, $N$, so as to realize the best use of the intrinsic stability of the crystal. Or, carrying this idea further, given transistors with parameters varying over a specified range, design the network so that the over-all stability is as near as possible to that of the crystal.

The criterion for sustained oscillations in the circuit of Figure 2.1 is

$$Z_0 + Z_1 = 0$$

(2.5)
and, consequently,

\[ \text{Re}[Z_0] = -\text{Re}[Z_1], \quad (2.6) \]
\[ \text{Im}[Z_0] = -\text{Im}[Z_1]. \quad (2.7) \]

Since some form of amplitude limiting is always present, equation 2.6 determines the amplitude of oscillations and equation 2.7 determines the frequency.

Equations 2.5, 2.6 and 2.7 are strictly correct only for a linear system because the ordinary concept of impedance is valid only for linear systems. However, if the system is almost linear, a negligible error results from the use of these equations. In one of the simplest situations, \( Z_0 \) consists of a series RLC circuit operating near its resonant frequency. Then, since the \( \text{Re}[Z_0] \) is constant, the frequency stability may be investigated by considering the imaginary components only. Suppose that \( Z_2 \) is changed by some external means. This will result in a change of \( Z_1 \), but because of the amplitude limiting, only the imaginary part of \( Z_1 \) changes. Equilibrium must be restored according to the relation

\[ \Delta X_o + \Delta X_1 = 0, \quad (2.8) \]

in which \( X_o \) and \( X_1 \) are the imaginary parts of \( Z_o \) and \( Z_1 \), respectively. In terms of the assumed variables, equation 2.8 becomes

\[ \frac{\partial X_o}{\partial \omega} \Delta \omega + \frac{\partial X_1}{\partial Z_2} \Delta Z_2 + \frac{\partial X_1}{\partial \omega} \Delta \omega = 0 \quad (2.9) \]

from which it follows that
Let $D_2 = \frac{\partial\omega}{\partial Z_2}$, which is a measure of the absolute frequency stability (more precisely, the absolute frequency instability) with respect to changes of $Z_2$. Also, let $K_{12} = \frac{\partial X_1}{\partial Z_2}, K_{01} = \frac{\partial X_0}{\partial \omega}$ and $K_{11} = \frac{\partial X_1}{\partial \omega}$. In terms of these definitions, equation 2.10 becomes

$$D_2 = \frac{-K_{12}}{K_{01} + K_{11}} .$$

This result emphasizes the fact that the over-all stability depends jointly on the properties of the resonator and the remainder of the circuit. In the situation of interest here, $K_{01} > K_{11}$ and, therefore,

$$D_2 \approx \frac{K_{12}}{K_{01}} .$$

Moreover, $K_{01}$ must be regarded as fixed, and so it becomes evident that the circuit design problem is to make $K_{12}$ small without upsetting equation 2.6.

If a certain circuit modification multiplies $K_{12}$ by some factor, then, so far as $D_2$ is concerned, this is equivalent to dividing $K_{01}$ by the same factor. For example, the bridge arrangement of the Meacham "bridge-stabilized oscillator" is usually considered to produce an increase of the "effective $Q$" of the crystal. Physically, however, the crystal is not changed, and according to the above argument, it should be quite as satisfactory to think of the bridge as improving the amplifier instead of the crystal.
The frequency stability of an oscillator is significantly affected by the nonlinear behavior of elements in the oscillator circuit. Therefore, it is to be expected that a linear analysis of an oscillator circuit cannot provide a complete description of the variation of operating frequency. In spite of this fact, it is found that the first-order effects on frequency stability as described by a linear analysis are adequate for most oscillators which are highly stable in frequency and nearly sinusoidal in operation.

The first step in the linear analysis of an oscillator is the selection of linear parameters which give an adequate mathematical description of the performance of the oscillator. It is often convenient, but not necessary, to draw an equivalent circuit consisting of various resistances, inductances, capacitances and sources to represent these parameters. The linear parameters are defined in terms of the voltages and currents of fundamental frequency in the oscillator. Thus, the driving-point impedance of a two-terminal element is defined as the complex ratio of fundamental voltage to fundamental current. This impedance is a complicated function of both the voltage and current waveshape and the frequency for the element. It can be specified and measured only under operating conditions for the circuit.

According to the above definition of impedance, a nonlinear pure resistor may be described by a linear impedance having both a real and an imaginary part. This is possible because the fundamental component of voltage in a nonlinear resistor may be out of phase with the fundamental component of current. As mentioned above, this impedance is a complicated function of both the voltage and the current of the resistor.

If the nonlinear element being analyzed has more than two terminals, the linear parameters selected to describe the operation cannot, in general, be
specified uniquely because the principle of superposition is no longer applicable. In this case, the selection of the parameters may be based on measurements taken when the element is operating under as nearly linear conditions as possible.

The linear parameters which are selected may be used to write a set of homogeneous algebraic equations describing the steady-state performance of the oscillator system. The condition for self-sustained oscillations may be obtained by setting the determinant of this set of equations equal to zero. Two equations of equilibrium result from this procedure. One is obtained by setting the real part of the determinant equal to zero; the other is obtained by setting the imaginary part of the determinant equal to zero. The oscillator frequency may be obtained from these equations of equilibrium if all other parameter values are known. (However, no information may be derived concerning the amplitude of the oscillations.) The closeness to which the predicted frequency approximates the actual frequency is dependent upon the accuracy with which the linear mathematical system describes the performance of the oscillator system.

The variation of frequency produced by a variation in a given parameter, $s$, of the circuit may be obtained from the two equations of equilibrium. All parameters other than the frequency of oscillation, $\omega$, the real part, $s_r$, of the variable parameter and the imaginary part, $s_i$, of the variable parameter are held fixed. The two equations of equilibrium may then be written as

\begin{align}
    f_1 (\omega, s_r, s_i) &= 0, \\
    f_2 (\omega, s_r, s_i) &= 0.
\end{align}

(2.13)  

(2.14)
The frequency variation of the circuit may be obtained through determining the locus of $s_r + js_i$ for variable frequency. The equation of this locus may be obtained by eliminating $\omega$ from equations 2.13 and 2.14.

The relative frequency stability of the oscillator with respect to variations in $s$ may be defined as

$$S_s = \frac{\omega}{|s|} \lim_{\Delta \omega \to 0} \frac{|\Delta s|}{|\Delta \omega|} = \frac{\omega}{|s|} \left| \frac{ds}{d\omega} \right| = \left| \frac{\omega}{s} \frac{ds}{d\omega} \right|. \quad (2.15)$$

Here, $\left| \frac{ds}{d\omega} \right|$ is the rate of change of arc length along the locus of the complex variable $s$ with respect to frequency $\omega$. The relative frequency stability, $S_s$, is taken to be a measure of the tendency of the oscillator system to resist changes in frequency.

The simple negative-resistance oscillator shown in Figure 2.2 may be analyzed to provide a demonstration of the procedure described previously.

Figure 2.2. Equivalent Circuit of Negative-Resistance Oscillator.

The elements $R_1$, $L$ and $C$ are assumed to be linear, and the linear parameters describing their operation are assumed to be the impedances $R_1$, $j\omega L$ and $-j/\omega C$ respectively. It is assumed that the element $R_2$ may be nonlinear and may vary in a random manner. The linear parameter describing the operation of the element
R_2 is taken to be the impedance \( R_{2r} + jR_{2i} \). Any possible relationship between the fundamental voltage and fundamental current of \( R_2 \) may be described by proper values of the real and imaginary parts of this impedance.

The equation describing the operation of this circuit is

\[
I \left( R_{2r} + jR_{2i} + R_1 + j\omega L - \frac{j}{\omega C} \right) = 0. \quad (2.16)
\]

The condition for sustained oscillations is

\[
R_{2r} + jR_{2i} + R_1 + j\omega L - \frac{j}{\omega C} = 0. \quad (2.17)
\]

The real and imaginary parts of equation 2.17 may be set equal to zero to provide two equations of equilibrium:

\[
R_{2r} + R_1 = 0, \quad (2.18)
\]

\[
R_{2i} + \omega L - \frac{1}{\omega C} = 0. \quad (2.19)
\]

Finally, equations 2.18 and 2.19 may be rewritten as

\[
R_{2r} = - R_1, \quad (2.20)
\]

\[
\omega = \sqrt{\frac{1}{LC} + \left( \frac{R_{2i}}{2L} \right)^2} - \frac{R_{2i}}{2L}. \quad (2.21)
\]

It is evident from equation 2.20 that the real part of the impedance describing \( R_2 \) must be negative and exactly equal in magnitude to \( R_1 \). If the element \( R_2 \) is not able to adjust itself to meet this condition by varying the waveforms and frequency of the voltage and current in the circuit, then the oscillator will cease operation. The operating frequency is identical with the series-resonant
frequency of the \( R_1LC \) combination only if \( R_{21} = 0 \). This special case requires that the fundamental component of current in the element \( R_2 \) be exactly 180 degrees out of phase with the fundamental component of voltage.

The locus of the variable parameter \( R_{2r} + jR_{21} \) as found by use of equations 2.18 and 2.19 is shown in Figure 2.3. This locus is exactly the negative of the locus of the impedance of the \( R_1LC \) series combination for variable frequency.

The relative frequency stability of the oscillator, evaluated at the series-resonant frequency of the \( R_1LC \) combination, is found by differentiating equation 2.19:

\[
S_s' = \omega \frac{dR_2}{dR} = \frac{1}{\sqrt{LC}} \left( L + \frac{1}{C \omega^2} \right) = 2 \sqrt{\frac{L}{C}} \tag{2.22}
\]

or,

\[
\frac{S_s'}{R_2} = S_s = \frac{\omega}{R_2} \frac{dR_2}{dR} = 2 \sqrt{\frac{L}{C}} = 2Q. \tag{2.23}
\]

Equation 2.22 apparently shows that frequency stability does not depend on the value of the resonator resistance \( R_1 \). Equation 2.23 seems to indicate that frequency stability is very much dependent on the value of \( R_1 \). The question as to which result gives a more accurate picture depends on the nature of the response of the nonlinear circuit to the basic variation, such as temperature or supply voltage, which is actually causing the frequency deviation. Equation 2.23 should be used if a given increment in the primary variable, such as temperature, etc., tends to produce a constant phase shift in the impedance describing the fundamental components of the nonlinear resistance \( R_2 \). Equation 2.22 should be used.
if this given increment in the primary variable produces a given change in the imaginary component of the impedance describing the resistor $R_2$, regardless of the value of the resistance $R_2$. In some situations perhaps neither definition gives an accurate description of the nature of the nonlinear element. However, it is very important to note that in both of these cases the frequency stability can be improved by increasing the $L/C$ ratio. Therefore, it seems reasonable to suppose that an increase in this ratio results in an increase in frequency stability regardless of the nature of the nonlinear resistor used in the circuit. It seems just as important to emphasize that a decrease in the value of the resonator resistance, or a corresponding increase in the value of $Q$, does not necessarily result in an increase in frequency stability. Whether the stability improves or not depends upon the type of negative-resistance element used in the oscillator and the conditions under which it is operated.

![Diagram](attachment:Figure_2.3.png)

Figure 2.3. Locus of $R_{2r} + jR_{2i}$ for Variable Frequency.
A more complicated analysis problem is the grounded-base transistor oscillator circuit shown in Figure 2.4.

![Equivalent Circuit of Grounded-Base Transistor Oscillator](image)

Figure 2.4. Equivalent Circuit of Grounded-Base Transistor Oscillator.

A set of homogeneous equations describing the steady-state performance of this circuit is

\[
0 = I_c (r_b + R_b) + I_e (j\omega L - j/\omega C + r_b + R_b + r_e + R_e),
\]

\[
0 = I_c (r_c + R_c + r_b + R_b) + I_e (r_b + R_b + 2ar_c). \tag{2.24}
\]

The condition for sustained oscillations is obtained by setting the determinant of this set of equations equal to zero.

\[
0 = (j\omega L - j/\omega C + r_e + R_e)(r_c + R_c + r_b + R_b)
\]

\[
+ (r_b + R_b) [R_c + (1 - a)r_c]. \tag{2.25}
\]

For perfectly linear operation, setting the real and imaginary parts, respectively, equal to zero yields the following two constraints on the operation of the circuit.

-17-
The relative frequency stability of the oscillator with respect to variation in any parameter may be determined by use of equation 2.25. For instance, suppose that all parameters remain fixed except \( r_b \). The parameter \( r_b \) is allowed to assume complex values \( r_{b\text{r}} + jr_{b\text{i}} \) in keeping with the constraints imposed by equation 2.25. These constraints may be stated explicitly by setting the real and imaginary parts of equation 2.25 equal to zero.

\[
0 = (r_{b\text{r}} + R_b) [r_e + R_e + R_c + (1 - a)r_c] + r_e + R_e)(r_c + R_c)
- r_{b\text{i}}(\omega L - \frac{1}{\omega C})
\]  
(2.27)

\[
0 = r_{b\text{i}} [r_e + R_e + R_c + (1 - a)r_c] + (r_c + R_c + r_{b\text{r}} + R_b)(\omega L - \frac{1}{\omega C})
\]  
(2.28)

The equation of the locus of \( r_{b\text{r}} + jr_{b\text{i}} \) for variable frequency may be found by eliminating \( \omega \) from equations 2.27 and 2.28.

\[
0 = r_{b\text{i}}^2 [r_e + R_e + R_c + (1 - a)r_c] + [(r_{b\text{r}} + R_b) [r_e + R_e + R_c + (1 - a)r_c] + (r_e + R_e)(r_c + R_c)](r_c + R_c + r_{b\text{r}} + R_b)
\]  
(2.29)

This locus is seen to be an ellipse with focii on the \( r_{b\text{r}} \) axis.

The relative frequency stability with respect to variations in \( r_b \) calculated at the frequency \( \omega = 1/\sqrt{LC} \) is found from equation 2.28.
\[
S_s = \frac{\omega}{r_b} \left| \frac{d^2 r}{d\omega^2} \right| = 2 \sqrt{\frac{L}{C}} \left( \frac{1}{r_e + R_e} \right) \left( 1 + \frac{R_b}{r_b} \right) \left( 1 + \frac{r_b + R_b}{r_c + R_c} \right) (2.30)
\]

A similar analysis may be carried out to determine the relative frequency stability with respect to variations in other parameters. The results of an analysis of this type are shown in Table 2.1. Table 2.2 shows the results of a similar analysis for the circuit of Figure 2.5.

### TABLE 2.1

RELATIVE FREQUENCY STABILITY OF THE TUNED-EMITTER OSCILLATOR OF FIGURE 2.4 EVALUATED AT THE FREQUENCY \( \omega = 1/\sqrt{LC} \)

<table>
<thead>
<tr>
<th>Variable</th>
<th>Relative Frequency Stability</th>
</tr>
</thead>
<tbody>
<tr>
<td>( r_b )</td>
<td>( \frac{2 \sqrt{L}}{r_b} \left</td>
</tr>
<tr>
<td>( r_c )</td>
<td>( \frac{2 \sqrt{L}}{r_c} \left( 1 + \frac{r_c + R_c}{r_b + R_b} \right) )</td>
</tr>
<tr>
<td>( r_e )</td>
<td>( \frac{2 \sqrt{L}}{r_e} )</td>
</tr>
<tr>
<td>( a )</td>
<td>( \frac{2 \sqrt{L}}{ar_c} \left( 1 + \frac{r_c + R_c}{r_b + R_b} \right) )</td>
</tr>
</tbody>
</table>

The formulas shown in Tables 2.1 and 2.2 may be used to design and compare the performance of oscillator circuits of the type shown in Figures 2.4 and 2.5 with respect to frequency stability. When using these formulas, however, it
TABLE 2.2

RELATIVE FREQUENCY STABILITY OF THE TUNED-COLLECTOR OSCILLATOR
OF FIGURE 2.5 EVALUATED AT THE FREQUENCY $\omega = 1/\sqrt{LC}$

<table>
<thead>
<tr>
<th>Variable</th>
<th>Relative Frequency Stability</th>
</tr>
</thead>
<tbody>
<tr>
<td>$r_b$</td>
<td>$\frac{2}{r_b \sqrt{LC}} \left( \frac{r_b + R_b}{r_c + R_c} \right) \left( 1 + \frac{r_b + R_b}{r_e + R_e} \right)$</td>
</tr>
<tr>
<td>$r_c$</td>
<td>$\frac{2}{r_c \sqrt{LC}} \left( 1 + \frac{r_e + R_e}{r_b + R_b} \right)$</td>
</tr>
<tr>
<td>$r_e$</td>
<td>$\frac{2}{r_e \sqrt{LC}} \left( 1 + \frac{r_e + R_e}{r_b + R_b} \right)$</td>
</tr>
<tr>
<td>$a$</td>
<td>$\frac{2}{2 r_c \sqrt{LC}} \left( 1 + \frac{r_e + R_e}{r_b + R_b} \right)$</td>
</tr>
</tbody>
</table>

should be remembered that the equivalent circuits that have been used do not provide an exact description of the performance of the physical oscillator. Furthermore, the frequency stability has been evaluated only for a small phase shift in a single element. In practice, variations of an external parameter such as bias voltage may cause variations in several or all of the parameters of the equivalent circuit.

In spite of these facts, this type of analysis may provide information which is useful in the design of transistor oscillators that are almost
sinusoidal in operation. In other words, an oscillator that is designed to be stable with respect to independent variations in all of the parameters of its equivalent circuit is expected to be stable with respect to collective variations in all parameters, assuming again that the operation is almost linear.

Frequency-stability tests of a negative-resistance oscillator have been made. Several transistors were used to evaluate the applicability of the type of analysis presented here. The experimentally determined performance was found to agree qualitatively with the predicted performance. The results of one such test are included here to show the general nature of the agreement between the predicted performance of the oscillator and the experimentally determined performance. A low-frequency, low-Q circuit was used for ease in measurement. The analysis should be applicable over as wide a frequency range as the equivalent circuit is valid. At higher frequencies a more complicated equivalent circuit undoubtedly will have to be used.
Equal values of external resistance were used in the emitter, base and collector leads. Therefore, the location of the LC tank was the only difference between the two circuits for this test. The oscillations were maintained close to the threshold condition to insure as nearly sinusoidal operation as possible.

A type 1768 point-contact transistor having the following measured parameter values was used in these tests.

\[
\begin{align*}
a &= 2.25 \\
r_c &= 3,500 \text{ ohms} \\
r_b &= 110 \text{ ohms} \\
r_e &= 40 \text{ ohms}
\end{align*}
\]

Each of the external resistances—\(R_v\), \(R_c\) and \(R_e\)—was 450 ohms. The tank coil \(L\) had an inductance of 0.5 henry, and the capacitor \(C\) had a capacitance of 0.0534 microfarad.

The relative frequency stabilities for the above parameter values, as calculated by use of Tables 2.1 and 2.2, are shown in Table 2.3.

**TABLE 2.3**

CALculated RELATIVE FREQUENCY STABILITY OF ELEMENTARY NEGATIVE-RESISTANCE TRANSISTOR OSCILLATORS

<table>
<thead>
<tr>
<th>Variable</th>
<th>(r_b)</th>
<th>(r_c)</th>
<th>(r_e)</th>
<th>(a)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Tuned Emitter (S_s)</td>
<td>73</td>
<td>37</td>
<td>150</td>
<td>6.2</td>
</tr>
<tr>
<td>Tuned Collector (S_s)</td>
<td>17</td>
<td>8.6</td>
<td>36</td>
<td>1.5</td>
</tr>
</tbody>
</table>

An inspection of Table 2.3 shows that the frequency of the tuned-emitter oscillator should be roughly four times as insensitive to phase shifts in the
parameters of the equivalent circuit as the frequency of the tuned-collector oscillator. It should be remembered that this prediction is made only for an oscillator circuit using the particular parameter values shown above. Entirely different predictions might be expected to result from a different circuit design. An inspection of Table 2.3 also shows that the frequency of oscillation of both circuits is more sensitive to phase shifts in $a$ and $r_e$ than to phase shifts in $r_b$ and $r_c$.

Measurements of the variation of operating frequency due to variations in bias were made for the two oscillator circuits described above. Circuit diagrams for the circuits used in the measurements are shown in Figures 2.6 and 2.7. In all cases, variations in bias were made from an operating point of $V_c = 14$ volts and $I_e = 2$ milliamperes. The results of these frequency-stability measurements are shown in Figures 2.8 and 2.9.

![Circuit Diagram](image)

**Figure 2.6. Tuned-Emitter Oscillator Circuit.**

An inspection of the curves of Figures 2.8 and 2.9 shows that the frequency of the tuned-emitter oscillator, as predicted, is several times as stable as that of the tuned-collector oscillator. Similar results were obtained for several transistors used in the same type circuit, provided that the operation was maintained close to the threshold of oscillation.

-23-
Figure 2.7. Tuned-Collector Oscillator Circuit.

Figure 2.8. Performance of the Circuit in Figure 2.6.
This same type of analysis may be applied to any of the two-terminal oscillators, and similar tables may be derived to indicate the instability with respect to any variable. Similar procedures may also be applied to feedback oscillators. To determine the stability of an actual oscillator from such a table, the transistor parameters must be known as a function of each basic variable such as voltage or temperature. The over-all frequency stability may, of course, be expressed in terms of the basic variables. However, it can be readily seen that this procedure would yield extremely complicated expressions. Even if such an expression were simple enough to be readily usable, it would not completely describe the oscillator performance since several factors such as stray capacitances, distortion and others do not appear in the expressions.

The use of a high-frequency equivalent circuit would further complicate the expressions for stability since additional transistor parameters would be introduced. Introduction of expressions for nonlinearities would not be permissible in some cases since some of the basic expressions are true only for linear operation. This method of analysis fails to indicate how well a given practical
oscillator meets the condition of linearity and thus fails to indicate the degree of accuracy of the results obtained. Therefore, the linear analysis is considered to have only limited usefulness. It does, however, provide an insight into some of the basic principles of oscillator design.

The advantage of this analysis is that the necessary parameter values are readily obtained from measurement or from published data on the transistors. Likewise, the parameters are of such nature that their variations with voltage and temperature can readily be obtained.

One of the disadvantages is that different equations must be used for each oscillator configuration.

C. The Isocline Diagram

In the previous section, the input characteristic of the transistor was considered to be linear. For almost linear operation to occur, it is necessary that the amplitude of oscillation be small. When this condition is not met, the results obtained from the linear analysis may be greatly in error. One method of taking the nonlinearity into account is to use the isocline diagram. By using this diagram, it is possible to determine the harmonic content of the oscillator output and also to study the effect of various circuit parameters, such as the resonator Q, on stability.

It was pointed out in the preceding section that, insofar as frequency stability alone is concerned, increasing the Q of the resonator of an oscillator does not necessarily result in improved performance. Rather, in this case, it is the L/C ratio that is basically more important. Although this fact has been recognized for a long time, a popular notion persists that high Q and a high degree of stability are inseparable. For this reason, it is worthwhile to include
some further analytical and experimental results. A word of warning is likewise appropriate. High-Q resonators are often desirable for other reasons, and it must also be remembered that these results apply specifically only to circuits of the series-mode type.

It is shown in equation 2.26, of the preceding section, that the circuit of Figure 2.10 will oscillate, provided

\[
a > \left(1 + \frac{R + r_e}{R_b + R_b}\right) \left(1 + \frac{r_c}{r_c}\right) + \frac{r_e + R}{r_c}.
\] (2.31)

![Transistor Oscillator Circuit](image)

Figure 2.10. Transistor Oscillator Circuit.

If \(r_c\) is very large compared to the other resistances, equation 2.31 yields the approximation

\[
a > 1 + \frac{R + r_e}{R_b + r_b}.
\] (2.32)

Thus, if the emitter-circuit resistance is sufficiently low or the base-circuit resistance sufficiently high, the circuit will oscillate.

The apparent resistance in series with the resonant circuit is negative when inequality of equation 2.32 holds. An idealized static V-I characteristic of
the circuit to the right of points 1 and 2 in Figure 2.10 is shown in Figure 2.11, in which the static operating point has been taken as the origin.

![Figure 2.11. Idealized Volt-Ampere Characteristic.](image)

When the resistance $R$ is taken into account, the combined characteristic appears as in Figure 2.12. If this characteristic has a region of negative slope, the circuit will oscillate at a frequency determined primarily by $L$ and $C$. Figure 2.12 also constitutes an isocline diagram by means of which the current and voltage waveshapes of the circuit may be found. For details of this procedure the reader is referred to the literature (for example, Edson, *Vacuum Tube Oscillators*, John Wiley, 1953). Briefly, the isocline diagram is a graphical, parametric representation of the equation for voltage equilibrium, which reads

\[ F(i) + Ri + L \frac{di}{dt} + \int \frac{idt}{C} = 0 , \tag{2.33} \]
F(i) representing the transistor characteristic. Eliminating t from equation 2.33 gives

\[ \frac{dv}{dt} + \frac{L}{C} v + F(i) + RI = 0. \]  \hspace{1cm} (2.34)

A closed path that satisfies equation 30 is called a limit cycle and represents steady-state oscillations. The limit cycle may be determined from the isocline diagram, after which the desired waveshapes may be found.

Again referring to Figure 2.12, a limit cycle and the corresponding current wave are shown. Figure 2.13 shows the result of increasing the L/C ratio. Notice that the distortion is greatly reduced and that the operating path is more nearly confined to the region of negative slope. The relative frequency stability is not obvious, but a careful analysis shows that the system of Figure 2.12 is more sensitive to small changes of the F(i) characteristic than that of Figure 2.13. Also, it may be seen from equation 2.34 that decreasing R has the same qualitative effect as decreasing L/C. Thus, an increase of Q will result in an improvement or an impairment of frequency stability depending upon how the increase is obtained.

Some experimental data have been taken on laboratory oscillators of the type shown in Figure 2.10, using Western Electric type-1768 point-contact transistors. The influence of harmonic content on frequency stability is apparent from the curves of Figure 2.14. As the collector voltage, \( V_c \), is increased to approximately 5.5 volts, the amplitude of oscillation increases almost linearly. In this range the waveshape is nearly sinusoidal, and the frequency is affected very little by changes in amplitude. For higher values of \( V_c \), the circuit becomes more nonlinear, the harmonic content of the current increases sharply and
Figure 2.12. Isocline Diagram for a Small L/C Ratio for the Circuit in Figure 2.10.

Figure 2.13. Isocline Diagram for a Large L/C Ratio for the Circuit in Figure 2.10.
its effect on frequency is apparent. For still higher values of $V_c$, the amplitude and harmonic content become less sensitive to small changes in $V_c$. The frequency is, therefore, also less sensitive to changes of $V_c$ in this region. Figure 2.15, which shows the variation in frequency as $V_c$ is changed, also shows the relative effects of changing $Q$ and changing the $L/C$ ratio. These results entirely confirm the analysis presented above.

By similar methods, any type of two-terminal oscillator may be analyzed. The only information needed is the input volt-ampere characteristic of the oscillator exclusive of the frequency-determining element. This information, however, is not always readily obtained.

It is possible by use of the isocline diagram to determine the exact frequency of oscillation. Practically, however, the limitations of graphical analysis render this method almost useless for studying frequency variations of the magnitude encountered in a reasonably stable oscillator. Also, the isocline diagram cannot be easily related to quantities that can be observed or measured in the laboratory. When multiloop circuits and high-order differential equations are involved, this method appears formidable indeed. This analysis, by itself, also fails to indicate the necessary circuit modifications to yield improved stability. However, it may, in some cases, be used as a tool for improving an existing design.

D. The Groszkowski Analysis of a Nonlinear Oscillator

In the circuit of Figure 2.1, the condition for sustained oscillation was given by equation 2.5. When this equation is exactly satisfied, current can flow without any signal source since the net impedance around the loop is zero.
Figure 2.14. Amplitude and Frequency as Functions of Collector Voltage.

Figure 2.15. The Effects of Changes in Q and Changes in Inductance on Frequency Variations.
If the system were linear and the condition of equation 2.5 were met, the circuit would not oscillate spontaneously; but once started by some external means, it would generate sustained oscillations of constant amplitude. If some parameter were changed so that the condition of equation 2.5 could not be met at any frequency, the oscillations would either die out altogether or increase in amplitude indefinitely. Such a pinpoint condition would be exceedingly difficult to establish in practice and would be of little use. Practical oscillators are constructed so that a very small disturbance, even thermal noise, starts growing oscillations, but the growth causes the system to change in such a way that an equilibrium is reached for a finite amplitude of oscillations. If some parameter is changed, a new equilibrium results. Since the adjustment to equilibrium is automatic, practical oscillators are amplitude stable. Amplitude stability cannot be achieved without using some nonlinear element or device. Perhaps the simplest situation occurs when the input volt-ampere characteristic is a single-valued function of the current. A typical characteristic of this kind is shown in Figure 2.16, which represents the emitter characteristic for a particular Western Electric type 1768 transistor when used as shown in Figure 2.17. The dynamic input resistance is given by the slope of this characteristic. For small variations of current near an operating point such as 0 in Figure 2.16, the dynamic resistance is negative and oscillations will begin if \( R_1 \), the magnitude of this resistance, is greater than \( R_0 \). Amplitude limiting depends on the fact that the effective value of \( R_1 \) decreases with increasing amplitude of oscillations. An exact analysis of this oscillator is quite involved; however, in practical situations simplifying approximations can be made which permit a satisfactory explanation of its behavior. The most important assumption
Figure 2.16. Typical Negative-Resistance Characteristic.

Figure 2.17. Circuit for Obtaining Negative Resistance.
is that the current in the circuit is nearly sinusoidal. For each value of current there is a corresponding value of the fundamental component of voltage. The ratio of this voltage to the fundamental component of current gives $R_1$, provided these two are in phase. Assuming for the moment that they are in phase, $R_1$ can be measured easily by means of the circuit in Figure 2.18.

![Circuit Diagram](image)

**Figure 2.18. Circuit for Studying Input Impedance.**

To measure $R_1$, a harmonic wave analyzer, WA, and a precision 1000-ohm potentiometer are used in the emitter circuit. The wave analyzer is tuned to the fundamental frequency, and the 1000-ohm potentiometer is adjusted until a null is indicated. Then the total resistance across which the voltage is being measured is zero and the resistance between the potentiometer arm and the emitter has the negative value of the input resistance of the transistor. The fact that almost a perfect null is obtained indicates that the input impedance is practically a pure resistance at the fundamental frequency. Of course, this is true only at low frequencies. Also, the driving current must be nearly a pure sine wave for accurate results to be obtained.

The variation of $R_1$ with current for a particular type 1768 transistor is shown in Figure 2.19 for several values of collector supply voltage. (Emitter
bias current is constant.) These curves may be used to determine the amplitude of oscillations corresponding to a specified value of the tank-circuit resistance, $R_o$. In fact, they can be interpreted as representing the variation of oscillating current as $R_o$ is varied. In this sense, they give only an approximation because the actual oscillating current is not quite sinusoidal.

![Graph](image)

Figure 2.19. Variation of $R_1$ as a Function of Collector Supply Voltage.

However, when the distortion is small, the approximation is acceptable. Amplitude stability results from the fact that if the current is smaller than that corresponding to $R_o$, the net effective resistance of the circuit is negative and the oscillations therefore increase in amplitude, whereas the converse is true if the current is larger than the equilibrium value. Since the variation of $R_1$ is the result of distortion, the amplitude limiting is likewise regarded as a result of distortion.

The operating frequency of this circuit is determined by the condition that the net circuit reactance must be zero. If the current in the circuit is truly
sinusoidal and if the volt-ampere characteristic of $Z_1$ is truly single-valued, analysis shows that the fundamental component of the voltage across $Z_1$ remains in phase with the current in spite of the distortion. This phase relationship still holds when the current wave is distorted, provided the current wave can be expressed as a cosine series. In other words, if the reactance of $Z_o$ is zero not only at the fundamental frequency but also at its harmonics, the operating frequency is not affected by the nonlinear nature of $Z_1$. If, as is usually the case, $Z_o$ has a reactive component at the harmonic frequencies, the voltage across $Z_1$ is not in phase with the current at the fundamental frequency. Under these circumstances, $Z_1$ may be regarded as having a reactive component at the fundamental frequency, and the distortion causes the frequency to be "pulled" away from the natural resonant frequency of the tank circuit.

In terms of the harmonic currents, the condition on the reactances is given by Groszkowski's equation,\footnote{This equation is derived in the appendix.}

$$\sum_{n=1}^{\infty} n I_n^2 x_n = 0, \quad (2.35)$$

in which $n$ is the order of the harmonic current $I_n$, and $x_n$ is the reactance of the passive circuit at this frequency. In terms of the harmonic voltages, equation 2.35 becomes

$$\sum_{n=1}^{\infty} n V_n^2 \frac{x_n}{\| I_n \|^2} = 0, \quad (2.36)$$
in which \(|n|\) is the magnitude of the impedance at the nth harmonic. Equation 2.36 may also be written in terms of susceptance as

\[
\sum_{n=1}^{\infty} nV_n^2 b_n = 0, \quad (2.37)
\]

since \(x_n^2 / |z_n|^2 = -b_n\).

Equation 2.36 may be rearranged to give the following more convenient result.

\[
x_1 = -\sum_{n=2}^{\infty} n \left( \frac{V_n}{V_1} \right)^2 \left| \frac{z_1}{z_n} \right|^2 x_n \quad (2.38)
\]

This is the reactance that \(Z_0\) must have at the operating frequency. As an example, let \(Z_0\) be a simple series circuit \(R_o, L_o, C_o\). Then

\[
Z_0 = R_o + J(\omega L_o - \frac{1}{\omega C_o}) = R_o \left[ 1 + JQ_o \left( \frac{1}{\omega o} - \frac{\omega}{\omega o} \right) \right], \quad (2.39)
\]

where

\[
\omega o^2 = \frac{1}{L_o C_o}, \quad (2.40)
\]

\[
Q_o = \frac{\omega_o L_o}{R_o} \quad (2.41)
\]

and

-38-
\[ x_n = R_0 Q_0 \left( \frac{n\omega}{\omega_0} - \frac{\omega_0}{n\omega_0} \right) = R_0 Q_0 (n \cdot \frac{1}{n}). \]  

(2.42)

Except for very small \( Q_0 \),

\[ z_1 \approx R_0 \]  

(2.43)

and

\[ z_n \approx x_n \text{ for } n \geq 2. \]  

(2.44)

Substituting these results into equation 2.38 gives

\[ X_0 = x_1 = - \sum_{n=2}^{\infty} \frac{V_n}{V_1} \left( \frac{\omega_0 L}{Q_0} \right)^2 \frac{n^2}{n^2 - 1}. \]  

(2.45)

If \( X_0 \) is small, then to a good approximation

\[ X_0 \approx 2\omega_0 L_0 \left( \frac{\omega - \omega_0}{\omega_0} \right) = 2L_0 \Delta \omega \]  

(2.46)

and so equation 2.45 yields

\[ \Delta \omega = \frac{\omega_0}{2Q_0^2} \sum_{n=2}^{\infty} \frac{V_n^2}{V_1} \frac{n^2}{n^2 - 1}. \]  

(2.47)

According to equation 2.47, a circuit operating with 30 per cent second harmonic, 20 per cent third harmonic and 10 per cent fourth harmonic and with a tank circuit \( Q \) of 100 would be "pulled" away from the resonant frequency of the tank by about 0.001 per cent. It should be noted that these figures represent relatively little distortion of the current. The second-harmonic
current, for example, would be only 0.2 per cent of the fundamental. Any change in the circuit external to the tank that tends to increase the amplitude will ordinarily increase the distortion as required by the limiting action, and this in turn will cause the frequency to change. It is clear, therefore, that the frequency stability will be improved if amplitude control can be obtained without depending on distortion.

Unfortunately, practical circuits always contain parasitic reactances that may cause the impedance to harmonics to be much less than is indicated by equation 2.39. The resulting increase in harmonic currents will cause the actual frequency deviation to be greater than is given by equation 2.47. Parasitic reactances may also cause the volt-ampere characteristic of $Z_1$ to be a double-valued function of the current. In this case Groszkowski's equation does not apply, but it seems likely that the effect of distortion would be increased. Thus, distortion may play an important part in oscillator stability even at high frequencies.

One of the problems arising from this type of analysis is the classification of nonlinear characteristics. One method of classification is by power-series representation. A system is then characterized as having second-order curvature, third-order curvature, etc. This procedure is very satisfactory for some purposes, but when the order of curvature is high, the number of variables to be considered is so large that it becomes unwieldy. Power-series representation is most useful for a nearly linear device. It has been found, however, that a transistor oscillator operating on the "linear" part of a negative-resistance characteristic tends to depend on critical adjustments for satisfactory operation. When the operating range includes sharp bends in the characteristic,
circuit adjustments are much less critical, amplitude stability is improved and frequency stability is approximately the same as in the case of nearly linear operation. Consequently, there is a need for an analysis suited to this type operation.

One method of analysis that appears promising is based on a Fourier series representation in which the negative-resistance characteristic is approximated by means of straight-line segments. Before describing this method, it is appropriate to review the approximations and assumptions involved.

First, instead of assuming the oscillation to be nearly linear it is assumed to be nearly sinusoidal. In the case of a series-mode oscillator this assumption implies a nearly sinusoidal current, and in the case of a parallel-mode oscillator it implies a nearly sinusoidal voltage. Second, energy storage in the nonlinear part of the circuit is neglected. (Some possible means for including energy storage have been considered, but a discussion of these will be omitted because at this point it would only complicate matters.) Third, it is assumed that oscillations of constant amplitude occur when (a) the real part of the impedance of the nonlinear circuit equals the negative of the real part of the impedance of the resonator, both calculated at the fundamental frequency, and (b) Groszkowski's equation is satisfied. Finally, it is convenient, but not essential, to assume the resonator to consist of a simple tuned circuit in which case it must have a reasonably high Q.

It has not been pointed out previously that the condition on the real parts of the impedances is an approximation. Actually, the fundamental requirement is that the energy dissipated per cycle in the resonator must equal the energy supplied by the active circuit, that is, by the negative resistance. If the current
were perfectly sinusoidal, harmonic voltages would supply no power to the resonator and no approximation would be involved. Otherwise, it is probable that power supplied at the harmonic frequencies must be taken into account in the criterion for constant amplitude. The existence of a complete treatment of this particular phase of the oscillator problem is not known by this project. For the present purpose, this issue will be avoided by taking refuge in the assumption of nearly sinusoidal currents. Under these circumstances, the effect of the harmonics must be small, whatever its nature.

The procedure for calculating the amplitude and frequency of an oscillator can now be described. The volt-ampere characteristics of the negative resistance and the constants of the tuned circuit are assumed to be known. For convenience, the discussion will be based on series-mode operation.

The volt-ampere characteristic to be considered is shown in Figure 2.20. This may be thought of as an idealized emitter characteristic for a point-contact transistor. The coordinates are incremental current and voltage, and the origin, therefore, represents the quiescent operating point. The dotted line is the load line which represents the alternating-current resistance of the resonator, \( R_0 \). For the purpose of this illustration, the direct currents are disregarded.

As the first step, assume a sinusoidal current having an amplitude corresponding to the intersection of the load line with the emitter characteristic. Call this amplitude \( I_1 \). Next, plot the voltage wave and determine the amplitude of the fundamental component, \( V_1 \), and the magnitude of the input resistance, \( |R_1| = \frac{V_1}{I_1} \). If the load resistance is less than \( R_1 \), make a second calculation using a slightly larger value for \( I_1 \). Continue in this way until a value of \( I_1 \)
is found such that $|R_1| = R_0$, and then make a complete harmonic analysis of the corresponding voltage wave. Since the frequency of the current is assumed to be the resonant frequency of the tank, the impedance can be determined for each harmonic voltage and, consequently, an approximate harmonic analysis of the current wave can be found.

![Emitter Characteristic Diagram](image)

Figure 2.20. Idealized Emitter Characteristic.

Next, the whole procedure is repeated using the distorted current wave instead of the pure sinusoid. If the current distortion is small, as has been assumed, the second voltage wave will differ very little from the first, and it is evident that this process will converge rapidly to a solution. Having found the voltage wave, Groszkowski's equation can be used to find the frequency and the problem is completely solved.

Theoretically, the above procedure can be refined so as to give as accurate a solution as is desired. Also, the actual volt-ampere characteristic can be used, since the straight-line approximation in no way affects the principles involved.
Actually, it is not proposed to solve oscillator problems by the method just described, but the principles it illustrates, together with the assumptions stated earlier, form the basis for an analysis which yields important information about the relation between amplitude limiting and frequency stability. The success of this analysis hinges on two important points, which should be restated at this time in somewhat more general terms. First, only an approximate solution is expected but the approximation will be good enough to predict pulling effects with adequate accuracy. Otherwise, it would be necessary to consider too many small details which would defeat the purpose of the method. Second, it is assumed that the characteristics of the tank circuit, whether or not it is a simple series circuit, are such that the current is very nearly sinusoidal. Under these circumstances, the actual voltage wave is essentially the same as the one obtained from a pure sinusoid. Finally, since the parallel-mode oscillator is the dual of the series-mode oscillator, the results apply to both types when the appropriate exchange of variables is made.

The notation to be used is illustrated in Figure 2.21. Two volt-ampere characteristics to be considered are shown in part (a) and the corresponding voltage waveforms are shown in part (b). It is convenient to plot \(-v\) instead of \(v\) and to let \(\omega_0 t = \theta\) since the algebra is thereby simplified. The solid lines apply to an asymmetrical characteristic, such as the one in Figure 2.20, and the dotted lines show the modifications for a symmetrical characteristic, which represents certain types of push-pull oscillators. The asymmetrical characteristic is designated type I and the symmetrical characteristic is designated type II. \(R_0\) is the resistance corresponding to the linear part of the characteristic. Let

\[ i = K I_o \cos \theta = I_1 \cos \theta. \]
Figure 2.21. (a) Volt-Ampere Characteristics. (b) Voltage Waveforms.

Then, for the type I characteristic,

\[
-v = \begin{cases} 
R_o I_o, & 0 \leq \alpha \leq \theta \\
K R_o I_o, & \alpha \leq \theta \leq \pi 
\end{cases} 
\]  \hspace{1cm} (2.48)

and

\[
-v(-\theta) = -v(\theta). \]  \hspace{1cm} (2.49)

Note that \( K \cos \alpha = 1 \). \hspace{1cm} (2.50)

In terms of its Fourier series,

\[
-v = \frac{a_0}{2} + \sum_{n=1}^{\infty} a_n \cos n\theta 
\]  \hspace{1cm} (2.51)
with
\[ a_n = \frac{1}{\pi} \int_0^{2\pi} (-v) \cos n\theta \, d\theta. \]  \hspace{1cm} (2.52)

Substituting for \(-v\) from equation 2.48 and taking advantage of the fact that \(-v\) is an even function of \(\theta\) gives
\[ a_n = \frac{2R_0 I_0}{\pi} \left[ \int_0^\alpha \cos n\theta \, d\theta + \int_\alpha^\pi K \cos \theta \cos n\theta \, d\theta \right]. \quad (2.53) \]

Equation 2.53 yields:
\[ a_1 = KR_0 I_0 \left[ 1 + \frac{\sin 2\alpha - 2\alpha}{2\pi} \right] = -v_1 \]  \hspace{1cm} (2.54)

\[ a_n = \frac{KR_0 I_0}{\pi n} \left[ \frac{\sin (n+1)\alpha}{n+1} - \frac{\sin (n-1)\alpha}{n-1} \right] = -v_n, \quad n = 2, 3, 4... \quad (2.55) \]

For the type II characteristic, a similar analysis gives:
\[ a_1 = KR_0 I_0 \left[ 1 + \frac{\sin 2\alpha - 2\alpha}{\pi} \right] \]  \hspace{1cm} (2.56)

and
\[ a_n = \frac{2KR_0 I_0}{\pi n} \left[ \frac{\sin (n+1)\alpha}{n+1} - \frac{\sin (n-1)\alpha}{n-1} \right] \quad n = 3, 5, 7... \quad (2.57) \]

A useful interpretation of these results requires a knowledge of the manner in which the volt-ampere characteristics change under the influence of oscillator disturbances. The ways in which a type I characteristic may vary are illustrated in Figure 2.22. A change from \(a\) to \(b\) changes \(I_o\) but \(K\) will not change if \(R_1\) remains fixed. In other words, this change will affect only the amplitude of the oscillations. A change from \(a\) to \(c\) changes \(R_o\) and so, in
Figure 2.22. Variations of a Type I Characteristic.

In general, would affect both amplitude and distortion. Reference to equations 2.54 and 2.55 shows that the relative distortion depends on $\alpha$ and, hence, on $K$ but not on $I_o$. Therefore, a change from $a$ to $c$ is equivalent to a change in the external resistance except for a difference in amplitude. Likewise, a change from $a$ to $d$ is equivalent to a change of the external resistance combined with a change of amplitude. Therefore, the stability of an oscillator with respect to changes in its negative-resistance characteristic is essentially the same as its stability with respect to a change of the external (linear) circuit resistance.

Equation 2.54 yields

$$\frac{-R_1}{R_o} = \left[ 1 + \frac{\sin 2\alpha - 2\alpha}{2\pi} \right]$$

(2.58)

for type I and equation 2.56 gives

$$\frac{-R_1}{R_o} = \left[ 1 + \frac{\sin 2\alpha - 2\alpha}{\pi} \right]$$

(2.59)
for type II. These results are plotted in Figure 2.23 as functions of \( K \). The frequency shift caused by distortion depends also on the properties of the resonator. In the case of a single tuned circuit, the frequency is given by equation 2.47 which may be written as

\[
f = f_0 \left[ 1 - \frac{1}{2Q_0^2} \sum_{n=2}^{\infty} \left( \frac{v_n}{v_1} \right)^2 \frac{n^2}{n^2 - 1} \right].
\]  

(2.60)

Hence, the quantity inside the summation sign is seen to be a kind of "pulling" factor for the oscillator. Call this quantity \( P^2 \). Then equation 2.60 becomes

\[
f = f_0 \left[ 1 - \frac{P^2}{2Q_0^2} \right].
\]  

(2.61)

The pulling factors for the type I and type II oscillators are also plotted in Figure 2.23. Harmonics only through the eighth have been included in calculating \( P^2 \). Neglecting the higher harmonics introduces appreciable errors for \( K > 6 \).

The use of these performance curves may be illustrated by means of a few numerical examples.

1. A series-mode oscillator having a type I characteristic has a negative resistance of 1000 ohms over its linear part. The "break-point" \( I_0 \) corresponds to an incremental current of 0.5 mA. The tank circuit has a resistance of 800 ohms, has a \( Q \) of 10 and is resonant at 1000 cps. Find the amplitude and frequency of the oscillations.
Figure 2.23. Normalized Performance Curves for Type I and Type II Oscillators.
Solution:

\[-R_1/R_o = 0.8 \text{ and Figure 2.23 gives } K = 2, P^2 = 0.05.\]

Therefore,

\[V_1 = 2 \times 0.5 \times 10^{-3} \times 800 = 0.8 \text{ volt (peak)}\]

and

\[f = 1000 \left(1 - \frac{0.05}{2 \times 100}\right) = 1000 \left(1 - 0.00025\right) = 999.75 \text{ cps.}\]

2. A ten per cent increase in the supply voltage of the oscillator in example 1 changes its parameters to \[-R_o = 1200 \text{ ohms and } I_o = 0.5 \text{ ma.} \] Find the new amplitude and frequency.

Solution:

\[-R_1/R_o = \frac{800}{1200} = 0.666. \text{ Figure 2.23 gives } K = 3.8 \text{ and } P^2 = 0.12.\]

Therefore,

\[V_1 = 3.8 \times 0.5 \times 10^{-3} \times 800 = 1.52 \text{ volts (peak)}\]

and

\[f = 1000 \left(1 - \frac{0.12}{2 \times 100}\right) = 1000 \left(1 - 0.0006\right) = 999.4 \text{ cps.}\]

3. If the oscillator in example 1 has no type II characteristics find the amplitude and frequency.

Solution:

As before,

\[-R_1/R_o = 0.8, \text{ but now } K = 1.45 \text{ and } P^2 = 0.023.\]

Therefore,

\[V_1 = 1.45 \times 0.5 \times 10^{-3} \times 800 = 0.58 \text{ volts (peak)}\]
and

\[ f = 1000 \left(1 - \frac{0.023}{2 \times 100}\right) = 1000 \left(1 - 0.00012\right) = 999.88 \text{ cps}. \]

4. Changing the supply voltage for this oscillator as in example 2 gives the following results:

\[ -\frac{R_1}{R_0} = 0.666, \quad K = 1.8 \quad \text{and} \quad P^2 = 0.05. \]

Therefore,

\[ V_1 = 1.8 \times 0.5 \times 10^{-3} \times 800 = 0.72 \text{ volts (peak)} \]

\[ f = 1000 \left(1 - \frac{0.05}{2 \times 100}\right) = 1000 \left(1 - 0.00025\right) = 999.75 \text{ cps}. \]

These calculations indicate that the symmetrical oscillator has an amplitude stability about 40 per cent better than the asymmetrical circuit and a frequency stability nearly three times as good.

The characteristics of Figure 2.21 may be modified to obtain more general results as shown in Figure 2.24. The asymmetrical case is still called a type I characteristic, and the symmetrical case is called type II.

In Figure 2.24 (a) the part of the volt-ampere characteristic that lies outside the negative-resistance region represents a linear positive resistance of value \(MR_o\), \(R_o\) being the magnitude of the negative resistance. A negative value of \(M\) corresponds to a resistance that is negative over the whole range but has different values in the two regions. The characteristic in Figure 2.21 (a) is seen to correspond to \(M = 0\).

A family of performance curves for a type I oscillator is shown in Figure 2.25, and a family for a type II oscillator is shown in Figure 2.26. It should
Figure 2.24. (a) Volt-Ampere Characteristics. (b) Voltage Waveforms. (M = 0).

be remembered that the pulling-factor curves depend on the tank-circuit configuration as well as on the distortion and that the curves shown here apply to a simple tuned circuit.

There are so many variables to consider in the design and operation of an oscillator that it is difficult to determine the real significance of these curves. They do not point clearly to any particular collection of "do's" and "don'ts" for oscillator designers. On the other hand, they certainly do show the way in which amplitude stability and frequency stability are related in negative-resistance oscillators, and a few observations in this connection are appropriate.

The curves of Figures 2.25 and 2.26 may be used in the following way. Variations of transistor characteristics are equivalent to variations of $-R_i/R_o$ and $I_o$. If both these quantities vary appreciably (as the result of a variation of temperature, supply voltage or other factors), the variation of $I_o$ must be
Figure 2.25. Normalized Performance Curves for a Type I Oscillator.
Figure 2.26. Normalized Performance Curves for a Type II Oscillator.
taken into account in determining amplitude variations. On the other hand, the pulling factor, \( P^2 \), depends on \( K \) but not \( I_0 \). Hence, frequency stability can be investigated without regard to variations of \( I_0 \). These factors are considered in detail in the appendix where a specific design problem is discussed. For the present purpose it will be convenient to assume that variations of \( I_0 \) are comparatively small and, hence, negligible.

Suppose, now, that various values of \(-R_1/R_0\) are encountered. For any selected value the corresponding value of \( K \) is found from the curves, and this \( K \) value yields the corresponding \( P^2 \) value. In this way, it is found that for small \( M \), \( P^2 \) varies slowly for values of \(-R_1/R_0\) near 0.7, but \( K \) varies rapidly. For large \( M \), the converse is true. Thus, the large \( M \)-values lead to relatively good amplitude stability and relatively poor frequency stability whereas the small \( M \)-values lead to relatively good frequency stability and relatively poor amplitude stability.

Another important conclusion that can be drawn is that \(-R_1/R_0\) should be as close to unity as possible. This requirement is offset by the fact that small variations of \(-R_0\) may cause the oscillations to die if a value too close to unity is used for design center. In this case, poor amplitude stability may result from excessive variations of \(-R_1/R_0\). A value between 0.7 and 0.8 appears to be a good compromise for general-purpose design. In every case the best value can be found only by examining the actual range of variation of parameters caused by the disturbing environment.

These conclusions are more clearly evident from the curves in Figures 2.27 and 2.28. In these curves, \( P^2 \) is plotted as a function of \( K \). Two families of curves are given in each case: one family with \( M \) as a parameter and the other
Figure 2.27. Normalized Performance Curves for a Type I Oscillator.
Figure 2.28. Normalized Performance Curves for a Type II Oscillator.
for $-R_1/R_0$ as a parameter. Variations of $-R_1/R_0$ correspond to motion along the $M = \text{constant}$ lines. Since this form of presentation of the data is probably more useful than that of Figures 2.25 and 2.26, the parts of the curves for $K$ values between one and two are shown to an expanded scale in Figures 2.29 and 2.30. All of these curves may be used for both series-mode and parallel-mode oscillators. For the latter, $-R_1/R_0$ is replaced by $-G_1/G_0$ and $K$ represents the voltage amplitude ratio instead of the current amplitude ratio.

The next question that must be considered is that of the actual shapes of volt-ampere characteristics of transistors and their relation to circuit design. Actual characteristics depend not only on the transistors themselves but also on the external circuits other than the tank circuit. A survey of the possibilities has been made for a particular type of transistor, and it is probably permissible to assume these results to be typical. At first, it was hoped that this information could be obtained from the published curves. Unfortunately, the data in the critical regions are given with very poor accuracy and, therefore, it was necessary to make point-by-point measurements. Figures 2.31, 2.32 and 2.33 show the general effects of various circuit parameters. It was desired to find characteristics that could be approximated with reasonable accuracy by the straight-line segments so that the theory already developed could be tested on a design problem. This problem is presented in the appendix.

The results are considered to be very satisfactory. In spite of the many approximations that must be made, the amplitude of the oscillations was correctly predicted to within a factor of two, and the frequency stability with respect to supply voltage was predicted to within a factor of about three. It is probable that a more exact analysis of the available data would yield even
Figure 2.29. Normalized Performance Curves for a Type I Oscillator (Expanded).

-0.5 -0.4 -0.3 -0.2 -0.1 0 0.1 0.2 0.3 0.4 0.5

- \frac{R_1}{R_0} = 0.7

- \frac{R_1}{R_0} = 0.6

- \frac{R_1}{R_0} = 0.8

- \frac{R_1}{R_0} = 0.9

- \frac{R_1}{R_0} = 1.0

M = 0

M = 0.2

M = 0.4

M = 0.6

M = 0.8

M = 1.0

M = 1.4

M = 2.0

K

p2
Figure 2.30. Normalized Performance Curves for a Type II Oscillator (Expanded).
Figure 2.31. Effect of Collector Voltage on Input Characteristic.
Figure 2.32. Effect of Emitter Resistance on Input Characteristic.
Figure 2.33. Effect of Collector Resistance on Input Characteristic.
better results. This was not attempted for two reasons: first, the order of magnitude of the frequency variation (50 parts per million) was such that it would be necessary to verify the linearity of the tuning capacitor; and second, it was desirable to explore the problem at higher frequencies before considering any further refinements.

Tests of this design method at higher frequencies indicated that factors other than distortion have appreciable effect on frequency stability. (Higher frequency as used here is relative to alpha cut-off.) Accurate predictions of frequency stability and amplitude of oscillation have been possible only as long as the frequency of operation has been less than one per cent of the alpha cut-off. Above this frequency, it is necessary that such factors as transistor-capacitance variations and transit-time variations be considered. It is probable that additional study would result in methods of analysis to include these factors and possibly indicate a method of neutralizing the effects.

E. A General Design Procedure for Linear Oscillators

As pointed out in section II-B, one of the major difficulties of the linear analysis approach is to determine whether or not the assumptions are sufficiently accurate to give satisfactory results. Also, the equations are too complicated for most practical applications. The following procedure, to some extent, solves these problems by providing a design with a minimum of distortion while, at the same time, expresses the results in simple equations that can be used for practical oscillator design. It is not a theory in itself but rather a design procedure which leads to the development of stable oscillators. As will be seen, this procedure is applicable only to the feedback type of oscillator, whereas the Groszkowski analysis presented above is applicable only to the negative-resistance oscillator in its present form.
For a low-distortion oscillator, two primary factors affect the ultimate stability. The first is the phase shift in the amplifier portion of the oscillator, particularly the change in phase shift when external parameters such as voltage and temperature are changed. The second factor is the change in frequency required to cause an equal and opposite change in phase shift in the frequency-controlling feedback loop. The second factor must be considered because the total loop phase shift must be zero.

These two factors are easily separated in oscillators having only external feedback loops. This separation cannot be made as easily for the negative-resistance oscillators. In some circuits, such as the Meacham bridge oscillator, which use both positive and negative feedback, the negative feedback can be considered as either a part of the amplifier or as a part of the frequency-controlling feedback loop. If the negative feedback is considered to be part of the amplifier, it serves to stabilize the phase shift of the amplifier. If it is considered to be a part of the frequency-controlling loop, it serves to reduce the required phase shift in the frequency-controlling element for a given phase shift in the amplifier. Either analysis leads to the same result.

The block diagram for a feedback oscillator is shown in Figure 2.34. The amplifier may be converted to an equivalent Thevenin voltage generator with a voltage, $K_{e1}/\theta$, in series with a resistance, $r_o$, where $r_o$ is to be considered as a part of the feedback network. The input impedance of the amplifier is considered to be infinite since any reactive component may be canceled at any one frequency and voltage, and the resistive component may be included as a part of $R$. The phase angle, $\theta$, is assumed to be a function of all external parameters such as temperature and voltage. It will be assumed, however, that
$\theta$ has been made zero for the normal static condition of the external parameters. $K$ will be assumed to be a real, positive constant so that no distortion of the waveform can occur. Since the frequency-controlling network is passive, the oscillator output will be a pure sinusoidal waveform. In an actual oscillator some distortion must occur in order that amplitude limiting will result. However, by carefully selecting the gain of the feedback network, $k$, this distortion can be kept small so that the above-mentioned assumptions will be approximately correct.

The primary factors affecting stability now become $\phi(f)$ and $\psi(s)$ where $s$ is a general external parameter affecting the amplifier phase shift. The objective is to make $d\phi / ds$ as small as possible and $d\psi / df$ as large as possible.

Several methods are available for reducing the changes in transistor amplifier phase shift for changes in voltage and temperature. One method is to operate the transistor well below its alpha cut-off frequency. Internal positive and/or negative feedback may also be considered as a possible means of improving the amplifier characteristic. However, for this analysis, the use of
over-all negative feedback will be considered as a part of the frequency-controlling feedback loop rather than a part of the amplifier. Clipping or other amplitude control means may in some cases be used to improve the amplifier phase-shift characteristics, provided the over-all distortion is kept low in order that the analysis which is to follow may apply in its simplest form.

The factors influencing $\varphi(f)$ are the geometry and element values of the frequency-controlling network. The choice of network geometry is limited only by practical values of $K$ and $r_o$. For most transistors, $r_o$ is much too large for use with practical networks. Therefore, the use of a transformer is usually necessary. If the values of $K$ and $r_o$ for the transistors alone are compared with the values for the transistor with a transformer, the quantity $K^2/r_o$ will be found to remain constant. This is shown in Figure 2.35 where $K'$ and $r_o'$ are the characteristics of the transistor without the transformer.

Either $K$ or $r_o$ may be selected for best performance with any particular feedback network, however, $K^2/r_o$ must be held constant. For a practical transformer-coupled amplifier, it is necessary that any transformer loss and phase shift be included in the characteristics, $K/\theta$ and $r_o$. This may be accomplished by measuring the amplifier characteristics with a representative transformer in use. The amplifier and transformer combination may then be completely characterized in terms of $\theta$ and $K^2/r_o = A$ = four times the available power output for one-volt input.
To illustrate the method by which a practical oscillator may be designed, the network shown in Figure 2.36 is analyzed as follows:

Let \( jX = jX_1 - jX_2 \). Then,

\[
e_1 = e_2 \frac{R}{R + R_1 + jX + r_0}
\]

or

\[
\frac{e_1}{e_2} = \frac{k\theta}{R} = k \tan^{-1} \frac{X}{R + r_0 + R_1}.
\]

Near series resonance, \( d\phi/df \) increases as \( (R + r_0) \) decreases. The minimum value of \( (R + r_0) \) that will permit oscillations may be found as follows:

For \( X = 0 \),

\[
e_1 = \frac{Re_2}{r_0 + R_1 + R} = \frac{Re_1K}{r_0 + R_1 + R},
\]

but
\[
\frac{K^2}{r_0} = A \text{ or } K = r_0^{1/2} A^{1/2}.
\]

Therefore,

\[
R r_0^{1/2} A^{1/2} = r_0 + R_1 + R
\]
or

\[
R = \frac{r_0 + R_1}{r_0^{1/2} A^{1/2} - 1}
\]  \hfill (2.64)

from which

\[
R + r_0 = \frac{r_0 + R_1}{r_0^{1/2} A^{1/2} - 1} + r_0
\]

\[(R + r_0) \text{ is minimum when } \frac{d(R + r_0)}{dr_0} = 0\]
or when

\[
2r_0^{3/2} A^{1/2} - 3r_0 = R_1 = 0.
\]  \hfill (2.65)

By substituting \(R_1\) as determined from the resonant element chosen for the oscillator and \(A\) as determined by the amplifier, \(r_0\) can be found from equation 2.65. Figure 2.37 shows the variation of \(r_0\) with \(A\) as a parameter with \(R_1\) constant at 100 ohms. For each value of \(r_0\) and \(A\), \(K\) is determined from \(K^2/r_0 = A\). \(R\) is found from equation 2.64. The variations of \(K\) and \(R\) for \(R_1 = 100\) ohms are also shown in Figure 2.37.
If $r_o$ and $R$ could both be made zero and still maintain oscillations, the change in frequency for a small amplifier phase shift, $\Delta \theta$, would be:

$$\Delta f = \frac{(\Delta \theta) (f_o)}{2Q}, \text{ for } \Delta \theta \text{ small}$$

(2.66)

where: $Q = X_1/R_1$ and $f_o = \frac{1}{\sqrt{LC}}$ for the resonant element. For values of $r_o$ and $R$ other than zero, equation 2.66 must be corrected by making $Q = X_1/(R_1 + R + r_o)$.

The relative change in frequency for a practical circuit compared to the change for an unloaded resonant circuit for the same $\Delta \theta$ is

$$\rho \phi = \frac{R_1}{R + r_o + R_1}.$$  

(2.67)

Thus, $\rho \phi$ is a figure of merit for the network. If the values of $\rho \phi$ are compared for two different networks, then the relative stabilities of oscillators using the two networks can be determined since:

$$\frac{\left( \frac{S_v}{v} \right)_1}{\left( \frac{S_v}{v} \right)_2} = \frac{\left( \rho \phi \right)_1}{\left( \rho \phi \right)_2}.$$  

(2.68)
Figure 2.37. Design Curves for Oscillator of Figure 2.36.
for the same amplifier and resonant element. Equation 2.67 expresses \( p_0 \) only for the network of Figure 2.36. For other networks, an appropriate expression for \( p_0 \) would have to be found.

The \( p_0 \) of the circuit of Figure 2.36 is plotted in Figure 2.37 \((R_1 = 100 \text{ ohms})\).

To design a practical circuit to use the network of Figure 2.36, a suitable amplifier is first chosen. \( A \) is found experimentally. A suitable resonant element is chosen (crystal, series LC circuit, or other). \( R_1 \) is determined from the resonant element. The correct value of \( r_0 \) is then found from equation 2.65. \( R \) can then be found from equation 2.64. \( X \) as a function of frequency can be calculated for the resonant element. Then, \( \phi \) as a function of frequency can be found from equation 2.63. If the amplifier phase shift, \( \theta \), is known as a function of voltage (or temperature, etc.), then the frequency of oscillation as a function of voltage can be found by making \( \phi \) equal to \( \theta \) for all voltages.

This method yields a theoretical stability which may be slightly better than can be obtained in a practical circuit since distortion has been neglected and since \( R \) cannot be made as small as the value calculated in equation 2.64. If \( R \) is made exactly equal to the minimum value, when a phase shift, \( \theta \), occurs, \( R \) will no longer be large enough to permit sustained oscillations.

For some transistors, making \( R \) smaller than the value calculated in equation 2.64 will still produce oscillations but at a reduced amplitude. This is because the nonlinear characteristic of the transistor is often such that the gain, \( K' \), is much greater for low amplitudes. From one standpoint, this is undesirable since it implies that the transistor transfer characteristic will produce appreciable distortion. However, it does provide better amplitude limiting.
Closer examination of equation 2.63 indicates that the amount of improvement to be expected from the proper selection of R may be small for a practical oscillator using this network configuration. This is particularly true when the input impedance of the amplifier is low since the maximum value of R that can physically exist is the actual input resistance of the amplifier. Thus, the improvement in stability obtained by externally unloading the crystal by R would be found by first substituting the amplifier input resistance into equation 2.63 and calculating the stability and then substituting R (minimum) into equation 2.63 and again calculating the stability. The ratio of the two stabilities will be the improvement due to the proper selection of R.

More important than the correctness of the magnitude of the stability as given by this method is the comparison that can be made between networks of different configurations. This method is also very valuable as a tool in choosing the constants for any particular feedback network. For example, in the circuit of Figure 2.36, one might assume that the value of R was relatively unimportant. However, as shown previously and substantiated later by experimental results, this assumption is incorrect.

Another network of interest is shown in Figure 2.38. The equations for this network are as follows:

\[ K = 1 + \sqrt{1 + \frac{R_1 A}{K}} / 2, \]  
\[ R = \frac{4K^2 + R_1 A(K + 1)}{A(K - 1)} \]  
\[ \rho_s = \frac{200(R + 2r_0)}{(R - R_1)(R + R_1 + 4r_0) - 1} \]
Figure 2.38. Negative-Feedback Network.

Figure 2.39 shows the variations of R, K, r₀ and ρ₀ for values of A from 0.1 to 10.0. R₁ has again been chosen as 100 ohms for illustrative purposes. The curves of Figure 2.39 are used the same way as the curves of Figure 2.37. Likewise, the application of the equations is the same.

The importance of R₁ is not directly indicated by the equations for either network. In general, for these networks and other similar networks, the frequency stability is increased as R₁ is decreased. For example, using equations 2.69, 2.70 and 2.71 and a typical value of A of 2, the following results are obtained:

For R₁ = 1000 ohms, ρ₀ = 0.795,
For R₁ = 100 ohms, ρ₀ = 2.5 and
For R₁ = 10 ohms, ρ₀ = 7.1.

If a crystal is used as the series-resonant element in Figure 2.38, the maximum stability obtainable is almost ten times as great for a series-resonant impedance of 10 ohms as it is for 1000 ohms, assuming that the equivalent L and
Figure 2.39. Design Curves for Oscillator of Figure 2.38.
C do not change. This is as would be expected since a decrease in the series-resonant impedance indicates an increase in Q.

It should be observed that for the network of Figure 2.36, $p_0$ can never exceed unity. That is, no Q multiplication can occur. However, for the network of Figure 2.38, $p_0$ is usually greater than unity for appreciable amplifier gains. Thus, the network of Figure 2.38 should provide greater stability than the network of Figure 2.36 under similar conditions.

The validity of the above equations is illustrated by the design examples given in section VII-C.

It should be remembered that the two networks, mentioned above for illustrative purposes, are not necessarily the best networks that can be found for oscillator design purposes. It is possible that networks may be developed which will yield greatly improved stabilities.

F. The Empirical Approach

A part of the work of this project has been directed toward developing new oscillator circuits and new variations of basic circuits. It has been found that practically all circuits have a vacuum-tube equivalent. In particular it has been found that most circuits are variations of the familiar Hartley, Colpitts or negative-resistance circuits. In one approach, the vacuum-tube equivalent has been purposely ignored in hope that a better understanding of the transistor circuit would result. Many circuits have, therefore, been investigated without regard to any theory. This has in some cases resulted in improved stabilities, while in other cases it has led to the conclusion that certain circuits are not well adapted to use with transistors. In general, the results have been similar to the predictions of the available theories.
The empirical approach has led to a number of conclusions concerning transistor oscillators. These conclusions, which concern bias stabilization, component selection and application, temperature compensation and stabilization by distortion, have been reserved for section VI of this report. Such conclusions are referred to as empirical for one of two reasons: (1) they have not been explained satisfactorily by theory or (2) they are of such elementary nature that theoretical explanation is not necessary.

A number of other empirical conclusions have been reached concerning specific oscillators. Each of these will be discussed in the appropriate place in section VII.

Some of the specific types of circuits which have resulted from the empirical approach are: (1) the junction-transistor oscillator using zero base current, (2) oscillator circuits using diode clipping, (3) distortion-compensated circuits, (4) transformer neutralized circuits and (5) combination feedback and negative-resistance circuits.

G. Vacuum-Tube Analogies

Many of the oscillators which have been constructed have been patterned after their vacuum-tube equivalents. Notable among these are the Hartley, Colpitts, Pierce and Miller circuits. The vacuum-tube theories and analysis have been applied to most of these circuits; however, in most cases, satisfactory results have not been obtained. Several reasons are apparent. First, the vacuum tube is essentially a device which gives power output for a voltage input. In other words, the power gain may be considered infinite for most vacuum-tube analysis, whereas the transistor is definitely limited as to power gain. Second, the nature of the limiting action is different for the vacuum
tube and the transistor. Third, the transistors generally have been operated nearer their frequency cut-off limit than have the vacuum tubes. For these and other reasons, the stabilization procedures suggested by Llewellyn, Jefferson and others do not always give satisfactory results with transistors. In some cases, satisfactory results have been obtained at frequencies which are very low compared to the alpha cut-off frequency of the transistor, that is, at frequencies where the transistor parameters are nonreactive.

On the other hand, vacuum-tube analogies have been useful in that they suggest possible transistor oscillator designs. In some few cases, the vacuum-tube theories have been directly applicable to the design of stable transistor oscillators. An example of this is the Meacham bridge oscillator circuit.

Many negative-feedback and AGC arrangements have been tried with transistor oscillators. These have failed in most cases because of the limited power gain.

A number of problems become apparent when a transistor is substituted for a vacuum tube. One such problem is that a quartz crystal does not operate satisfactorily when used in its parallel mode with transistors. This immediately indicates that some vacuum-tube circuits such as the Miller and parallel-mode dynatron may not perform satisfactorily with transistors. Such has been found to be true in actual laboratory experiments. In class-A vacuum-tube oscillator circuits, the grid resistor may usually be made very large if necessary to improve stability; however, with transistors the base resistor is limited in size by the base current required. As mentioned in section II-F, several transistor oscillators have been designed so that the average base current is zero. These were patterned after the vacuum-tube circuits.
It has been concluded that vacuum-tube analogies are, in general, useful as an initial guide but not particularly useful in final stabilization.

H. Relative Advantages of the Design Procedures

The theories and design procedures discussed in the preceding sections are useful in practical oscillator design to varying degrees. In this section all transistor oscillators will be divided into two categories (1) negative-resistance oscillators and (2) feedback oscillators.

1. Negative-Resistance Oscillators

Five of the previously discussed procedures are applicable to negative-resistance oscillators. All five procedures have limitations of various kinds. The linear circuit approach is the most comprehensive, but also the most difficult to apply. In addition, the approximations involved are likely to result in serious errors. The isocline diagram method is capable of good accuracy but, as are most graphical methods, is very time consuming. Also, it does not directly indicate methods of improving stability. The empirical approach and the vacuum-tube analogies are not in themselves design procedures but are sometimes useful in conjunction with other methods. The Groszkowski analysis has proved useful in designing stable oscillators but is greatly limited as to accuracy at high frequencies. It has the advantage of directly leading to practical oscillator designs while at the same time indicating the necessary changes to improve the frequency stability. The analysis as presented in this report is not complete because of the termination of the project. However, indications are that the Groszkowski analysis can be extended to higher frequencies by taking into account the nonlinear capacitance variations and transit-time variations of transistors.
Even in its present form, the Groszkowski procedure is recommended as the best available method of designing negative-resistance oscillators.

2. Feedback Oscillators

Three of the previously discussed procedures are applicable to the design of feedback oscillators. These are the general design procedure for linear oscillators, the empirical approach and the vacuum-tube analogies. The linear-circuit approach and the isocline diagram are both adaptable to feedback-oscillator design but have not been considered because of their complexity and probable limited usefulness. The Groszkowski analysis promises to be useful in feedback-oscillator design, but time has not permitted this aspect to be studied.

The empirical approach and the vacuum-tube analogies are again very useful although, as mentioned previously, they are not in themselves design procedures. The most successful method of designing feedback oscillators has been the general design procedure for linear oscillators. This has resulted in the best stabilities thus far obtained (with the exception of the Meacham oscillator) for oscillators without compensation. Compensation applied to these oscillators would, of course, result in further improvements in frequency stabilities.

The general design procedure for linear oscillators is recommended as the best available method of designing feedback oscillators.
III. PROPERTIES OF RESONATORS

A. Quartz Crystals, Series and Parallel Modes

An important part of this work has been the application of crystal control to transistor oscillators. In order to approach this problem intelligently, an understanding of the equivalent circuit of quartz crystals is necessary.

The most commonly accepted and simplest equivalent circuit of a quartz crystal is shown in Figure 3.1. As can be seen from this figure,

![Figure 3.1. Equivalent Circuit of a Quartz Crystal.](image)

the crystal exhibits both a series resonance (resonance) and a parallel resonance (antiresonance). The crystal has the property that $L_1$ is very large (0.1 to 100 henries) and $C_1$ very small (0.001 to 0.1 micromicrofarad). The series arm resistance, $R_1$, usually has a value between 10 ohms and 5000 ohms, depending on the type of crystal and the frequency. The total shunt capacitance, $C_0$, is usually between 3 and 30 micromicrofarads.

Some of the more important relations for quartz crystals are as follows:
\[ \omega_0 = \frac{1}{\sqrt{L_1 C_1}} = \text{resonant frequency of the series arm.} \quad (3.1) \]

\[ Y_o = \frac{1}{R_1} + j \omega_0 C_0 = \text{admittance when the series arm is resonant.} \quad (3.2) \]

\[ \omega_r \approx \sqrt{\frac{1}{L_1 C_1} + \frac{R_1^2 C_0}{L_1^2 C_1}} = \text{series-resonant frequency.} \quad (3.3) \]

\[ \omega_a \approx \omega_0 \left(1 + \frac{C_0}{2C_1}\right) = \text{parallel-resonant frequency.} \quad (3.4) \]

\[ PI = \frac{\omega_0 L_1 - \frac{1}{\omega_r C_1}}{\omega_0^2 R_1} = \text{performance index = parallel-resonant resistance of the crystal and total shunt capacitance.} \quad (3.5) \]

\[ Q \approx \frac{\omega_0 L_1}{R_1} = Q \text{ at series resonance.} \quad (3.6) \]

Since these relations are readily available in the literature, they will not be derived here.

The primary reason for using a crystal in an oscillator circuit is to obtain a large change in phase for a small change in frequency. The ability of a crystal to do this is indicated by the fact that the crystal Q is very large at series resonance. A study of the above equations will show that in order to maintain this high Q it is necessary that the crystal operate into a low impedance when used in its series mode or operate into a high impedance when used in its parallel mode. In addition, it can be seen that \( C_0 \) tends to decrease the effective Q of the series resonance, though not to a large extent. The Q of the crystal is the greatest when it is operated at the frequency \( \omega_0 \).
In most series-mode transistor oscillators, it is possible to operate the crystal near $\omega_0$ and into a low impedance, thus maintaining the high $Q$. $C_o$ is usually increased very little by the transistor circuit.

When a crystal is operated in its antiresonant mode, any shunt resistance will reduce the effective $Q$. Since $PI$ is very high for a crystal, it is difficult to design a transistor oscillator circuit to make effective use of the parallel mode. This is because the transistor is inherently a low-impedance device. In addition, most circuit arrangements which require a parallel-resonant circuit also require a d-c path for bias currents. The crystal must, therefore, be shunted by a d-c path to provide the bias.

Crystals may also be operated at frequencies other than the series- or parallel-resonant frequencies. This ordinarily results in a reduction of $Q$.

An attempt has been made to correlate crystal characteristics with transistor oscillator performance. This has been a difficult process because most of the crystals available to the project have been experimental types or unidentified types. Crystal impedance meters have been available only on a very limited basis. However, certain conclusions concerning the use of crystals in transistor oscillator circuits may be stated.

1. Crystals do not ordinarily operate satisfactorily in their parallel mode in transistor oscillators.

2. Where parallel resonances are required, impedance-inverting networks may be used with a series-mode crystal (section III-E).

3. A crystal may be operated as an inductance; however, some reduction in $Q$ can be expected to result.

4. The crystal should be operated as near $\omega_0$ as possible when used in its series mode.
5. Crystals should operate into as low an impedance as possible when used in the series mode.

6. Precautions are sometimes necessary to prevent crystals from operating in overtone modes in transistor oscillators.

7. Variations in $R_1$ with temperature often cause larger frequency changes than changes in $\omega_0$ with temperature since most transistor oscillators are very dependent on the resonator impedance.

8. Crystal units may be less stable at elevated temperatures (even when used in an oven) than they are at uncontrolled room temperature.

9. The value of $R_1$ is often very dependent on drive level. This is particularly true at the very low power levels encountered in transistor oscillators.

10. The power-level rating of practically all crystals is sufficiently large for use with most transistor oscillators.

11. In some oscillators, the over-all stability is limited by the crystal's temperature stability rather than by the transistor circuit. Preferably, the stability characteristics of a crystal should be known before attempting to incorporate the crystal in a transistor oscillator.

12. Measurements made on a crystal impedance meter cannot always be relied upon when designing crystal-controlled oscillators because of the difference in the minimum impedance-meter drive level and the drive levels encountered in most transistor oscillators.

Many excellent references are available on crystal resonators in the lower frequency range. Use of these references has been made throughout the performance of the project.
B. Lumped Resonant Elements

Many of the oscillator circuits which have been investigated have used lumped inductance and capacitance as the frequency-controlling elements. Such resonant circuits are not nearly as simple as they first appear to be. Some of the problems which have been encountered are as follows:

1. The effective high-frequency Q of an inductance is more often limited by the distributed capacitance than by the high-frequency resistance.

2. The change in Q of an inductance may in some cases be more important than changes in inductance values.

3. The Q of LC circuits is often limited by capacitor losses rather than by the Q of the inductance.

4. Inductances containing magnetic cores often are sufficiently nonlinear to greatly limit the stability.

5. Spurious resonances are very difficult to eliminate and may be very troublesome in transistor oscillators, especially the negative-resistance type of oscillator.

6. The resonant impedance of LC circuits is very important in transistor oscillator design.

7. High Q is usually desirable although it is not always the most important factor in LC resonator design.

Temperature compensation of LC resonators is usually possible over a fairly wide temperature range. Since sufficient information is available in the literature, this subject has not been investigated in the laboratory. Briefly, the procedure is to determine the variation in frequency with temperature and then to select suitable negative- or positive-temperature coefficient
capacitors to compensate for the variation. It may also be necessary to com-
pensate for variations in resonant impedance by using Thermistors or other tem-
perature sensitive devices.

The capacitive branch of a tuned circuit should always be investigated to
determine its effect on the Q of the circuit. The losses of some capacitors
are dependent on temperature and this may cause additional circuit instability.
There have also been some indications of capacitor nonlinearity.

The Q of resonant circuits is often greatly reduced when they are mounted
in close proximity to other components. This effect may be reduced to some
extent by keeping the diameter of the inductance small. The inductive field
may in some cases affect the performance of the transistor if it is placed too
close to the inductance. This possibility has not yet been investigated since
it was usually satisfactory to keep the components separated.

The effective Q of resonant circuits is usually greatly reduced by the
loading effect of the transistor. This effect may be partially overcome by
giving careful attention to impedance levels in designing the oscillator.
Usually, a compromise among L/C ratio, Q and resonant impedance is necessary.

The requirements on lumped resonant elements are so varied that no gen-
eral design method can be stated.

C. The Importance of Q and Resonant Impedance

In some oscillators, as pointed out previously, the Q of the frequency-
determining element is not necessarily the most important factor governing sta-
bility. This is particularly true in the negative-resistance oscillators where
the resonant impedance and L/C ratio both affect the stability. As an example,
with the common-collector point-contact oscillator, a parallel-resonant circuit
may be used in the base to control the frequency. The requirement for oscillation is that the parallel-resonant impedance of the tuned circuit be greater than the negative base resistance. However, best stability is usually obtained when this impedance is only a few per cent greater than the negative resistance. Best stability is also obtained when the L/C ratio is small. Therefore, for fixed impedance level and small L/C ratio, the inductance is required to be so small that it is difficult to obtain a high Q. In this case, the proper impedance match is more important than high Q.

Another example of poor stability resulting from increased Q is a situation in which the increase in Q is obtained by using a magnetic core. In this case, the nonlinearity of the core may cause greater instability than would be obtained by not using the core and accepting the lower Q.

In most feedback oscillators, the resonator impedance and L/C ratio are of lesser importance than Q. However, it is usually desirable that the resonator impedance be of the same order of magnitude as either the amplifier input impedance or the output impedance so that only one coupling transformer will be necessary. Variations in resonator impedance are still of major importance since such variations will affect the amplitude of oscillations and, thus, change the amount of distortion which in turn will change the frequency of the oscillation. In the case of crystal resonators, desirable properties are high Qs and constant resonant impedances. Of course, the resonant frequency must also be constant for high stability.

D. Impedance Inversion

Some transistor oscillator configurations require the use of a parallel-resonant circuit as the frequency-determining element. As mentioned in section III-A, quartz crystals do not operate well in their parallel mode in
transistor oscillators because of their very high resonant impedance. Thus, a network is desirable whereby the crystal can operate in its series mode while at the same time the impedance at the network terminals represents a parallel resonance. This may be accomplished by using a quarter-wave transmission line as shown in Figure 3.2.

![Figure 3.2. A Quarter-Wave Transmission Line as an Impedance Inverter.](image)

The relation among the impedances is

$$Z_o^2 = Z_1 Z_r$$  \hspace{1cm} (3.7)

where $Z_o$ is the characteristic impedance of the transmission line, $Z_r$ is the impedance of the crystal and $Z_1$ is the terminal impedance of the complete circuit, assuming that the transmission line is exactly one-quarter wavelength long.

If the crystal is operating at series resonance, $Z_r$ will be the series-resonant impedance, $R_s$, of the crystal. As an example, if a crystal with a series-resonant impedance of 100 ohms is to be used to represent a parallel-resonant circuit with an impedance of 10,000 ohms, the required impedance of the transmission line is 1000 ohms.

-88-
Two of the disadvantages of this arrangement are the excessive length required if a distributed line is used and the difficulty in obtaining lines with sufficiently high impedance. Both of these difficulties are partially overcome by using lumped-constant lines. However, this practice often results in many spurious resonances which may produce oscillations at unwanted frequencies. These spurious resonances may, in many cases, be eliminated by shunting the crystal with a resistance. The resistance also provides a d-c path through the network so that bias may be supplied to the transistor.

The simplest form of the lumped-constant line is the three-element pi-section network shown in Figure 3.3. It is not even necessary that all of these elements appear as physically lumped units in the circuit. For example, $C_1$ may be the transistor capacitance and stray circuit capacitance, $C_2$ may be the crystal-holder capacitance and $L_1$ may even be the inductance of the connecting wire (only at very high frequencies).

![Figure 3.3. Pi-Network Impedance Inverter.](image)

The use of impedance-inverting networks in transistor oscillators has not been generally successful. This has partly been due to the high resonant impedance of most of the available crystals (100 to 1000 ohms) and the high
parallel-resonant impedance (1000 to 10,000 ohms) required by the transistor circuits. Since most of the oscillators to which this principle is applicable are of the negative-resistance type, much difficulty has been encountered from spurious oscillations. It has been found almost impossible to eliminate all spurious resonances from even the simple pi-network. However, since most of these resonances occur at higher frequencies, the use of transistors of limited frequency response helps to control the oscillations.

A more extensive discussion of this subject is available in the literature.
IV. TYPICAL TRANSISTOR OSCILLATOR CIRCUITS

A. Negative-Resistance Oscillators

There are three basic negative-resistance oscillator configurations. These are shown in Figure 4.1. Bias supplies are omitted from the circuits for simplicity. In each of the circuits, definite relations must exist among $R_e$, $R_c$, $R_b$ and the transistor parameters for oscillations to occur. All three of these circuits may be used as LC controlled oscillators. Figure 4.1 (c) offers no difficulties in supplying bias to the transistor since the inductance provides a d-c path. To supply bias to the transistor in Figures 4.1 (a) and (b), it is necessary to use an additional impedance shunting the LC circuit. Since the LC circuit is operating at series resonance and therefore has a low impedance, its $Q$ is not greatly reduced by shunting it with a resistor.

![Figure 4.1. Basic Negative-Resistance Oscillator Configurations.](image)

Figures 4.1 (a) and (b) may also be operated with crystal control by shunting the crystal with a bias impedance. However, when a crystal is used in Figure 4.1 (c), it must also be shunted by a bias impedance since the
crystal is an open circuit at direct current. But, the crystal, when operated at antiresonance, has a very high impedance. Shunting it with a bias impedance will greatly lower its Q. Also, the mismatch between the crystal and the transistor is very great. In some cases, the circuit will not even oscillate. This is one of the circuits where it is desirable to use the impedance-inverting circuits previously discussed. The crystal, operating in its series mode, may then be paralleled by a resistance to provide bias current for the transistor.

Extensive experimental data have been obtained on the negative-resistance type of oscillator. Some of the more stable circuits will be discussed in section VII.

B. Free-Running Feedback Oscillators

Several typical free-running feedback oscillators are shown in Figure 4.2. The tuned-collector, Pierce, Hartley, Colpitts, tuned-collector tuned-base and tuned-base circuits correspond to the vacuum-tube tuned-plate, Pierce, Hartley, Colpitts, TOTP and tuned-grid oscillators respectively. The basic principles of operation of these circuits will not be discussed here since they are very similar to the vacuum-tube equivalents, which are fully described in the literature.

The oscillator of Figure 4.2 (a) has no vacuum-tube equivalent since it requires a configuration which does not have phase reversal but yet has both voltage and current gain. It can, therefore, be used only with point-contact transistors.

The oscillator of Figure 4.2 (b) required that the amplifier operate at a frequency where its phase shift approaches 90 degrees. Its configuration appears similar to Figure 4.2 (a), but its operation is entirely different. In
Figure 4.2. Typical Free-Running Feedback Oscillator Circuits.
Figure 4.2 (b), the input impedance of the emitter is low, therefore, the collector operates into a parallel-resonant circuit consisting of L and C. A phase shift of almost 90 degrees occurs through the capacitor, C. Thus, the phase condition for oscillation can be met. Current gain is obtained since the emitter is driven by a series-resonant branch while the collector sees a parallel resonance. In this way, the amplitude condition for oscillation can also be met.

Some basic oscillator configurations other than those shown are known to exist and have in some cases been used. Also, many modifications of the circuits shown are possible. Detailed discussions will appear in section VII in connection with the practical oscillator circuits.

C. Crystal-Controlled Feedback Oscillators

All of the circuits of Figure 4.2 may be crystal-controlled without changing the configuration. Most of them require the use of the tuned circuit shown in addition to the crystal. With some of the circuits, the crystal replaces the LC combination.

Figures 4.2 (b), (c), (e), (f) and (h) may be crystal-controlled by inserting the crystal at the points indicated by the X. The LC combination must be resonated near the crystal frequency. In all of these circuits, the crystal operates in its series mode and acts as a bandpass filter. These circuits may also be made to operate at crystal overtones or any other frequency where the crystal has a low-impedance resonance.

Figure 4.2 (a) may be crystal-controlled by replacing L and C with a crystal shunted by a bias impedance. Since the crystal must operate in its parallel mode, the Q will be greatly reduced by the bias impedance. This is another one of the circuits with which the impedance-inverting network may be used to advantage.
Figure 4.2 (d) may be crystal-controlled by replacing the L and C by the crystal. Again the operation is similar to the vacuum-tube Pierce oscillator. By proper adjustment, this circuit may be made to operate at crystal overtones.

Figure 4.2 (g) may be crystal-controlled by replacing either LC circuit with the crystal shunted by a suitable bias impedance. Since the crystal must operate in its parallel mode in this circuit, the stability will be poor for the reasons mentioned previously. When the base circuit is replaced by a crystal, this circuit becomes the conventional Miller oscillator.

Additional crystal-controlled oscillator circuits are shown in Figure 4.3.

![Figure 4.3. Additional Crystal-Controlled Circuits.](image)

Figures 4.3 (a) and (b) appear at first to be negative-resistance oscillators; however, transistors with current gains greater than unity are not
required for these circuits. They operate on the principle that a resonant
circuit can provide current gain (resonant rise in current). The operation is
similar to that of Figure 4.2 (b). The transistors in these circuits are ac-
tually operated at zero d-c base current, that is, the crystal in Figure 4.3
(a) and the capacitor in Figure 4.3 (b) are not shunted by a d-c bias imped-
ance.

Figure 4.3 (c) is an example of a circuit capable of effectively in-
creasing the Q of the crystal and thus giving improved stability. Otherwise,
the circuit is similar to Figure 4.2 (c).

Figure 4.3 (d) is a simplified Pierce oscillator. Only four components
(transistor, crystal, resistor and power supply) are required when the proper
transistor is used. This circuit also operates the transistor at zero base
current. By the addition of a capacitor across the resistor, oscillations can
be obtained at crystal overtones.
V. SELECTION OF TRANSISTORS

It is usually possible to state the required transistor type and desirable transistor characteristics for each of the previously mentioned circuit configurations (except when certain modifications of the circuits are used). This information is given in Table 5.1. References to specific transistor types as listed in Table 5.2 are based on transistors that have been used in the particular circuits. It is not necessarily implied that the transistor type used has all of the desired characteristics. Not all of the transistors listed were available at the time that most of the circuits were initially investigated. Consequently, only the more promising circuits were investigated again at a later date.

It has been concluded from laboratory data that several of the transistor types tested are not suitable for any type of oscillator at any frequency where stability is of major importance. These transistors include the TA-153, TA-165 and RD-2525A.

Among the point-contact transistors, the best for use as an oscillator below 1 mc has been found to be the WE 1729. It also has been found fairly suitable for use as an amplifier in such circuits as the Meacham bridge oscillator since it is capable of providing greater gain than most of the junction transistors at frequencies as high as 1 mc. The 2N33 is the only point-contact transistor that has been found suitable for use at higher frequencies. It has been used at frequencies as high as 95 mc with very limited stability.

For the first 18 months of the study, the only junction transistors found suitable for use as oscillators were the CK 720's (except possibly the TI 900 and TI 901 on which data have not been available). The GE ZJ1 was of some
TABLE 5.1
DESIRABLE CHARACTERISTICS OF TRANSISTORS FOR OSCILLATOR PROTOTYPES

<table>
<thead>
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<th>Necessary Characteristics</th>
<th>Transistors Used† † †</th>
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† Desired and necessary characteristics:
1. Point contact only (alpha must be greater than unity).
2. Junction only (difficult to stabilize with point contact).
3. Point contact or junction.
4. High alpha.
5. Cut-off frequency well above operating frequency.
6. Cut-off frequency well below operating frequency.
7. Transistor parameters should be highly stable with respect to changes in voltage and temperature.
8. Transistor parameter variations with respect to voltage and temperature should be moderate.
9. Transistor parameter variations of lesser importance.
10. May require specially chosen transistors.
11. Impedance levels of transistors should be high.
12. Impedance levels of transistors should be low.

† † † Numbers refer to Table 5.2.

-98-
## TABLE 5.2

**TRANSISTOR TYPES AVAILABLE TO THE PROJECT**

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</tr>
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†Notes:

1. This transistor has not been used in laboratory work because of fragile terminals. Characteristics almost identical to WE 1698.
2. Project has been unable to obtain manufacturer's or other data on these transistors.
3. Measured characteristics differed very greatly from manufacturer's data. Gain very low, apparently because of age and lack of hermetic sealing.
5. Apparently similar to TI 904 except lower gain and worse frequency response.
6. Low gain, poor frequency response.
7. Except for 2N44, these transistors were investigated only very briefly because of forthcoming termination of project. The 2N44 was investigated in many circuits previously designed for the CK 720.
usefulness but was inferior to the GE 2N44's which were received later. The TI 900 and 901 were used to some extent when it was learned that they were silicon junction transistors. However, they were later replaced by TI 904.

The most stable uncompensated single-transistor oscillators which have been designed used Philco L-5100 transistors. Indications are that this transistor is excellently suited for any oscillator requiring a junction transistor except where signal levels are a limiting factor. The major disadvantage of this transistor is that the available power output is very low. It was found that the characteristics of all samples of this transistor were almost identical.

The TI 904's were found to provide operation almost as stable as the L-5100's. In addition, they provide much greater power output.

The last four transistors listed in Table 5.2 were not investigated to any appreciable extent because they were only recently received.

Junction transistors have usually been found more reliable and uniform than point-contact transistors and are, therefore, recommended for use in circuits where either type may be used.

The following conclusions are of general importance in practically all oscillators and serve as a guide in the choice of transistors.

1. The transit time should not change appreciably with voltage and temperature.

2. The element capacitances should not change appreciably with voltage and temperature.

3. The gain should not change appreciably with voltage and temperature.

4. Different samples of the same transistor type should show uniform characteristics.

5. The transistor should be capable of providing the required power output.
Concerning requirements 1 and 2, practically all transistors have been found to have similar percentage wise variations of transit time and capacitance with voltage. The silicon transistors generally show slightly less variation with temperature. Thus, the only practical way of meeting requirements 1 and 2 is to use transistors having small transit times and capacitances. This implies that the transistor cut-off frequency should be high compared to the operating frequency, except that a few oscillator circuits require the converse to be true. These latter oscillators are likely to have very poor stabilities.

Variations in gain for changes in voltage and temperature apparently depend more on manufacturing techniques (or materials used) than on the type classifications of the transistors. It has not been possible to predict the performance of a transistor from the published data. The measurement of parameters as functions of voltage and temperature is always necessary. In some cases, predictions of stability with changes in voltage can be made from published curves when they are available. The published curves on point-contact transistors have not been found useful because of the great variations between individual samples. Junction transistors exhibit greater uniformity.

Recommendations of transistor types cannot be made except on the basis of the types used by the project. Among those listed in Table 5.2, the most desirable for general oscillator use were the Philco L-5100 and the TI 904.

The selection of transistors on the basis of required power output can readily be made from manufacturer's data.
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VI. OSCILLATOR STABILIZATION

A. Bias Compensation

This chapter suggests methods whereby an already fairly stable oscillator can often be made more stable by various compensating means. One of the simplest of these methods is bias compensation by which the oscillator is stabilized with respect to voltage changes. Bias compensation may be divided into two categories (1) linear compensation and (2) nonlinear compensation.

In order to use linear bias compensation, it is necessary that the oscillator be designed to have two separate bias sources. It is further necessary that the frequency variations caused by changes in the bias sources be in opposite directions (unless complicated bridge biasing systems are to be used).

The method of applying linear bias compensation is best illustrated by the example which follows.

It will be assumed that the oscillator has separate emitter and collector bias sources and that the frequency variations are as shown in Figure 6.1. It will further be assumed that the input resistances (dc) at the emitter and collector are constant. (If the emitter input resistance is very small, it is not necessary that it be constant.)

From Figure 6.1, the following relationships are obtained.

\[
\frac{\Delta f}{\Delta V_c} = \frac{1}{5} = 0.2 \text{ cycle per volt}
\]

and

\[
\frac{\Delta f}{\Delta I_e} = -\frac{1}{0.5} = -2.0 \text{ cycles per milliampere.}
\]

It is now only necessary to design a resistive voltage divider that will provide 0.1-milliampere increase in emitter current for each volt increase in
collector voltage. If a nominal collector supply voltage of 25 volts is assumed, a one cps change in frequency occurs for a 20 per cent increase in collector voltage. To balance this change, the emitter current must be caused to increase by 0.5 ma. Since the system is assumed to be linear, this must represent a 20 per cent increase in emitter current. Thus, the nominal emitter current is chosen to be 2.5 ma. The circuit is as shown in Figure 6.2. $R_1$ and $R_2$ should be made sufficiently large so that the emitter is current controlled. The actual values of $R_1$ and $R_2$ are chosen on the basis of the available supply voltage, $V$, and the collector current, $I_c$.

![Figure 6.1. Frequency Versus Voltage and Current for the Example.](image-url)

If the relation between voltage (or current) and frequency had not been linear or if the d-c collector input characteristic had not been a constant resistance, nonlinear compensation would have been necessary. The term nonlinear compensation here implies that Varistors or other nonlinear resistances must be
used in the biasing divider. The procedure is similar to the above except somewhat more complicated. Since no one procedure will apply to all cases, an example will not be given here.

In optimum situations, an improvement in voltage stability as great as one order of magnitude can be expected from this procedure where it is applicable. In some cases, this procedure is not applicable, for example, when the frequency is already independent of one of the source voltages. In such situations, distortion compensation can usually be used to improve the stability. Since the bias compensation procedure depends upon cancellation effects, the degree of improvement is often dependent upon a number of other variables such as temperature, aging of transistors and, in some cases, operating frequency. Likewise, frequency variations at supply voltages greatly different from the design center values may be greater than for the uncompensated oscillator.

![Circuit Diagram](image)

Figure 6.2. Circuit for the Bias Compensation Example.

B. Component Selection and Application

Component selection may often be thought of as a stabilization procedure since the stability is often limited by the passive components rather than the transistor. In many cases, any components available have been used; sometimes by necessity and at other times under the philosophy that the design should be such that the stability is fairly independent of component variations.
Several general statements can be made concerning the types of components used by the project. The following list also contains some practical construction notes.

1. Composition resistors are usually satisfactory for most applications.

2. Wire-wound resistors should not be used in critical locations unless their characteristics are completely known. Such resistors may have large reactive components at high frequencies. Their temperature coefficients are generally lower than those of composition resistors, but in some cases they may be larger.

3. The characteristics of printed and tape resistors may either be desirable or undesirable for oscillator applications. Their characteristics should be carefully measured before they are used.

4. Mica capacitors are usually satisfactory even for use in frequency-determining elements. The losses are usually very low.

5. Ceramic capacitors should not be used in oscillator applications (except possibly for bypasses) without first investigating their characteristics. Samples tested by the project have usually been found to have very poor temperature characteristics and excessive losses. In addition, many of them have nonlinear voltage characteristics which can increase oscillator distortion.

6. Paper capacitors have usually been found to have more losses than mica capacitors but have been found to be more linear than ceramic capacitors. The temperature characteristics are usually good.

7. Commercial inductances have been found to be quite variable in characteristics. One of the major problems has been spurious resonances. It is recommended that inductances not be used as r-f chokes in transistor oscillator circuits unless impedances higher than obtainable with resistors are absolutely
necessary. In such cases, the use of a parallel-resonant circuit is preferable to an r-f choke if a bandpass arrangement is permissible.

8. Inductances designed for use as r-f chokes should not be used in resonant circuits unless a low Q is desirable.

9. Inductances designed for use in resonant circuits should not be used as r-f chokes as sharp self-resonances are likely to exist.

10. Ceramic forms have been found to be the best for inductances to be used in resonant circuits. Practically all other materials show a greater variation in inductance with temperature.

11. Inductances with magnetic cores are not usually satisfactory for use in frequency-determining elements because of the nonlinearities and poor temperature characteristics.

12. When designing crystal-controlled oscillators, the characteristics of the quartz crystal should be carefully investigated. Variations in resonant impedance with temperature have been found particularly bad in some crystals. This consideration is usually more important for transistor oscillators than for vacuum-tube oscillators because of the limited power gain of the transistor. Also, variations in frequency with ambient temperature have in some cases been exaggerated by using the crystal in an oven when it was apparently designed for use at room temperature.

13. When Varistors are used for voltage stabilization, the temperature characteristics should also be considered. Tungsten lamps operated at fairly high temperatures have been found to be less sensitive to ambient temperature than most of the low-temperature commercial Varistors.

14. Tungsten lamp characteristics are subject to wide variations. Thus, it is usually necessary to determine the characteristics of each lamp individually.
15. The mechanical construction of tungsten lamps is very important when they are to be used in r-f bridges. Lamps with large inductances or wrapped terminal connections should be avoided. Most of the switchboard type of lamps are usable to 1 mc or higher in frequency.

16. Mechanical rigidity is an important consideration at higher frequencies. Components should not be self-supported on long leads. The transistors have been found to be the least microphonic component used in oscillator construction. Quartz crystals have been observed to be appreciably microphonic.

17. Attempts to miniaturize transistor oscillators have often resulted in reduced stability. This has been, in some cases, due to unwanted couplings and, in other cases, to the poorer quality of some miniature components. Some miniature components are, on the other hand, more stable and more reliable than their larger counterparts.

18. Variable capacitors are usually microphonic and subject to humidity effects. The moving contacts are sometimes not suitable for use in low-impedance circuits. With some miniature variable capacitors, the losses have been substantially increased by the spreading of solder flux across the contacts while making connections. This may be avoided by properly cleaning the contacts and by using a solder which does not contain flux.

19. Components often become very erratic from a single overheating during soldering. Great care should be taken to avoid this.

20. Poor solder connections are more troublesome in transistor oscillators than in vacuum-tube oscillators. This is due to the necessarily greater care against overheating and also to the lower impedances to be found in most transistor oscillators.
Some of the above suggestions have been previously stressed for all types of electronic construction. They are mentioned again here because of their even greater importance in transistor circuitry.

C. Temperature Compensation

Temperature compensation may be applied to a transistor oscillator in many ways. General over-all temperature compensation is sometimes possible and often simpler than compensating each of the individual causes of the instability. Of course, no compensation can be as effective as finding the cause of the instability and correcting it at its source.

The over-all compensation will be considered first. The first step of the procedure is to determine the variations in frequency with temperature. The second step is to find a component in the oscillator which is capable of causing an equal variation in frequency, for example, a tuning capacitor. The third step is to select a substitute component with the proper temperature characteristic, for example, a positive or negative temperature coefficient capacitor. The major difficulty with this method is in finding components which vary in the proper manner with temperature. Another difficulty is that the frequency-temperature characteristic of a transistor oscillator is often dependent on the rate of heating or cooling. Usually, the very best compensation that can be achieved by this method is one order of magnitude of improvement. A typical improvement may be by a factor of only two or three.

Compensating each circuit component against temperature is another approach to the temperature stability problem. The first step is to select highly stable passive elements (or compensate poorer elements until they are sufficiently stable). The selection of resistors, capacitors and inductors presents no
particularly difficult problem. The compensation of quartz crystals is somewhat more difficult although it can usually be accomplished by combinations of Thermistors and high-temperature-coefficient capacitors. (A better approach for laboratory purposes is to use a highly stable crystal oven with a crystal designed to operate at the oven temperature.) Temperature compensation of the transistor offers the greatest problem. This subject has been treated at length in the literature, although it appears that most of the methods suggested leave much to be desired. This subject has not been investigated to any great extent by the project.

Compensation of diodes, Varistors and other such components is beyond the scope of the project.

D. Stabilization by Distortion

Several methods of using distortion to stabilize transistor oscillators against frequency variations due to voltage changes have been investigated. The simplest of these methods is shown in Figure 6.3.

![Figure 6.3. Diode Stabilized Feedback Oscillator.](image)
Two crystal diodes are used connected front-to-back across the amplifier output with bias voltages as shown. The capacitor is used to prevent upsetting the d-c transistor biasing. If the two diodes were perfect, the output waveform would be symmetrically clipped at a level determined by the bias voltages in series with the diodes. Typical diode characteristics are shown in Figure 6.4. If sufficient signal current is available from the amplifier, the silicon diode will provide satisfactory clipping without series bias voltages as can be seen from the figure. The silicon junction diode is to be preferred for its volt-ampere characteristic whereas the germanium point-contact diode has better high-frequency characteristics.

The theory for this type of stabilization is based on the fact that a large part of the frequency shift that occurs when the supply voltage is changed is due to changes in the amount of distortion. The purpose of the diodes is to cause severe distortion but, at the same time, make the distortion less dependent

---

**Figure 6.4. Typical Diode Characteristics.**
on supply voltage. In almost all cases, this method will improve the voltage stability if the output amplitude is sufficiently large to be clipped satisfactorily. In some oscillator circuits, overcompensation is possible. Where this is true, the distortion may be controlled by inserting a variable resistance in series with the diodes.

A serious objection to this method of compensation is that the output voltage is greatly reduced, but it has the advantage that the output amplitude stability is greatly improved. The addition of the diodes usually has some effect on the temperature stability since the diodes themselves are relatively sensitive to temperature. In some circuits, the diodes actually improve the temperature stability.

Another method of stabilizing an oscillator by distortion is shown in Figure 6.5.

Figure 6.5. Another Method of Stabilization by Distortion.

This method requires only one diode (the transistor emitter-base junction acts as a second diode). No capacity coupling is used in this circuit. As a result the d-c operating point is such that overcompensation is possible. This
is advantageous as it permits using a resistor, $R$, to control the amount of distortion and, thus, the amount of frequency correction. A value for $R$ can often be found such that no appreciable change in frequency occurs for any small change in supply voltage in the vicinity of the nominal voltage. The compensation is generally not perfect over the complete voltage operating range. Typical performance curves for a free-running Hartley oscillator using this type of compensation are shown in Figure 6.6. A single voltage source was used for both emitter and collector bias to obtain these curves. The curves show that the additional distortion introduced by the diode causes a decrease in frequency as would be predicted from the theory. The particular circuit for which these curves apply will be discussed in section VII.

The compensating methods discussed here may be applied to either free-running or crystal-controlled feedback oscillators and also, in most cases, to negative-resistance oscillators. The use of compensating diodes occasionally results in spurious oscillations. This is likely to occur only when more than one tuned circuit is used in the oscillator.

E. Bridge Stabilization

Several bridge circuits for improving the frequency stability of transistor oscillators have been investigated. The circuit which has given the best frequency stability of all the circuits constructed used the Meacham bridge circuit. A discussion of the Meacham bridge will not be included here since adequate material is available in the literature on this circuit. Examples of transistorized Meacham bridge oscillators will be included in section VII.

Another bridge circuit which has been used to improve frequency stability was shown in Figures 2.38 and 4.3 (c). This circuit is not always recognized as
Figure 6.6. Effect of Compensating Diode on Free-Running Hartley Oscillator.
a bridge since two of the bridge arms are transformer windings. The design of
this circuit was discussed in section II-E.

In each of the two circuits mentioned above, the bridge is used in the r-f
portion. This arrangement results both in improved frequency-voltage stability
and in frequency-temperature stability except where temperature instability is
caused primarily by the temperature characteristics of the resonant element.
The Meacham bridge circuit also greatly improves the output amplitude stability.
For satisfactory operation of the Meacham bridge, however, more gain is required
than is available from one transistor amplifier stage. Also, the power output
required for the bridge may be as high as 40 milliwatts depending on the choice
of Varistor. The transformer bridge has no specific minimum power requirement
for oscillation.

The use of a bridge circuit for bias stabilization is sometimes advanta-
geous. This is true especially where the oscillator frequency changes in the
same direction when either the collector voltage or the emitter current is in-
creased. Figure 6.7 shows an arrangement which may be used for this purpose.
R represents fixed resistors and V represents Varistors. The values to be used
for the resistors and Varistors depend upon the characteristics of the particu-
lar oscillator with which the bridge is to be used.

Thermistors may also be used in this circuit to provide temperature stabili-
sation. It is also possible to combine both the voltage and temperature sta-
bilization in one bridge.

Other bridge circuits have been investigated but have been found to be of
relatively little importance except in special applications.
F. Transistor and Crystal Replaceability

Most of the stabilization and compensation procedures which have been proposed will affect the replaceability of transistors and crystals. The changes are not always obvious or predictable. However, some conclusions may be stated concerning the expected results. For example, the stabilization by distortion usually improves the transistor replaceability although some less active transistors may fail to produce oscillations. For the transistors that do operate properly, the frequency stability will almost certainly be improved. Crystal replaceability is affected very little by the distortion compensation.

Stabilization by the use of r-f bridges almost invariably improves the transistor replaceability while the crystal replaceability is almost always made worse. The reason is that the r-f bridge makes the circuit relatively independent of the amplifier characteristics, but the bridge balance depends
critically upon the crystal impedance. Substitution of crystals with imped-
ances differing by as much as 50 per cent may cause the circuit to cease os-
cillating until the crystal or other bridge components are readjusted. How-
ever, once the proper adjustments have been made reliable and stable opera-
tion is obtained.

Bias compensation may either increase or decrease the transistor replace-
ability. For transistors with characteristics differing only slightly from the
transistor for which the circuit was designed, an improvement in stability over
that obtained without compensation may be expected. When transistors of greatly
different characteristics are used, the stability may be worse than that ob-
tained without the compensation. For these reasons, bias compensation is more
successful with junction transistors than with point-contact transistors. With
regard to crystal replaceability, the trend is toward an improvement. However,
the effect may be either way.

With temperature compensation and d-c bridge stabilization, either an in-
crease or a decrease in transistor replaceability may be obtained. The trend
is toward an improvement for junction transistors. The effect on crystal re-
placeability is only slight.

The proper selection of components usually improves replaceability of both
transistors and crystals. "Selection" here refers to the type of component and
not to specific component values.

Even though the replaceability may be made worse in some cases, the various
compensation and stabilization procedures may usually be used to improve greatly
the frequency stability of oscillators, provided the circuits are designed aft-
er a particular transistor and crystal combination has been chosen.
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VII. RECOMMENDED DESIGNS

A. Properties of Oscillators Investigated

Table 7.1 shows the relative characteristics of some of the oscillator circuits investigated by this project. The inclusion of an oscillator circuit in this table does not necessarily indicate that it is a superior or recommended type, but merely indicates that it is representative of particular types. The purpose of the table is to compare as many different oscillator configurations as possible and, in some cases, to show the relative results of modifications.

The circuits for these oscillators are shown later in section VII-C.

B. Selection and Recommendations of Circuits

Table 7.1 may be used as a guide in the selection of suitable oscillator circuits for particular applications. It does not give quantitative information since this will vary with the type of transistors and even with the particular transistor within a type. The quantitative performance also depends on the physical layout of the circuit and the care with which it is assembled. Of course, the quality of the passive components used will always affect the performance.

After an initial choice of possible circuits is made from Table 7.1, the final choice can best be made from the data given in section VII-C. In that section, typical quantitative performance will be given for particular transistor types.

As a further aid to circuit selection, the recommended circuits for various requirements are shown in Table 7.2. Where frequency is mentioned in this table it is relative to alpha cut-off, thus a "low frequency" surface barrier transistor oscillator may operate at a much higher frequency than a "high frequency" point-contact transistor oscillator.

-119-
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<td>Crystal Replaceability</td>
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<td>&gt; qf CO</td>
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</table>

1Refers to alpha cut-off.

**NOTE:** Explanation of symbols used in the table.

- $S_v$: voltage-frequency stability as defined in section II-A. For crystal-controlled oscillators; $P < 1$, $F = 1-10$, $G = 10-100$ and $E = > 100$. For free-running oscillators; $P = < 0.01$, $F = 0.01-0.1$, $G = 0.1-1.0$ and $E = > 1.0$.

- $S$: temperature-frequency stability as defined in section II-A. $P$ means that the frequency changes caused by the transistor and other elements are greater than the expected changes due to temperature variation of non-temperature controlled LC elements. $F$ means that the frequency changes caused by the transistor and other elements are greater than the expected changes due to temperature variation of the crystal (if the oscillator is free-running, the stability can usually be improved by placing the LC circuit in an oven). $G$ means that the temperature stability is sufficiently high that the use of a crystal oven is necessary to realize the maximum stability of the oscillator.

- Power Output is defined as follows: $V$ means voltage only available (load impedance greater than 10,000 ohms). $P_A$ means that appreciable power is available by simple circuit modifications. $P$ means that appreciable power output is available without modification of the circuit.

- Output Isolation is related to the effect of the load on the frequency. $P$ means that changes in load greatly affect the frequency. $F$ means that the frequency is affected slightly by changes in loading. $G$ means that the loading has negligible effect on the frequency.

- Complexity refers to the number of components used in the circuit. $S$ means that few components are required, but no special components are used. $SC$ means that few components are required, but some special components are used. $C$ means that a large number of components are used or that some of the components are highly specialized.

- Transistor Replaceability: $P$ means that the circuit usually has to be adjusted when transistors are replaced. $G$ means that the circuit usually requires no readjustment when transistors are replaced.

- Crystal Replaceability: $P$ means that the circuit usually has to be readjusted when crystals or tuned circuits are replaced. $G$ means that all crystals or tuned circuits within normal tolerances will operate properly in the circuit without readjustment of circuit values.

- Output Voltage Stability: $P$ means that the output voltage varies more than the supply voltage. $F$ means that the output voltage is practically unaffected by supply voltage variations. $G$ means that the output voltage varies less than the supply voltage.

- Ease of Design: $S$ means that little or no mathematics and only elementary laboratory adjustments are involved. $SL$ means simple to design but that appreciable laboratory adjustment may be required. $C$ means that a complex design procedure is involved.

- Supply Voltage: $L$ means less than 5 volts. $M$ means 5 to 15 volts, $H$ means 1 to 40 and $VH$ means greater than 40 volts.

- Frequency Range is the maximum frequency (in terms of alpha cut-off) for reasonable stability.

- Type of Transistor is either PC (Point Contact) or Jun (Junction) or Any (either type may be used).
<table>
<thead>
<tr>
<th>Primary Interest</th>
<th>Desirable Characteristics of Circuit</th>
<th>Undesirable Characteristics of Circuit</th>
<th>Figure Number</th>
</tr>
</thead>
<tbody>
<tr>
<td>Good frequency stability with p.c.t.</td>
<td>Good frequency stability, constant output amplitude, may be controlled or free-raning, uses one voltage supply, simple to design, requires no special components.</td>
<td>High supply voltage, low power output, operates well only below alpha cut-off frequency, requires two diodes, alpha of transistor must be greater than two.</td>
<td>7.19</td>
</tr>
<tr>
<td>Very good frequency stability and high output voltage with p.c.t.</td>
<td>Very good frequency stability, high output voltage, uses no diodes, single supply voltage.</td>
<td>Poor output amplitude stability, high supply voltage, requires a special transformer, operates only below alpha cut-off frequency, crystal control only, adjustment for best stability sometimes difficult.</td>
<td>7.16</td>
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<tr>
<td>High-frequency operation with moderate output voltage using p.c.t.</td>
<td>Operates well at frequencies beyond two times alpha cut-off, moderate output voltage, uses no diodes, uses no transformers or other special components, single supply voltage.</td>
<td>Poor frequency stability, free-running only, high supply voltage, poor output voltage, amplitude stability cannot readily be crystal-controlled.</td>
<td>7.10</td>
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<tr>
<td>Constant output voltage with high frequency using p.c.t.</td>
<td>Constant output voltage, good frequency stability, requires no transformers or other special components, simple to design and adjust.</td>
<td></td>
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</tr>
<tr>
<td>Exceptional frequency stability with low supply voltage.</td>
<td>Exceptional frequency stability, low supply voltage, requires no diodes.</td>
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<tr>
<td>Very high frequency junction transistor oscillator.</td>
<td>Operates to beyond four times alpha cut-off, may be crystal-controlled or free-raning, uses no diodes, uses no transformers or other special components. Figure 7.40 requires only one supply voltage, simple to design.</td>
<td>Only moderate stability, high supply voltage, poor output amplitude stability, operates only beyond alpha cut-off frequency, Figures 7.36 and 7.38 and require two supply voltages.</td>
<td>7.27</td>
</tr>
<tr>
<td>Good frequency stability at medium frequencies with junction transistor.</td>
<td>Good frequency stability, medium frequency operation, requires no transformers, may be controlled or free-raning, good output amplitude stability, simple to design.</td>
<td>High supply voltage, requires one or two diodes, output voltage low for Figure 7.32, and 7.34.</td>
<td>7.30</td>
</tr>
<tr>
<td>Good frequency stability at medium frequencies with junction transistor.</td>
<td>Good frequency stability, medium frequency operation may be crystal-controlled or free-raning, has high output voltage, uses one voltage supply, simple to design.</td>
<td>High supply voltage, requires one or two diodes, requires special output, may have to be readjusted for best stability for each different transistor.</td>
<td>7.21</td>
</tr>
<tr>
<td>Exceptional frequency stability with appreciable output power from medium voltage.</td>
<td>Exceptional frequency stability, appreciable output power, moderate supply voltage, good transistor replaceability, uses no diodes, very good output amplitude stability.</td>
<td>Requires two or three transistors, requires two or three special transformers, circuit fairly complicated, suitable amplifier may be difficult to design.</td>
<td>7.46</td>
</tr>
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</table>
The circuits shown on the following pages are representative of those that have been studied. The transistor types and the circuit values shown in the diagrams are those actually used in the tests. Below each circuit are shown the performance data for this particular choice of components. Also included with some of the circuits are recommendations for modifications that should improve the frequency and/or voltage stabilities.

Some of the suggestions on the following pages have not been tried in the laboratory because of lack of sufficient time. However, the information given should be reliable since it is based on laboratory experience with the components and procedures as well as the various theories and discussions which have preceded.

Most of the circuits shown can be readily adapted to newer types of transistors by choosing transistors with the characteristics suggested. Adjustments in the other component values will usually be required. Where the methods of making such adjustments are not indicated, they are usually obvious to anyone familiar with transistor oscillator circuits.

Some of the circuits are inferior to others in practically all respects. These are included for completeness and for their shortcomings which are not obvious.

The special terminology used is explained as follows:

The Prototype is the diagram from section IV which most closely resembles the circuit.

The Circuit Type is classified as feedback or negative resistance and crystal-controlled or free-running. When both crystal-controlled and free-
running are indicated, a series LC circuit may be used to replace the crystal or a crystal may be used to replace the series LC circuit (or the inductive arm of the series LC circuit).

The **Transistor Class** is either point-contact or junction.

The **Approximate Frequency** is the frequency range over which the circuit will oscillate with reasonable stability. The range over which it can be made to oscillate is usually much greater. Frequency is given in terms of alpha cut-off. Numbers in parentheses are the frequencies at which the stability data were obtained.

For definitions of **Stability**, see section II-A.

The **Transistor Types** are the transistors from the list in section V, which should operate properly in the circuit. When "any point-contact" is stated, it is implied that alpha must be greater than unity.

The **Design Procedure** is the type of procedure which has been or may be used to design the circuit. (References are to other parts of this report.)

Under **Power Output**, "voltage only" indicates that only high-impedance loads should be connected to the oscillator. Loads requiring appreciable current will affect the frequency or amplitude stability unless they are purely resistive and constant in magnitude. Unless otherwise stated, the peak-to-peak amplitude of the voltage output is between 50 and 200 per cent of the d-c supply voltage. "Low voltage only" indicates less than two volts peak-to-peak output.

The power supply voltage is not listed since it will depend upon the transistor type used. In some cases, the power supply voltage for the particular transistor used is shown in the diagram.

The **Desired Transistor Characteristics** indicate the transistor characteristics which are considered to be most important in obtaining good frequency stability.
The recommended output terminals are indicated by the letters X-X. These are not usually the only points from which the output may be taken, but they are the points considered most desirable from the consideration of output amplitude and loading.

In the notes, the temperature stability is often given as good, fair or poor. The meanings are as follows: good--the temperature stability is sufficiently high so that the use of a crystal oven is necessary to realize the maximum stability of the oscillator; fair--the frequency changes caused by the transistor and other elements are greater than the expected changes due to temperature variation of the crystal (if the oscillator is free-running, the stability can usually be improved by placing the LC circuit in an oven); and poor--the frequency changes caused by the transistor and other elements are greater than the expected changes due to temperature variation of non-temperature-controlled LC elements.
Figure 7.1. Collector-Controlled Free-Running Oscillator.

Prototype: 4.1 (b).

Circuit Type: Negative resistance, crystal-controlled or free-running.

Transistor Class: Point-contact only.

Approximate Frequency: 0 to alpha cut-off.

Stability: Quantitative data not available.

Transistor Types: Any point-contact.

Design Procedure: Linear circuit theory or Groszkowski analysis.

Power Output: Voltage only, unless a transformer is used in the emitter circuit.

Desired Transistor Characteristics: High alpha, constant h parameters.

Notes:

The frequency stability of this circuit is very poor. The resonant circuit does not provide satisfactory control over the oscillations. The primary reason for this is that impedance matching between the resonant circuit and the transistor is very difficult to obtain. The resonant impedance of the tuned circuit must be less than the negative resistance at the points a-a. This negative resistance may be anywhere from a few hundred ohms to 100,000 ohms depending on the frequency, transistor type, voltage and other factors. The resonant impedance of a typical series LC circuit is only a few ohms. This impedance cannot be raised by series resistance since the Q would be greatly lowered. As the frequency is raised, a better impedance match is obtained but the stability is usually worse because of the much greater variation in the complex transistor parameters with voltage and temperature.

The relative temperature stability of this circuit is poor.

This circuit is not recommended as a stable oscillator.
Prototype: 4.1 (b).
Circuit Type: Negative resistance, crystal-controlled only.
Transistor Class: Point-contact only.
Approximate Frequency: 0 to alpha cut-off (50 kc).
Stability: \( S_v = 0.5 \).
Transistor Types: Any point-contact.
Design Procedure: Linear circuit theory or Groszkowski analysis.
Power Output: Voltage only, unless a transformer is used in the emitter circuit.
Desired Transistor Characteristics: High alpha, constant \( h \) parameters.

Notes:

It was desired to use crystal control at low frequencies with a circuit of the type shown in Figure 7.1. The only crystal available for 50 kc was one with a series-resonant impedance of 40,000 ohms. The negative resistance at the collector was less than this value, therefore, oscillations could not be obtained in the circuit of Figure 7.1. By adding the parallel-resonant circuit in the base lead, the negative resistance could be raised to almost 100,000 ohms. Good impedance match was obtained with this circuit. However, the stability was poor because of the relatively poor crystal and relatively high frequency for the WE 1768. No difficulty should be experienced in obtaining much better stability with a WE 1729 and frequencies as high as 1 mc. If the resonant impedance of the base circuit is kept low, no oscillations will occur when the crystal is removed.

The temperature stability of this circuit was fair.

This circuit is not recommended where stability is important.
Figure 7.3. Tuned-Emitter Oscillator.

Prototype: 4.1 (a).

Circuit Type: Negative resistance, crystal-controlled or free-running.

Transistor Class: Point-contact only.

Approximate Frequency: 0 to alpha cut-off (50 kc).

Stability: \( S_v = 0.01 \) for \( E_e \), \( S_v = 0.03 \) for \( E_c \).

Transistor Types: Any point-contact.

Design Procedure: Linear circuit theory or Groszkowski analysis.

Power Output: Voltage only, unless a transformer is used in the collector circuit.

Desired Transistor Characteristics: High alpha, constant \( h \) parameters.

Notes:

The notes for this circuit are the same as for Figure 7.1.

This circuit is not recommended as a stable oscillator.
Prototype: 4.1 (a).

Circuit Type: Negative resistance, crystal-controlled only.

Transistor Class: Point-contact only.

Approximate Frequency: 0 to alpha cut-off (1 mc).

Stability: $S_v = 7$ for $E_c$.

Transistor Types: Any point-contact.

Design Procedure: Linear circuit theory.

Power Output: Voltage only; low power is available by transformer coupling to collector circuit.

Desired Transistor Characteristics: High alpha, constant alpha and $h$ parameters.

Notes:

This circuit is a modification of Figure 7.3 to permit high-impedance crystal control and to include bias compensation. With low-impedance crystals, the LC circuit in the base may be replaced by a resistor.

The temperature stability of this circuit is good.

Because the circuit values must be matched to particular transistors, this circuit is not recommended for general use.
Prototype: 4.1 (c).

Circuit Type: Negative resistance, free-running.

Transistor Class: Point-contact only.

Approximate Frequency: 0 to greater than alpha cut-off.

Stability: Depends on transistor and frequency.

Transistor Types: Any point-contact.

Design Procedure: Linear circuit theory or Groszkowski analysis.

Power Output: Voltage only.

Desired Transistor Characteristics: High alpha, constant h parameters.

Notes:

The circuit shown is a general circuit type which has been investigated extensively. Some of the circuits on the following pages are practical modifications of this circuit. This basic circuit has been operated at frequencies from 1 kc to 95 mc with varying degrees of frequency stability. More data on this circuit may be found on the following pages and in the appendix.
Figure 7.6. 25-kc Oscillator Using Diode Amplitude Limiting.

Prototype: 4.1 (c).
Circuit Type: Negative resistance, free-running.
Transistor Class: Point-contact only.
Approximate Frequency: 0 to much less than alpha cut-off (25 kc).
Stability: $S_v = 1$
Transistor Types: Any point-contact.
Design Procedure: Linear circuit theory or Groszkowski analysis.
Power Output: Constant low voltage only.
Desired Transistor Characteristics: High alpha, constant, nonreactive $h_{pa}$ parameters.

Notes:

This circuit is a distortion-stabilized free-running modification of Figure 7.5. The stability obtained has, in some cases, approached that of a crystal-controlled oscillator. However, this degree of stability is possible only at frequencies very low as compared to the alpha cut-off frequency of the transistor. The available output voltage may be increased by removing the diodes, but this will reduce the stability by a factor of 5 or more.

The following points are of importance in designing this oscillator:

1. Better stability is obtained with high values of $Q$ although a $Q$ greater than 15 or 20 does not produce appreciable improvements. An optimum $Q$ does not exist.
2. Best stability is obtained with low values of resonant impedance (5,000 to 10,000 ohms). It should be noted that the resonant impedance of the tuned circuit must be greater than the negative resistance presented by the base of the transistor, otherwise, parallel-mode oscillations will not be obtained.

3. Higher values of $R_e$ usually produce the best stability while at the same time reducing the tendency to oscillate. Typical values for $R_e$ range from 2000 ohms to 8000 ohms. An optimum value of $R_e$ does exist in most instances.

Curves showing the effect of $R_e$ on the frequency stability are presented in Figure 7.7. The proper choice of $R_e$ for the particular transistor and tuned circuit would be near 4700 ohms as may be seen from the curves. It should be noted that these curves apply only to a particular transistor and tuned circuit. However, the shape of the curves is generally the same for other tuned circuits and transistors.

Figure 7.8 shows a typical set of curves for a frequency near the alpha cut-off of the transistor. The stability is no longer good over a wide range of supply voltage.

This circuit is recommended where a stable free-running oscillator is required at low frequencies.
Figure 7.7. Frequency Variation as a Function of Supply Voltage for Different Values of $R_e$. 

-133-
Figure 7.8. Frequency Variation as a Function of Supply Voltage for Different Values of $R_e$.
Figure 7.9. Impedance-Inversion Oscillator.

Prototype: 4.1 (c).
Circuit Type: Negative resistance, crystal-controlled only.
Transistor Class: Point-contact only.
Approximate Frequency: 0 to less than alpha cut-off.
Transistor Types: Any point-contact.
Design Procedure: Linear circuit theory or Groszkowski analysis.
Power Output: Low voltage only.
Desired Transistor Characteristics: High alpha, constant alpha and h parameters.

Notes:

This circuit is a modification of Figure 7.5 to permit crystal control and distortion stabilization. The circuit uses an artificial transmission line to obtain impedance inversion so that the crystal may be operated in its series mode. The d-c base current is obtained through the diodes. Spurious resonances are likely to result in this circuit unless extreme care is used on constructing the artificial line.

The temperature stability of this circuit is fair.

This circuit is not recommended because the artificial line is too bulky and because the line parameters must be made variable to accommodate different frequencies.
Figure 7.10. 11-mc Free-Running Oscillator.

Prototype: 4.1 (c).

Circuit Type: Negative resistance, free-running.

Transistor Class: Point-contact only.

Approximate Frequency: 0 to 10 times alpha cut-off (11 mc).

Stability: Very poor at high frequencies.

Transistor Types: Any point-contact.

Design Procedure: Linear circuit analysis or Groszkowski analysis.

Power Output: Low voltage output only at high frequencies.

Desired Transistor Characteristics: High alpha at high frequencies, constant alpha and h parameters at high frequencies.

Notes:

This circuit was designed to permit operation of transistors at frequencies well beyond their alpha cut-off. The circuit will operate only as long as the power gain of the transistor remains high. The collector inductance and emitter capacitance are best chosen experimentally for the highest operating frequency.

Both temperature and voltage stability are poor for this circuit.

Bias compensation is useful at low frequencies but is not successful at frequencies beyond about twice alpha cut-off.

Operation at the highest frequencies is possible only by hand picking the transistors.

This circuit is recommended for use only when operating a point-contact transistor far beyond alpha cut-off.
Prototype: 4.1 (c).

**Circuit Type:** Negative resistance, free-running only.

**Transistor Class:** Point-contact only.

**Approximate Frequency:** 20 to 95 mc with an RCA 2N33.

**Stability:** Quantitative data not available.

**Transistor Types:** 2N33.

**Design Procedure:** Empirical.

**Power Output:** Low voltage only at higher frequencies.

**Desired Transistor Characteristics:** High alpha, constant h parameters.

**Notes:**

This circuit is the same as Figure 7.5 except that particular component values are shown for the operation of the 2N33 at frequencies as high as 95 mc. L and C must be carefully chosen experimentally to obtain oscillations at this high frequency. The Q of the tuned circuit must be high and the parallel resonant impedance must be correct for the particular transistor (between 1000 and 10,000 ohms).

The stability of this circuit was not measured because of the lack of suitable equipment. The stability was, however, observed to be relatively poor by means of a high-frequency receiver.

This circuit is recommended only when very high frequency operation is necessary without regard to stability.
Prototype: 4.2 (h).

Circuit Type: Feedback, crystal-controlled.

Transistor Class: Point-contact or junction.

Approximate Frequency: 0 to greater than alpha cut-off (50 kc).

Stability: $S_v = 3$ for $E_c$, $S_v = 1$ for $I_e$.

Transistor Types: Any transistor.

Design Procedure: Empirical.

Power Output: Low power output is available if a third winding is added to the transformer.

Desired Transistor Characteristics: Constant alpha, small transit time, constant h parameters.

Notes:

This circuit was designed primarily to obtain crystal control with a high-impedance crystal. Therefore, it was necessary to place the crystal in the primary side of the transformer. Reasonable impedance match for very high impedance crystals (greater than 10,000 ohms) is possible in this position.

The temperature stability is fair.

Diode distortion stabilization may be used with this circuit to improve the frequency stability.

This circuit is not recommended except when the use of a very high impedance crystal is necessary at low frequencies. In this case, the stability can be greatly improved by using a modification of the general design procedure for linear oscillators as presented in section II-E.
Figure 7.13. Crystal-Controlled Oscillator with Transformer Feedback and Bias Compensation.

Prototype: h.2 (h).

Circuit Type: Feedback, crystal-controlled.

Transistor Class: Point-contact or junction.

Approximate Frequency: 0 to greater than alpha cut-off (50 kc).

Stability: $S_v = 3$ for $E_C$.

Transistor Types: Any transistor.

Design Procedure: Empirical.

Power Output: Low power output is available if a third winding is added to the transformer.

Desired Transistor Characteristics: Constant alpha, small transit line, constant $h$ parameters.

Notes:

The notes pertaining to the circuit of Figure 7.12 apply to this circuit. The resistor network shown was used to provide bias compensation. No over-all improvement resulted, but the stability was improved over some voltage ranges. The temperature stability was improved slightly by the addition of the bias network.

This circuit is not recommended for general use.
Prototype: 4.2 (h)

Circuit Type: Feedback, crystal-controlled.

Transistor Class: Point-contact only.

Approximate Frequency: 0 to 5 times alpha cut-off (1 mc).

Stability: $S_V = 6$ for $E_e$, $S_V = 2$ for $E_c$.

Transistor Types: Any point-contact.

Design Procedure: Empirical.

Power Output: Low power output is available by transformer coupling into the base circuit.

Desired Transistor Characteristics: Constant alpha, small transit time, constant $h$ parameters.

Notes:

This circuit is another modification of Figure 7.12. It makes possible operation at higher frequencies by using a combination of negative resistance and feedback. The higher frequencies permit the use of lower impedance crystals thus making it possible to place the crystal in the emitter circuit. Better impedance match is obtained in this configuration. No oscillations are obtained when the crystal is removed if the resonant impedance of the base circuit is low.

The temperature stability for this circuit is fair.

The frequency stability may be improved by distortion stabilization.

This circuit is not recommended for general use.
Prototype: 4.2 (h).

Circuit Type: Feedback, crystal-controlled.

Transistor Class: Point-contact only.

Approximate Frequency: 0 to 5 times alpha cut-off (1 mc).

Stability: $S_v = 3$ for $E_c$.

Transistor Types: Any point-contact.

Design Procedure: Empirical.

Power Output: Low power output is available by transformer coupling into the base circuit.

Desired Transistor Characteristics: Constant alpha, small transit time, constant $h$ parameters.

Notes:

This circuit is a modification of Figure 7.14 to permit single-battery operation. Bias compensation was not possible in this case. Furthermore, the addition of the bias network actually decreased the stability. This was caused primarily by the presence of both negative resistance and feedback. The temperature stability was greatly improved by the addition of the bias network.

This circuit is not recommended for general use.
Figure 7.16. Transformer Coupled 400-kc Oscillator.

Prototype: 4.2 (c).
Circuit Type: Feedback, free-running or crystal-controlled.
Transistor Class: Point-contact or junction.
Approximate Frequency: 0 to less than alpha cut-off (400 kc).
Stability: $S_v = 80$.
Transistor Types: Any transistor.
Design Procedure: Empirical.
Power Output: Power output available by adding a third winding to the transformer.
Desired Transistor Characteristics: Constant alpha, low transit time, non-reactive $h$ parameters.

Notes:

This circuit was designed to permit operation of the crystal exactly at the resonant frequency of the series branch. This is accomplished by using the transistor at low frequencies and effectively canceling the crystal shunt capacitance by $C$. Distortion stabilization with diodes was tried with this circuit but was unsuccessful because of spurious oscillations.

The temperature stability of this circuit is good.

This circuit is recommended for obtaining very good stability from a point-contact transistor oscillator.
Prototype: 4.2 (a).

Circuit Type: Feedback, free-running or crystal-controlled.

Transistor Class: Point-contact only.

Approximate Frequency: 0 to alpha cut-off (400 kc)

Stability: $S_v \approx 25$.

Transistor Types: Any point-contact transistor with high alpha.

Design Procedure: Empirical.

Power Output: Voltage only.

Desired Transistor Characteristics: High alpha, low transit time, nonreactive $h$ parameters.

Notes:

The transistor must provide current gain greater than unity in order for this circuit to oscillate. Bias compensation was used in the design. It was found that the bias compensation must be designed for each particular sample of the WE 1698 transistor. Poor stability is likely to result when the transistor or crystal is replaced. Compensation is good for only a narrow frequency range. Figure 7.18 shows the effect of crystal impedance on compensation.

The temperature stability for this circuit is fair.

This circuit is not recommended for general use.
Figure 7.18. Change in Frequency for a Given Change in Voltage as a Function of Crystal Resistance.
Prototype: 4.2 (a).

Circuit Type: Feedback, free-running or crystal-controlled.

Transistor Class: Point-contact only.

Approximate Frequency: 0 to less than alpha cut-off (400 kc).

Stability: $S_v = 20$-$100$, $S_t = 25$.

Transistor Types: Point-contact transistors with high alpha.

Design Procedure: Empirical.

Power Output: Low voltage only.

Desired Transistor Characteristics: High alpha, low transit time, nonreactive $h$ parameters.

Notes:

This circuit is a modification of the circuit of Figure 7.17. Distortion stabilization is incorporated. Although the best stability obtainable with some transistors is reduced, the average stability with several samples of the same transistor type is improved. Only the more active WE 1729 transistors operated properly in this circuit (alpha greater than 2 at operating frequency).

The temperature stability for this circuit is good.

This circuit is recommended for good stability and transistor replaceability at frequencies where the diode shunt capacitance is not too great.
Prototype: 7.2 (e).
Circuit Type: Feedback, free-running or crystal-controlled.
Transistor Class: Junction.
Approximate Frequency: 0 to greater than alpha cut-off (30 mc).
Stability: Quantitative data not available.
Transistor Types: L-5100.
Design Procedure: Empirical.
Power Output: Very low voltage only.
Desired Transistor Characteristics: Constant alpha, nonreactive h parameters.

Notes:

Circuits similar to this have been shown previously. This particular circuit is included because it is the circuit with which the highest frequency has been obtained using a junction transistor. The stability at the highest frequency (30 mc) was not measured because of the very low voltage output and the lack of suitable equipment. The stability appeared to be low.

This circuit is not recommended except where very high frequency operation is necessary.
Prototype: 4.2 (e).

Circuit Type: Feedback, free-running or crystal-controlled.

Transistor Class: Point-contact or junction.

Approximate Frequency: 0 to greater than alpha cut-off (600 kc).

Stability: See Figure 6.6.

Transistor Types: Any.

Design Procedure: Vacuum-tube analogy.

Power Output: Voltage only.

Desired Transistor Characteristics: Low transit time, constant phase shift.

Notes:

This circuit is a modified Hartley oscillator with distortion compensation. The position of the diode is such that the amplitude of oscillation is increased rather than decreased. R can be varied to control the amount of distortion and, thus, the amount of frequency correction. For curves on this oscillator, see section VI, Figure 6.6.

This oscillator may be crystal-controlled by replacing the feedback capacitor with a crystal.
Prototype: 4.2 (e).

Circuit Type: Feedback, free-running or crystal-controlled.

Transistor Class: Point-contact or Junction.

Approximate Frequency: 0 to greater than alpha cut-off.

Stability: See Figure 7.23.

Transistor Types: CK 720, 2N44, L-5100 (See Notes).

Design Procedure: Vacuum-tube analogy.

Power Output: Voltage only.

Desired Transistor Characteristics: Low transit time, constant phase shift.

Notes:

This circuit is a modification of the circuit of Figure 7.21. The notes to Figure 7.21 apply to this circuit.

This transistor operates with practically zero d-c base current. Therefore, transistors having power gain under this condition must be chosen.

The performance of this circuit is shown in Figure 7.23.
Figure 7.23. Performance Curves for Circuit of Figure 7.22.
Prototype: 4.2 (c).

Circuit Type: Feedback, crystal-controlled only.

Transistor Class: Junction only.

Approximate Frequency: 0 to less than alpha cut-off (1 mc).

Stability: $S_v = 10$ (for changes in collector voltage).

Transistor Types: Any junction transistor.

Design Procedure: Empirical.

Power Output: Appreciable power output available, depending upon the type of transistor used.

Desired Transistor Characteristics: High power gain, low phase shift, constant collector capacitance.

Notes:

This oscillator was designed for comparison with three circuits which are to follow. The following steps were taken in the design of this oscillator:

1. The Philco L-5100 transistor was selected because of its high power gain at 1000 kc.

2. The grounded emitter configuration was selected because of its high current gain. The amplifier was designed empirically, keeping in mind the

See section II-E for meanings of symbols.
dissipation limitations of the transistor. The input impedance was kept high to make use of maximum power gain. The input impedance was found to be 1.6K by measurement. The output impedance, $r_o$, for the amplifier and transformer was about 125 ohms. From equation 2.67, $\rho_p$ was found to be 0.055.

(3) The only requirements placed on the transformer were that it provide the necessary phase inversion and the necessary current gain to obtain oscillations.

(4) A crystal was connected as shown to operate in its series mode.

(5) The stability, $S_v$, for changes in collector voltage, was found experimentally to be 10.
Prototype: 4.2 (c).
Circuit Type: Feedback, crystal-controlled only.
Transistor Class: Junction only.
Approximate Frequency: 0 to less than alpha cut-off (1 mc).
Stability: $S_v = 80$ (for changes in collector voltage).
Transistor Types: Any junction transistor.
Design Procedure: Crystal impedance match.
Power Output: Appreciable power output available, depending upon the type of transistor used.
Desired Transistor Characteristics: High power gain, low phase shift, constant collector capacitance.

Notes:

This oscillator uses the same circuit as Figure 7.24.

The design procedure was as follows:

(1) The same amplifier and the same transistor were used as in Figure 7.24 so as to give a fair comparison of the stability with and without impedance match. $K'$ and $r_o'$ for the amplifier were found to be 167 and 15,000 ohms respectively. The turns ratio of the transformer was selected so that the reflected primary impedance would be 15,000 ohms while at the same time maintaining oscillations with a minimum value of $R$. The actual measured value of $r_o$ was 125 ohms. This differed from the theoretical value of 150 ohms because of the transformer losses. The value of $\rho_\phi$ was found to be 0.145.

†See section II-E for meanings of symbols.
(2) The crystal was connected as shown with $R$ equal to 17 ohms.

(3) The stability, $S_v$, for changes in collector voltage was found experimentally to be 80.

Proper impedance matching of the crystal resulted in an increase in the stability from 10 to 80.
Prototype: 4.2 (c).

Circuit Type: Feedback, crystal-controlled only.

Transistor Class: Junction only.

Approximate Frequency: 0 to less than alpha cut-off (1 mc).

Stability: $S_v = 125$ (for changes in collector voltage).

Transistor Types: Any junction transistor.

Design Procedure: General design procedure from section II-E.

Power Output: Appreciable power output available, depending upon the type of transistor used.

Desired Transistor Characteristics: High power gain, low phase shift, constant collector capacitance.

Notes:

Again the same circuit was used as shown in Figures 7.24 and 7.25. In this case, however, Figure 2.37 was used for the design. The procedure was as follows:

(1) The same amplifier and transistor were used as in the previous examples. $K/r_o$ was found to be 1.86.

(2) From Figure 2.37, $r_o$ was found to be 14 ohms and $R$ to be 28 ohms. The required turns ratio of the transformer, $K'/K$, was 33. The value of $p_0$ was 0.71.

† See section II-E for details.
(3) The constants actually used in the circuit were as shown in Figure 7.26.

(4) The voltage stability was found experimentally to be greater than 125.
Prototype: 4.3 (c).

Circuit Type: Feedback, crystal-controlled only.

Transistor Class: Junction only.

Approximate Frequency: 0 to less than alpha cut-off (1 mc).

Stability: $S_v = 250$ (for changes in collector voltage).

Transistor Types: Any Junction transistor.

Design Procedure: General design procedure from section II-E.

Power Output: Appreciable power output available, depending upon the type of transistor used.

Desired Transistor Characteristics: High power gain, low phase shift, constant collector capacitance.

Notes:

This circuit was designed by using the curves of Figure 2.39 as follows:

1. The same amplifier and transistor were used as in the previous examples. $K^2/r_o$ was again found to be 1.86.

2. From Figure 2.39, $r_3$ was found to be 60 ohms and $R$ to be 145 ohms. The required turns ratio of the transformer, $K'/K$, was 16. The value of $\rho_{\phi}$ was 2.4.

3. The constants actually used in the circuit were as shown in Figure 7.27. The choice of components was limited by the available transformer.

4. The voltage stability was found experimentally to be greater than 250.
This circuit was tested with 11 L-5100 transistors in addition to the original design transistor. The stability was about the same for each transistor. With no circuit or voltage changes, the following table shows the actual frequency of oscillation for each of the 12 transistors.

<table>
<thead>
<tr>
<th>Transistor No.</th>
<th>Frequency</th>
<th>Transistor No.</th>
<th>Frequency</th>
</tr>
</thead>
<tbody>
<tr>
<td>66</td>
<td>999903.4</td>
<td>94</td>
<td>999903.6</td>
</tr>
<tr>
<td>83</td>
<td>999903.1</td>
<td>95</td>
<td>999903.4</td>
</tr>
<tr>
<td>84</td>
<td>999903.6</td>
<td>96</td>
<td>999903.2</td>
</tr>
<tr>
<td>85</td>
<td>999903.2</td>
<td>97</td>
<td>999903.6</td>
</tr>
<tr>
<td>86</td>
<td>999903.7</td>
<td>98</td>
<td>999903.6</td>
</tr>
<tr>
<td>87</td>
<td>999903.8</td>
<td>99</td>
<td>999903.7</td>
</tr>
</tbody>
</table>

The maximum frequency difference was 0.7 cycle per second. The frequency of oscillation could have been adjusted to be the same in each case by varying the capacitance across the primary of the transformer.

In this circuit, as well as in the preceding three circuits, no attempt was made to improve the performance of the amplifier (that is, to reduce $d\delta/d\sigma$ for the amplifier). The object of the designs was to evaluate the general design procedure outlined in section II-E. By the use of this procedure, the voltage stability was improved from 10 to a maximum of over 250. The best stability was probably greatly in excess of 250, however, equipment for measuring such stabilities was not available at the time. The improvements in stability were approximately as predicted from the ratios of the values of $\rho_0$.

It is reasonable to assume that further improvements in stability can be accomplished by improving the phase characteristics of the amplifier. Also, the networks chosen in section II-E are not necessarily the best ones that can be found. It should be emphasized that this is a general design procedure which can be applied to any feedback network and any amplifier design.

In measuring the stability of this circuit, it was found that the basic stability of the oscillator was much better than the temperature stability of any crystals available to the project. The crystals were also of relatively low $Q$, which greatly reduced the voltage stability of the oscillators. With better crystals, this circuit should give stabilities in excess of 1000 (1 part in $10^9$ for one per cent voltage change).

This circuit is recommended for maximum stability from a single transistor oscillator.
Prototype: 4.3 (a).
Circuit Type: Feedback, free-running or crystal-controlled.
Transistor Class: Point-contact or junction (preferably junction).
Approximate Frequency: 0 to greater than alpha cut-off (400 kc).
Transistor Types: CK 720, 2N44, L-5100 (See Notes).
Design Procedure: Empirical.
Power Output: Voltage only.
Desired Transistor Characteristics: Transistor must have power gain at zero d-c base current.

Notes:

This circuit obtains current gain for oscillation by means of the resonant rise in current in a resonant circuit. The type of transistor used must be capable of supplying power gain when operating with zero base current. The resonant rise in current occurs between the base and the emitter.

The temperature stability of this circuit is good.

This circuit is not recommended for general use.
Prototype: 4.3 (a).

Circuit Type: Feedback, free-running or crystal-controlled.

Transistor Class: Point-contact or junction (preferably junction).

Approximate Frequency: 0 to greater than alpha cut-off (400 kc).

Stability: See Figure 7.30.

Transistor Types: CK 720, 2N44, L-5100 (See Notes).

Design Procedure: Empirical.

Power Output: Voltage only.

Desired Transistor Characteristics: Transistor must have power gain at zero d-c base current.

Notes:

This circuit is a simplified version of the circuit of Figure 7.28. The stability is reduced by the simplification. Also, the circuit does not perform as well with different transistors of the same type. (See Figure 7.30.)

This circuit is not recommended for general use.
Figure 7.30. Frequency-Versus-Voltage Characteristics for Several CK 720 Transistors.
Prototype: 4.3 (b).
Circuit Type: Feedback, free-running or crystal-controlled.
Transistor Class: Point-contact or junction (preferably junction).
Approximate Frequency: 0 to greater than alpha cut-off (400 kc).
Stability: Quantitative data not available.
Transistor Types: CK 720, 2N44, L-5100 (See Notes).
Design Procedure: Empirical.
Power Output: Voltage only.
Desired Transistor Characteristics: Transistor must have power gain at zero d-c base current.

Notes:

This circuit is another modification of Figure 7.28. The performance is almost identical to that of Figure 7.30.

This circuit is not recommended for general use.
Figure 7.32. Zero-Base-Current Oscillator With External Distortion.

Prototype: 4.3 (b).

Circuit Type: Feedback, free-running or crystal-controlled.

Transistor Class: Point-contact or junction (junction preferred).

Approximate Frequency: 0 to greater than alpha cut-off (400 kc).

Stability: See Figure 7.33.

Transistor Types: CK 720, 2N44, 1N67 (See Notes).

Design Procedure: Empirical.

Power Output: Low voltage only.

Desired Transistor Characteristics: Transistor must have power gain at zero d-c base current.

Notes:

This circuit is a free-running version of Figure 7.28. Amplitude and distortion stabilization are incorporated in this circuit. The output amplitude is greatly reduced by the diodes.

The temperature stability of this circuit is good.

This circuit is recommended as a free-running oscillator when the low output voltage from a high power supply voltage is satisfactory. See Figure 7.33 for performance curves. The circuit operates equally well with crystal control.
Figure 7.33. Performance Curves for Circuit of Figure 7.32.
Figure 7.34. Zero-Base-Current Oscillator With Controllable Distortion.

Prototype: 4.3 (b).

Circuit Type: Feedback, free-running or crystal-controlled.

Transistor Class: Point-contact or junction (junction preferred).

Approximate Frequency: 0 to greater than alpha cut-off (400 kc).

Stability: See Figure 7.35.

Transistor Types: CK 720, 2N44, L-5100 (See Notes).

Design Procedure: Empirical.

Power Output: Voltage only.

Desired Transistor Characteristics: Transistor must have power gain at zero d-c base current.

Notes:

This circuit shows another method of applying distortion stabilization to the circuit of Figure 7.28. This method does not reduce the output amplitude. Also, the distortion can be controlled by R.

The temperature stability of this circuit is fair.

This circuit is recommended as a free-running or crystal-controlled oscillator when a high power supply voltage is satisfactory. See Figure 7.35 for performance curves.
Figure 7.35. Performance Curves for Circuit of Figure 7.34.
Prototype: 4.2 (b).

Circuit Type: Feedback, free-running.

Transistor Class: Point-contact or junction (junction preferred).

Approximate Frequency: Only beyond alpha cut-off.

Stability: Quantitative data not available.

Transistor Types: Any.

Design Procedure: Empirical.

Power Output: Voltage only.

Desired Transistor Characteristics: Alpha constant with respect to temperature and voltage beyond alpha cut-off.

Notes:

This circuit is designed to make use of large transistor phase shifts (that is, high-frequency operation). For a transistor with characteristics as shown in Figure 7.37, oscillations at frequencies as high as 2 mc have been obtained. The stability, however, was not nearly as great as that of other circuits at lower frequencies.

This circuit is only moderately sensitive to temperature variations.

This circuit is recommended for high-frequency operation of junction transistors as free-running oscillators where stability is not of primary importance. The stability may be improved by the application of bias compensation.
Figure 7.37. Alpha Characteristics of a Typical Type CK 720 Junction Transistor.
Prototype: 4.2 (b).

Circuit Type: Feedback, crystal-controlled.

Transistor Class: Point-contact or junction (junction preferred).

Approximate Frequency: Only beyond alpha cut-off (400 kc).

Stability: See Figure 7.39.

Transistor Types: Any.

Design Procedure: Empirical.

Power Output: Voltage only.

Desired Transistor Characteristics: Alpha constant with respect to temperature and voltage beyond alpha cut-off.

Notes:

This circuit is a modification of Figure 7.36 to include crystal control. Figure 7.39 shows the frequency variations for changes in emitter and collector voltage.

This circuit is not recommended for general use.
Figure 7.39. Variation of Frequency With Collector Supply Voltage for Various Values of Emitter Supply Voltage.
Prototype: 4.2 (b).

Circuit Type: Feedback, crystal-controlled.

Transistor Class: Point-contact or junction (junction preferred).

Approximate Frequency: Only beyond alpha cut-off (400 kc).

Stability: $S_v = 10$, See Figure 7.41.

Transistor Types: Any.

Design Procedure: Empirical.

Power Output: Voltage only.

Desired Transistor Characteristics: Alpha constant with respect to temperature and voltage beyond alpha cut-off.

Notes:

From Figure 7.39, it can readily be seen that the circuit in Figure 7.38 can be bias-compensated. Figure 7.40 is the bias-compensated version of the circuit of Figure 7.38. Figure 7.41 shows the resulting variation in frequency with supply voltage. The variation in frequency with temperature is shown in Figure 7.42.

This circuit is recommended for high-frequency operation of junction transistors as crystal-controlled oscillators. It is possible, however, that the bias compensation may have to be adapted to each individual transistor.
Figure 7.41. Variation of Frequency With Supply Voltage for Two Values of $R$. 

$R = 560\Omega$

$R = 890\Omega$
Figure 7.42. Frequency-Versus-Temperature Characteristics for Circuit of Figure 7.40.
Prototype: 4.2 (a).

Circuit Type: Feedback, crystal-controlled only.

Transistor Class: Junction.

Approximate Frequency: 0 to less than alpha cut-off (1 mc).

Stability: $S_v = 7$.

Transistor Types: Any junction.

Design Procedure: Vacuum-tube analogy.

Power Output: Voltage only.

Desired Transistor Characteristics: Transistor must provide power gain with low d-c base current.

Notes:

This circuit is a transistorized version of the vacuum-tube Pierce oscillator. The circuit will operate over a wide range of crystal frequencies without readjustment of component values.

The temperature stability is fair.

This circuit is recommended when a wide-range low-voltage oscillator is required. Distortion stabilization cannot be used if the L-5100 transistor is employed. However, bias compensation may be helpful in improving the stability.
Prototype: 4.3 (d).
Circuit Type: Feedback, crystal-controlled only.
Transistor Class: Junction.
Approximate Frequency: 0 to less than alpha cut-off (1 mc).
Stability: $S_v = 3$.
Transistor Types: L-5100.
Design Procedure: Vacuum-tube analogy.
Power Output: Voltage only.
Desired Transistor Characteristics: Transistor must provide power gain at zero d-c base current.

Notes:

This circuit is a simplification of the one in Figure 7.43. The stability is reduced, but the frequency range of operation is about the same. This circuit will operate at power input levels as low as 0.001 milliwatt. If a low-impedance crystal is used and if the impedance of the voltage supply source is high, the resistor may be omitted from the circuit.

The temperature stability of this circuit is fair.

This circuit is not recommended except where the ultimate in simplicity is required.
Prototype: 4.2 (a).

Circuit Type: Overtone feedback, crystal-controlled only.

Transistor Class: Junction.

Approximate Frequency: 0 to less than alpha cut-off (3 mc).

Stability: $S_v = 35$ (collector not tuned to exact resonance).

Transistor Types: L-5100, CK 720, 2N44.

Design Procedure: Empirical.

Power Output: Voltage only.

Desired Transistor Characteristics: Low collector-base capacitance, high alpha.

Notes:

This circuit has been operated successfully as an overtone oscillator at the third, fifth and seventh overtones. In most cases, the stability has been better at the overtones than at the fundamental. The crystals used were not designed for overtone operation. Best stability is obtained when the collector tuned circuit is tuned away from resonance until the amplitude begins to decrease appreciably.

This circuit is recommended as an overtone oscillator.
Figure 7.46. 1000-kc Meacham Bridge Transistor Oscillator.
Prototype: Vacuum-tube Meacham bridge oscillator.
Circuit Type: Feedback, crystal-controlled only.
Transistor Class: Point-contact.
Approximate Frequency: 0 to less than alpha cut-off (1 mc).
Stability: $S_v = 500, S_t = 100, S = 3.3.$
Transistor Types: Any point-contact with high gain at operating frequency.
Design Procedure: Vacuum-tube analogies.
Power Output: Voltage only with circuit as shown.
Desired Transistor Characteristics: High gain at operating frequency, low phase shift.

Notes:

This circuit was patterned after the familiar two-stage Meacham bridge vacuum-tube oscillator. The amplifier was designed to obtain the best impedance match at the output and between the stages. The output impedance of the amplifier was adjusted to approximately 100 ohms. The input impedance of the amplifier was found to be greater than 100 ohms. The over-all voltage gain of the amplifier was measured to be 50. By using an equal-arm bridge, the bridge negative-feedback factor was 0.5. This gave an over-all phase-reduction factor of about 25.

One of the major difficulties encountered in designing the oscillator was reducing the gain of the amplifier to less than unity at the frequency where the phase shift first reached 180 degrees. If this were not done, the circuit would oscillate on the negative-feedback path of the bridge. The crystal, $X_2,$ was added to the circuit as a bandpass filter to prevent this type of oscillation. The crystal, $X_1,$ is approximately 25 times more effective in controlling the frequency than the crystal $X_2.$ It is necessary, however, that the two crystals be series resonant at very nearly the same frequency.

A vacuum-tube Meacham bridge oscillator was constructed for comparison with the transistor oscillator. The characteristics of the two circuits are presented in the table on the following page.

The operating conditions for the two oscillators were made as nearly the same as possible. The same crystal and oven were used. Room-temperature variations were approximately the same while the oscillators were under test.

The volume of space required for the vacuum-tube oscillator was many times that required for the transistor oscillator. The power required for the vacuum-tube oscillator was 130 times that required for the transistor oscillator. The voltage stability for the two oscillators was approximately the same. The short-time stability for the vacuum-tube oscillator was slightly better than for the transistor oscillator.
<table>
<thead>
<tr>
<th>Item to be Compared</th>
<th>Transistor Oscillator</th>
<th>Vacuum-Tube Oscillator</th>
</tr>
</thead>
<tbody>
<tr>
<td>Number and type of amplifying element used</td>
<td>two WE 1729</td>
<td>one 12BY7</td>
</tr>
<tr>
<td></td>
<td>crystal in JK03 oven</td>
<td>one 6S4</td>
</tr>
<tr>
<td>Type of frequency control</td>
<td>1000 kc</td>
<td>1000 kc</td>
</tr>
<tr>
<td>Frequency</td>
<td>3 v P. to P. to high</td>
<td>3 v P. to P. to high</td>
</tr>
<tr>
<td></td>
<td>impedance</td>
<td>impedance</td>
</tr>
<tr>
<td>Output</td>
<td>nine 1.5-v batteries</td>
<td>200-v, d-c regulated</td>
</tr>
<tr>
<td></td>
<td></td>
<td>supply</td>
</tr>
<tr>
<td>Power supply</td>
<td>0.12 watt exclusive</td>
<td>16 watts exclusive of</td>
</tr>
<tr>
<td></td>
<td>of oven</td>
<td>oven</td>
</tr>
<tr>
<td>Power requirements</td>
<td>± 25%</td>
<td>± 25%</td>
</tr>
<tr>
<td>Permissible voltage variation</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Frequency-versus-voltage stability</td>
<td>$S_v = 500$</td>
<td>$S_v = 500$</td>
</tr>
<tr>
<td>Short-time stability</td>
<td>$S = 2.5$</td>
<td>$S = 3.3$</td>
</tr>
</tbody>
</table>

The point-contact transistor version of the Meacham bridge oscillator is not recommended because of the difficulty to obtain freedom from spurious oscillations. It does, however, perform very satisfactorily once it has been properly adjusted.
Figure 7.47. Junction-Transistor Meacham Bridge Oscillator.
Prototype: Vacuum-tube Meacham bridge oscillator.

Circuit Type: Feedback, crystal-controlled only.

Transistor Class: Junction.

Approximate Frequency: 0 to less than alpha cut-off (1 mc).

Stability: $S_v = 300, S_t = 200, S = 10, S(10 \text{ minutes}) = 100$.

Transistor Types: RDX-300A and TI 904A.

Design Procedure: See Notes.

Power Output: 2-mw maximum.

Desired Transistor Characteristics: High gain, low phase shift, reasonably constant phase shift and gain.

Notes:

Several junction-transistor amplifiers were designed for tentative use in this circuit. The transistors which were tried are the RD-2525A, the CK 720, the RCA 2N34 and the GE ZJ1. The maximum gain obtained at 1000 kc for a two-transistor amplifier was 26 db. The gain for this amplifier, which used the RD-2525A transistors, was 8 db lower than the gain of the amplifier shown in Figure 7.46. The gain also varied greatly with the supply voltage. The greatest power output was obtained from a circuit designed for use with the GE ZJ1. However, the gain for two GE ZJ1 transistors was only 12 db. The circuits designed for use with the CK 720 and the RCA 2N34 gave less gain than the RD-2525A circuit and less power output than the GE ZJ1 circuit. Thus, none of the junction amplifier circuits using the transistors mentioned above had characteristics desirable for use in the Meacham bridge 1000-kc oscillator.

For the Meacham bridge circuit, the power output required from the amplifier may range from a minimum of 4 mw to a maximum of about 40 mw, depending on the type of lamp chosen for use as the amplitude regulator. With lamps which were available, 6 mw was considered to be the minimum power which would give reliable bridge operation. After an investigation of all of the junction transistors available to the project, it was concluded that only the RDX-300A tetrode transistor would provide sufficient power at 1000 kc and at the same time provide a reasonable gain. An amplifier was designed using a RDX-300A and an L-5100. However, the maximum power output from the L-5100 was not sufficient to supply the losses of the coupling transformer and drive the RDX-300A.

The circuit that was finally developed is shown in Figure 7.47. The TI 904A and the RDX-300A were the only transistors found suitable for this application among the junction transistors available. The gain of this amplifier is approximately the same as the gain of the amplifier shown in Figure 7.46. The rate of change in phase for changes in supply voltage is slightly greater than for the point-contact amplifier. This accounts for the slightly worse voltage stability. Bias compensation which occurred when the point-contact amplifier was operated from a single power supply did not occur for the junction amplifier. Bias compensation could, of course, be applied to the junction amplifier by using Varistors.
The over-all operation of the junction-transistor Meacham bridge oscillator is considered superior to that of the point-contact circuit for two reasons. First, the short-time stability was considerably better (1 part in $10^8$ for 10 minutes). Second, an isolated output was readily obtained from the second base of the RDX-300A. It was necessary, however, to add an amplifier stage as shown to provide a two-volt rms output for the particular application for which the oscillator was designed (frequency standard for the Berkeley Model 5570 Frequency Meter).

The crystal used with this circuit was series resonant at 999,891 cycles per second. In order to measure the stability, it was necessary to make this crystal operate at 1,000,000 cycles per second. This was accomplished by placing a 30.5-micromicrofarad capacitor in series with the crystal. The stability data given above for this oscillator are taken with the capacitor in series with the crystal. It is believed that the stability of the oscillator was much better without the capacitor, but measurements to substantiate this belief could not be made.
Prototype: 4.2 (c).
Circuit Type: Feedback, crystal-controlled.
Transistor Class: Junction only.
Approximate Frequency: 0 to greater than alpha cut-off (400 kc).
Stability: $S_y = 3$.
Transistor Types: Any junction.
Design Procedure: Empirical.
Power Output: Voltage only unless an additional winding is added to the transformer.
Desired Transistor Characteristics: High power gain, low phase shift.

Notes:

This circuit gives good output waveform but the frequency stability is relatively poor. This is primarily due to the poor impedance match to the crystal caused by the low transistor power gain and high power loss in the collector tuned circuit (with the particular components shown in the figure).

This circuit is not recommended unless a very high power gain transistor is used.
Figure 7.49. Pierce Oscillator With Isolation Stage.

Prototype: 4.2 (\(\delta\)).
Circuit Type: Feedback, crystal only.
Transistor Class: Junction only.
Approximate Frequency: 0 to greater than alpha cut-off (1 mc).
Stability: \(S_v = 75\).
Transistor Types: Any junction.
Design Procedure: Empirical.
Power Output: Power available depending upon the characteristics of the output stage.
Desired Transistor Characteristics: Low phase shift, constant \(h\) parameters, constant collector capacitance.

Notes:

This circuit is another modification of the Pierce oscillator circuit to include distortion compensation and an isolation amplifier. The frequency is affected very little by changes in loading. The circuit operates over a wide range of frequencies without the necessity of tuning.

The temperature stability of the circuit is fair.

This circuit is recommended for applications where two transistors may be used. However, the circuit should be investigated further since detailed information has not been obtained on it.
Prototype: Vacuum-tube Butler oscillator.

Circuit Type: Feedback, crystal-controlled only.

Transistor Class: Junction only.

Approximate Frequency: 0 to greater than alpha cut-off.

Stability: Quantitative data not available.

Transistor Types: Any junction.

Design Procedure: Vacuum-tube analogy.

Power Output: Voltage only.

Desired Transistor Characteristics: Low phase shift, constant alpha and phase shift.

Notes:

This circuit is a transistorized version of the Butler oscillator circuit. The stability was not measured but was observed to be very poor. A low-impedance crystal is required for satisfactory operation of the circuit.

This circuit is not recommended for general use because of its poor stability.
VIII. HIGH-FREQUENCY OPERATION OF TRANSISTORS

High-frequency operation is used here to indicate operation of transistors at frequencies well beyond their alpha cut-off. With the majority of the transistors that have been used, this means a frequency range from low frequencies to between 1 and 10 mc. With some of the available transistors (2N33 and L-5100), operation is possible to frequencies as high as 20 to 100 mc. Operation above the alpha cut-off frequency presents about the same problems regardless of the actual frequency involved except that at higher actual frequencies some of the basic problems are more severe.

Very few circuits have been presented that operate satisfactorily at frequencies beyond the alpha cut-off frequency of the transistor. Many such circuits have been investigated, and, in most cases, the stability has been very poor. The basic reason has been found to be the large transistor phase shifts accompanying high frequencies. In addition to being large, the phase shift also depends much more upon voltage and temperature. This imposes a basic limitation on the maximum stability that can be obtained without resorting to the stabilization methods discussed in section VI. Of course, most of these stabilization methods are still effective in improving the stability.

The decrease in stability at high frequencies has been found to be much worse with point-contact transistors than with junction transistors. This decrease in stability is apparently due to the greater phase shift because of greater transit time in the point-contact transistor. The major cause of instability in the junction transistors has been attributed to changes in collector and emitter capacitance with voltage and temperature. Since it is impossible for the associated circuit to alter this effect (except with the compensating methods previously discussed), the problem becomes one of providing the
smallest change in frequency for a given change in capacitance. This suggests paralleling the transistors with large, stable capacitors so that the total percentage of change in capacitance is small. This is a satisfactory solution at low frequencies where large capacitances can be used in the resonant circuits. However, at higher frequencies, the maximum total capacitance that can be used is definitely limited, thus establishing the minimum phase shift which must occur for a given change in voltage or temperature.

The only other alternative is to design frequency-controlling networks producing the greatest possible change in phase for a given change in frequency. Such circuits always result in the use of some type of bridge arrangement. Unless a bridge is used, the maximum effective Q that can be realized is the Q of the resonant circuit. (However, this Q is a limiting value which actually cannot be obtained.) The bridge circuits often give effective values for Q greater than the Q of the resonant circuit (some authors refer to this as Q multiplication). This can be accomplished only at the expense of attenuation. In fact, there is a definite relation between the necessary attenuation and the effective increase in Q. Since this is true, then it is possible that there exists a constant frequency-stability product for the active element of any oscillator. This suggests that any increase in frequency of oscillation must be accompanied by a decrease in frequency stability for the same quality of elements throughout the circuit.

All of these considerations have been investigated but sufficient information is not yet available to permit establishing a criterion.

Experience has shown that to obtain good stability the transistor must be used at frequencies below its alpha cut-off. This has been invariably true with all types of oscillators and transistors investigated.
IX. MISCELLANEOUS DATA

A. Transformer Design

Most stable transistor oscillators require careful consideration of impedance matching. Also, many circuits require a phase reversal external to the transistor to provide the proper phase for oscillations. One way in which either or both of these objectives may be accomplished is by the use of transformers, either autotransformers or transformers with two or more windings. Laboratory experience has indicated that autotransformers are more desirable where they can be used, since greater coupling is more readily obtained.

In practically all oscillator applications, the transformer is used with the high impedance side (usually primary) resonated. It is desirable to resonate both primary and secondary in some applications. When autotransformers are used, the entire winding is usually resonated.

Laboratory work has shown that at low frequencies the ideal-transformer equations are satisfactory as long as the coefficient of coupling is reasonably high. The coupled-circuit theory should be used if frequencies are above 1 mc or if tolerances must be less than about ten per cent. Where the ideal transformer approach is adequate, the voltage and current ratios are obtained directly from the turns ratio. The impedance ratio is obtained from the square of the turns ratio. At low frequencies, the stray capacitance may usually be neglected. In most of the designs presented in this report, the ideal-transformer equations have been used in obtaining impedance match.

Since very little use has been made of the coupled-circuit theory, it will not be discussed here. The subject is thoroughly discussed in the literature.
One difficulty which has been encountered at all frequencies, and which has not been solved satisfactorily, is impedance matching involving impedances higher than about 50,000 ohms. Most transformers lead to excessive losses at such impedance levels. The antiresonant impedance of the transformer windings usually does not exceed about 200,000 ohms and is more often a much lower value. Higher values of Q and less loss are made possible by the use of a suitable iron core, but the use of iron cores may be objectionable for other reasons discussed previously. Litz wire has been found useful only in certain frequency ranges to improve the transformer performance. Increasing the physical size of the transformer is a possible but often objectionable method of obtaining satisfactory operation at high-impedance levels.

At frequencies above 1 mc, transformer phase shift becomes an additional problem. Also, changes in loading are more likely to cause serious detuning of resonant transformers. Stray capacitances may cause spurious resonances to exist. At low frequencies, an oscilloscope may be used to analyze the performance of transformers effectively. At frequencies above 5 mc, other techniques must be used because of the phase shifts in the circuits of most oscilloscopes and because of the additional capacitive loading caused even by a r-f probe. For example, the 14-μf r-f probe of the Tektronix oscilloscope may represent a large percentage of the total tuning capacitance required for a 10-mc resonant transformer. The problems of high-frequency transformer design have not been studied extensively since few oscillators have been operated at frequencies above 10 mc. More study on this subject will be necessary to permit the design of stable transistor oscillators to operate at higher frequencies.
B. Frequency Measurements

The development of more stable transistor oscillators has necessitated the
investigation of accurate methods for measuring frequency.

The method of measuring frequency initially employed was to inject the un-
known frequency into a Berkeley Model 5570 Frequency Meter and to obtain a di-
rect frequency reading. The accuracy of the instrument, claimed by the manu-
facturer, was 1 part in $10^7 \pm 1$ count. The effective counting period was one
second. At frequencies below 10 mc, the accuracy was therefore limited by the
$\pm 1$ count. At higher frequencies, the accuracy was limited by the accuracy of
the internal oscillator of the instrument.

Since most of the laboratory investigations were at frequencies below 10
mc, it was necessary to modify the frequency meter to provide longer counting
periods. Accordingly, an attachment was developed to permit effective counting
periods of 10 and 100 seconds. The $\pm 1$ count limitation now provided accur-
cies as great at 1 part in $10^7$ at frequencies as low as 100 kc. Difficulties
were then experienced in obtaining reliable readings on oscillators known to
have this degree of stability. The trouble was traced to the standard oscilla-
tor built into the frequency meter. The instability of this oscillator could
not be easily corrected since it was partially caused by the high ambient tem-
perature of the laboratory which resulted in overheating of the instrument.

The frequency-measuring procedure first used was to calibrate the frequency
meter against WWV once every week or two. This resulted in errors as great as
2 or 3 parts in $10^6$ over the period of a week and errors as great as 1 part in
$10^6$ over the period of a few minutes. Therefore, it became necessary to compare
continuously the frequency of the frequency meter oscillator with WWV. The
Final Report, Project No. 236-206

oscillator frequency was corrected continuously while an unknown frequency was being measured. This provided accuracies as great as 1 part in $10^7$ when WWV reception was good, the laboratory temperature was moderate and the a-c line voltage was constant. However, all of these conditions were rarely achieved at the same time.

Accordingly, the transistorized Meacham bridge oscillator which was discussed in section VII-C was constructed in a form suitable for use in the frequency meter. However, it was found that the relative drift between two transistor oscillators was considerably less than the drift between either one and WWV. When the two transistor oscillators were operated in totally different environment and from different power sources, this was still true. Eventually, it was learned that the drift of the transistor Meacham oscillator as indicated by WWV was a function of the time of day. This and other facts led to the conclusion that the absolute frequency of WWV as received at the laboratory could be relied upon to an accuracy of only $3$ or $4$ parts in $10^8$ at the time of the year that the tests were being made. Presumably, this was caused by atmospheric conditions. Further investigation indicated that the transistorized Meacham oscillator could be relied upon to 1 part in $10^8$ over short periods of time. Further improvements in this oscillator were contemplated but were not completed at the termination of the contract.

To facilitate oscillator comparisons, frequency-voltage stability has usually been given in terms of the frequency change produced by a one per cent voltage variation. Before the improved frequency-measuring techniques were developed, it was often necessary to change an oscillator supply voltage by 25 per cent or more to detect a change in frequency. As an example, one oscillator
operating at 1 mc showed a frequency change of one cycle per second for a 50 per cent change in supply voltage. This would be equivalent to a 2 parts in $10^8$ change in frequency for a one per cent change in supply voltage only if the frequency could be proved to be a linear function of the supply voltage. This was established by substituting a less stable resonant element in the oscillator so that the change in frequency for a one per cent change in supply voltage could be measured. The frequency-voltage characteristic was examined for small increments of supply-voltage change over the voltage range originally covered by the 50 per cent change. Since this characteristic was found to be fairly linear, the original oscillator could be claimed to have a stability of 2 parts in $10^8$ for a one per cent voltage change. This procedure was often used to justify stating the stability of an oscillator in terms of a one per cent supply-voltage change. The one per cent supply-voltage change was chosen because this is a typical tolerance for a regulated power supply. When the voltage stability, $S_v$, as defined in section II-A is used in this report, it may be assumed that the value applies for voltage changes as small as one per cent.

By using combinations of the above-mentioned procedures, stabilities as great as 1 part in $10^9$ for one per cent change in supply voltage ($S_v = 1000$) may now be measured.

C. Parameter Measurements and Calculations

Various methods of designing transistor oscillators depend to different degrees upon the determination of transistor parameters. Several methods have been used to obtain this necessary information. Most of these methods depend upon a suitable choice of equivalent circuits, while some methods merely require
testing the transistor as an amplifier to obtain sufficient information for an oscillator design. In some cases the low-frequency parameters are satisfactory while in other cases the high-frequency characteristics are necessary. A choice is also necessary between small-signal parameters and large-signal parameters.

Generally, the measurements of interest are the low- and high-frequency small-signal parameters and the static characteristics. Several sets of parameters have been used for expressing these characteristics. The small-signal parameters usually involve the current gain, $\alpha$ and $(1 - \alpha)$, and the open-circuit impedances, $z_{11}$, $z_{12}$, $z_{21}$ and $z_{22}$, of the transistor when treated as an active linear four-pole network. For some purposes, the actual elements of the $T$ (or $\pi$) equivalent circuit, namely, $z_e$, $z_b$ and $z_c$ (or $y_{eb}$, $y_{ce}$ and $y_{cb}$), may be used. These quantities are relatively easy to measure at audio frequencies, but they are difficult to measure at high frequencies where the simple $T$ and $\pi$ equivalent circuits no longer represent the transistor accurately, and where transit time and internal reactance effects cannot be neglected.

In large-signal or nonlinear applications of the transistor, static curves are useful if not entirely necessary. In some commercial equipment, point-by-point methods of obtaining such curves have been replaced by rapid-sweep methods which present the desired information as a family of curves displayed on a CRO or as actual plots on a curve plotter.

The analysis of transistor circuits from the matrix viewpoint has become increasingly popular. This method conveniently eliminates the necessity of first replacing the transistor by its equivalent circuit. The transistor is treated as an active linear four-pole network so that its electrical behavior is completely described by two equations which relate the input and output.
currents and voltages. Since only two of the four variables are independent, the two equations can be written in a maximum of six permutations. Three of these—those involving the open-circuit impedance \((z)\) parameters, the short-circuit admittance \((y)\) parameters and the hybrid \((h)\) parameters—are useful in connection with transistor analysis and measurement.

The \(h\)-parameters of a grounded-base stage are usually selected as the parameters to be used as the basis for circuit analysis and design. They are identified by the equations

\[
\begin{align*}
v_1 &= i_1(h_{11}) + v_2(h_{12}) \\
i_2 &= i_1(h_{21}) + v_2(h_{22})
\end{align*}
\]

where

\[
\begin{align*}
h_{11} &= \left. \frac{\partial v_1}{\partial i_1} \right|_{v_2 = \text{constant}} \\
h_{12} &= \left. \frac{\partial v_1}{\partial v_2} \right|_{i_1 = \text{constant}} \\
h_{21} &= \left. \frac{\partial i_2}{\partial i_1} \right|_{v_2 = \text{constant}} \\
h_{22} &= \left. \frac{\partial i_2}{\partial v_2} \right|_{i_1 = \text{constant}}
\end{align*}
\]

The \(h\)-parameters require both open-circuit input and short-circuit output conditions. This is advantageous from the measurement-and-circuit-operation point of view because the transistor is inherently a low-input-impedance, high-output-impedance device in the more common circuit applications. Therefore, the required terminal conditions can be easily achieved so that the \(h\)-parameters can readily be measured with existing measuring equipment.
Figure 9.1 shows the equivalent circuit of a grounded-base stage whose identifying equations are

\[ v_{eb} = i_e(h_{11}) + v_{cb}(h_{12}) \]
\[ i_c = i_e(h_{21}) + v_{cb}(h_{22}) \]  \hspace{1cm} (9.3)

The h-parameters in equation 9.3 are related to the familiar z- and y-parameters in the following manner:

\[ h_{11} = \frac{\Delta_z}{z_{22}} = \frac{1}{y_{11}} \]
\[ h_{12} = \frac{z_{12}}{z_{22}} = -\frac{y_{12}}{y_{11}} \]  \hspace{1cm} (9.4)
\[ h_{21} = \frac{z_{21}}{z_{22}} = \frac{y_{21}}{y_{11}} \] and
\[ h_{22} = \frac{1}{z_{22}} = \frac{\Delta_y}{y_{11}} \]
where

\[ \Delta_z = z_{11}z_{22} - z_{12}z_{21} ; \quad \Delta_y = y_{11}y_{22} - y_{12}y_{21} ; \]

also,

\[ \alpha = -h_{21} = -\frac{y_{21}}{y_{11}} = \frac{z_{21}}{z_{22}} . \]  

(9.5)

Examination of equations 9.3 and 9.4 reveals the fact that the h-parameters may be used directly in matrix analysis without conversion. Moreover, they provide information on the important factors affecting transistor performance. For example, \( h_{21} \) is the negative of the current amplification factor \( \alpha \); \( h_{12} \) is the reverse voltage amplification factor and provides information on internal feedback performance; \( h_{11} \) and \( h_{22} \) are useful for impedance matching. Also, the h-parameters are said to be more nearly independent of each other than the more conventional equivalent circuit parameters. The so-called "active mutual" impedance \( z_m \) (or admittance \( y_m \)) found in the T (or \( \pi \)) equivalent circuit has no counterpart in the h-parameter equivalent circuit. Finally, the fact that the h-parameters can be readily transformed into the z- or y-parameters by means of simple algebraic relations adds to their usefulness. Figure 9.2 gives a tabulation of the important relations of the grounded-base z-, y- and h-parameter equivalent circuits.

The current amplification factor, \( \alpha \), has been defined as

\[ \alpha = \left. \frac{\partial i}{\partial v_{cb}} \right|_{v_{cb} = \text{constant}} , \]

(9.6)

so that at any one operating point and frequency, \( \alpha \) remains constant regardless of the way in which the transistor is connected.
### Z-Parameter

\[
\begin{align*}
    v_{eb} &= i_e(z_b + z_e) + i_c(z_b) \\
    v_{cb} &= i_e(z_b + z_m) + i_c(z_b + z_c)
\end{align*}
\]

### Y-Parameter

\[
\begin{align*}
    i_e &= v_{eb}(y_{eb} + y_{ce}) + v_{cb}(-y_{ce}) \\
    i_c &= v_{eb}(y_m - y_{ce}) + v_{cb}(y_{cb} + y_{ce})
\end{align*}
\]

### H-Parameter

\[
\begin{align*}
    v_{eb} &= i_e(h_{11}) + v_{cb}(h_{12}) \\
    i_c &= i_e(h_{21}) + v_{cb}(h_{22})
\end{align*}
\]

### General Formulæ

\[
\begin{align*}
    i_b + i_c + i_e &= 0 \\
    v_{cb} + v_{be} + v_{ec} &= 0 \\
    \alpha &= \frac{\partial i_c}{\partial i_e} |_{v_{cb} = \text{constant}} \\
    \alpha &= \frac{z_{21}}{z_{22}} = \frac{y_{21}}{y_{11}} = -h_{21}
\end{align*}
\]

### Z -> H Transformations

\[
\begin{align*}
    z_{11} &= h_{11} - (h_{21})(h_{12}) \\
    z_{12} &= \frac{h_{12}}{h_{22}} \\
    z_{21} &= \frac{h_{21}}{h_{22}} \\
    z_{22} &= \frac{1}{h_{22}}
\end{align*}
\]

### Y -> H Transformations

\[
\begin{align*}
    y_{11} &= \frac{1}{h_{11}} \\
    y_{12} &= \frac{h_{12}}{h_{11}} \\
    y_{21} &= \frac{h_{21}}{h_{11}} \\
    y_{22} &= h_{22} - (h_{21})(\frac{h_{12}}{h_{11}})
\end{align*}
\]

*Figure 9.2. Equivalent Circuits and Important Relations of a Grounded-Base Stage.*
Figure 9.3 shows the basic circuits of a test set constructed for measuring the small-signal h-parameters and \((1 - \alpha)\).

\[
\begin{align*}
(a) & \text{ MEASURES } h_{11}, h_{21}, (1-a). \\
(b) & \text{ MEASURES } h_{12}, h_{22}.
\end{align*}
\]

Figure 9.3. Basic Circuits for Measuring h-Parameters.

The d-c bias supply is not shown. Both circuits are incorporated into a single unit, and switching arrangements have been provided to enable the operator to make measurements easily and quickly. The parameter values are read as voltage ratios across precision resistors \(R_1, R_2, R_3\), using a sensitive vacuum-tube voltmeter.

This instrument has been useful in obtaining the magnitudes of the small-signal parameters at frequencies from 1 kc up to about 300 kc. The frequency range is limited primarily by the difficulty of obtaining high impedances over a wide band of frequencies without the use of tuned circuits.

In obtaining high-frequency characteristics, use has been made of the fact that some of the parameters are nearly independent of frequency. For example, \(\alpha\) is the only parameter of point-contact transistors that varies appreciably with frequency until the alpha cut-off frequency is reached. Hence, \(\alpha\) can be measured at the frequency of interest, and the other parameters can be measured at low frequencies.
In some instances, the information needed is more easily obtained by operating the transistor in a typical circuit rather than by using a test set. The design procedure of section II-E is an example. Here, the amplifier is designed from the manufacturer's data. The only data necessary are the impedances, gain and phase characteristics which are measured after the amplifier is constructed.

For the application of the design procedure involving the Groszkowski equation (section II-D), information is needed concerning the input impedance of the transistor. This information may be obtained by actual measurement, by graphical analysis of the static characteristic curves of the transistor or by linearization of the transistor characteristic curves. For oscillator design purposes, this information has usually been obtained by actual measurement. The following discussion illustrates the procedure.

It was desired to determine the magnitude of negative resistance obtainable, its nonlinear properties and the manner in which it changes with various parameters, such as bias and collector or base-circuit impedance for the circuit shown in Figure 2.18. The nonlinear properties were determined by measuring the harmonics in the emitter-voltage wave when the emitter was driven by a high-impedance sinusoidal-current source as described in section II-D.

A few selected curves showing the variation of negative input resistance and significant harmonic voltages with other circuit parameters are shown in Figures 9.4 through 9.8. All of these curves were taken by using a single type 1768 point-contact transistor.

Figures 9.4 and 9.5 show the manner in which the input resistance $R_1$ and the second-harmonic voltage $H_2$ vary with the amplitude of the signal current in the emitter circuit. The curves in Figure 9.4 were obtained with $R_c$ much
smaller than the transistor-collector resistance $r_c$, so that the collector current was almost independent of $R_c$. Under these conditions, the input resistance is negative and only slightly smaller in magnitude than $R_b$. This magnitude is practically constant and little distortion is present for signals up to 1 ma. Reference to the static collector family curves for the type 1768 transistor indicates that the operation should be essentially linear in this case.

Conditions for the curves of Figure 9.5 are the same as for Figure 9.4 except that $R_c$ was increased to 2,700 ohms, which is only slightly smaller than $r_c$. The magnitude of $R_1$ is reduced to approximately $R_b/2$. However, $R_1$ decreases appreciably, and the second harmonic component of the emitter voltage increases for larger signal amplitudes. The static curves show that in this case the collector is being driven into a nonlinear region when the emitter current exceeds approximately 0.7 ma rms. This is in agreement with the data plotted in Figure 9.5.

In order for a negative-resistance oscillator to have good amplitude stability, $R_1$ must decrease with increasing signal amplitude, and it is desirable that the harmonic voltages remain small. The circuits corresponding to Figures 9.4 and 9.5 are not particularly good in this respect.

One way in which this characteristic can be improved is seen by comparing the curves in Figures 9.6(a) and 9.6(b). The data in Figure 9.6(a) correspond to an $R_c$ of 390 ohms. The circuit used for Figure 9.6(b) was identical to Figure 9.6(a) except that a Mazda Lamp having a cold resistance of 400 ohms was used for $R_c$. The increase in the lamp resistance due to heating causes the input resistance to decrease much more rapidly with signal current than it does.
Figure 9.4. The Effect of Signal Current on $R_1$ and $H_2$, $R_c = 1k$.

$V_c = -8 \, v$

$R_b = 1 \, k$

$i_e = 3 \, ma$

$R_b = 1 \, k$

Figure 9.5. The Effect of Signal Current on $R_1$ and $H_2$, $R_c = 2.7k$.
Figure 9.6. An External Limiter Reduces Distortion.
with a linear $R_c$. Notice that for the same magnitude of $R_1$, the harmonic content in Figure 9.6(b) is smaller than that in Figure 9.6(a). For example, when $R_1 = 700$ ohms, the percentage of $H_2$ in Figure 9.6(b) is only half that in Figure 9.6(a).

Figures 9.7 and 9.8 show typical variations of $-R_1$ and percentage of $H_2$ when the signal amplitude is fixed but a d-c bias is changed. Both curves show regions of reasonably fast change of $R_1$ and small harmonic content. This suggests the possibility of using some kind of automatic gain control to limit the amplitude of oscillation.

For example, Figure 9.7 indicates that for $V_c = -5$ volts, $I_e = 2.5$ ma and $i_e = 0.3$ ma, $R_1 = -600$ ohms, and the harmonic voltage is less than five percent of the fundamental voltage. If an AGC circuit can be used to increase $I_e$ when $i_e$ exceeds 0.3 ma, the amplitude of oscillation may be stabilized at a value that produces small distortion. A similar result could be obtained by causing the collector-bias voltage to decrease when the amplitude of oscillation increases.

By measurements such as this, methods can usually be found for improving the stability of the resulting oscillators.

The following procedure describes a simple method of graphical analysis and a method of linearization for determination of the input volt-ampere characteristic of a grounded-collector point-contact transistor circuit.

The grounded-collector transistor circuit under consideration is shown in Figure 9.9.
Figure 9.7. Variation of $R_1$ and $H_2$ With Emitter Bias.

Figure 9.8. Variation of $R_1$ and $H_2$ With Collector Bias.
Equations describing the operation of this circuit are:

\[ V_c - V_e = I_e R_e - E_{cc} \quad \text{(9.7)} \]

\[ V_1 = -V_c - E_{cc} \quad \text{(9.8)} \]

and \[ I_b = -I_e - I_c \quad \text{(9.9)} \]

The desired input characteristic of this circuit shows the variation of \( I_b \) with \( V_1 \). The functional relationship between these two variables can be obtained by solution of equations 9.7, 9.8 and 9.9.

One obvious method of solving these equations graphically is as follows: The input characteristic (\( V_e \) versus \( I_e \) for constant \( I_c \)) and the forward characteristic (\( V_c \) versus \( I_e \) for constant values of \( I_c \)) can be used to plot the variation of \( V_c - V_e \) with \( I_e \) for constant values of \( I_c \). Load lines can then be constructed on this new set of curves for different values of \( R_e \) and \( E_{cc} \) by the use of equation 9.7.

Any point on one of these load lines corresponds to a particular value of \( I_e \) and \( I_c \) for the circuit of Figure 9.9. When these two currents are determined,
the value of $I_b$ for this particular operating condition is specified by equation 9.9.

The voltage $V_c$ can be found by referring the particular value of $I_e$ selected on the load line back to the forward characteristic. The voltage $V_1$ may then be obtained by use of equation 9.8. Calculation of the points for the entire input volt-ampere characteristic of the circuit may be completed by continued repetition of this procedure.

This graphical method is time-consuming because the forward characteristic and the input characteristic of the transistor must be combined to provide a new set of characteristic curves. Fortunately, a study of typical point-contact transistor characteristics shows that $V_e$ is small compared with $V_c$ for most operating conditions. Furthermore, a study of the behavior of the input volt-ampere characteristics for various load lines shows that the negative-resistance portion of the characteristic does not ordinarily occur in the region where $V_e$ is comparable in magnitude to $V_c$. This means that the voltage $V_e$ can ordinarily be neglected in equation 9.7, and, consequently, the load lines can be drawn on the forward characteristic according to the approximation

$$V_c = I_e R_e - E_{cc}$$  \hspace{1cm} (9.10)

Any point on the load line drawn on the forward characteristic by use of equation 9.10 specifies a particular value of $I_e$ and $I_c$ for the circuit of Figure 9.9. The current $I_b$ for this operating condition may then be found by use of equation 9.7. The voltage $V_1$ may be determined by combining equations 9.10 and 9.8.

$$V_1 = -V_c - E_{cc} = -I_e R_e$$  \hspace{1cm} (9.11)
The value of collector current $I_c$ for $V_1 = -I_e R_e = 0$ is most easily obtained by referring the value of $E_{cc}$ used to the output characteristic of the transistor. In fact, the number of points calculated for the input characteristic can be increased considerably by use of the output characteristic as well as the forward characteristic.

The forward characteristic and output characteristic curves for a typical point-contact transistor are shown in Figure 9.10. A load line drawn by the use of equation 9.10 for $R_e = 2000$ ohms and $E_{cc} = 14$ volts is shown on the forward characteristic. The transistor characteristics shown in Figure 9.10 were used to calculate the input volt-ampere characteristics of the grounded-collector transistor circuit as shown in Figure 9.11.

Inspection of the load line shown in Figure 9.10 shows that for large values of $V_1 = -I_e R_e$ the emitter current $I_e$ changes slowly while the collector current $I_c$ changes rapidly. The collector current $I_c$ first increases in magnitude and then decreases as the emitter current $I_e$ decreases. This causes a peak in the input characteristic of the current. The rapidly changing collector current has the greatest effect on the formation of this peak. Consequently, the peak occurs near the point where the load line crosses the "knee" of the forward characteristic curves. Actually, the peak occurs for a somewhat smaller value of $I_e$ than the value at the knee. This small shift is caused by the effect of the variation of $I_e$ on the base current $I_b$.

A reasonably accurate input characteristic for the grounded-collector circuit can usually be calculated very rapidly by the method just described, provided the transistor characteristic curves are available. In case the transistor curves are not available, an approximate idea of the shape of the input
Figure 9.10. Characteristic Curves for a Typical Point-Contact Transistor.
Figure 9.11. Effect of Collector Voltage on Input Characteristic.
The characteristic may be obtained by consideration of small-signal parameters for the transistor. The method given here for accomplishing this requires the linearization of the forward characteristic of the transistor and is subject to the approximations inherent in the graphical method discussed above as well as the rather serious errors inherent in the linearization process.

The idealized forward characteristic of the transistor is shown in Figure 9.12. The idealized characteristic is divided into three regions:

Region 1: \( I_e < 0. \)
Region 2: \( 0 < I_e < -\frac{I_c}{\alpha}. \) \hspace{1cm} (9.12)
Region 3: \( I_e > -\frac{I_c}{\alpha}. \)

The slopes of these straight lines are approximated by the small-signal parameter \( r_{21} \) measured in the appropriate regions.

The intercept of the lines for \( I_e = 0 \) may be obtained from the emitter-diode characteristic which is assumed to be described by the relation

\[ V_c = I_c r_{22}. \] \hspace{1cm} (9.13)

The small-signal parameter \( r_{22} \) ideally is measured for \( I_e = 0 \). An idea of the extent of the approximation involved in using equation 9.13 may be obtained by inspecting the \( I_e = 0 \) curve on the input characteristic shown in Figure 9.10.

The equations of the three straight lines used in the idealized forward characteristic are:
Figure 9.12. Idealized Forward Characteristics of a Point-Contact Transistor.
Region 1:

\[ V_c = I_e (r_{21})_1 + I_c r_{22} \]  \hspace{1cm} (9.14)

Region 2:

\[ V_c = I_e (r_{21})_2 + I_c r_{22} \]  \hspace{1cm} (9.15)

Region 3:

\[ V_c = I_c (r_{21})^3 + I_c \left[ \frac{(r_{21})_3}{\alpha} - \frac{(r_{21})_2}{\alpha} \right] \]  \hspace{1cm} (9.16)

The equation of the knee of the curves is

\[ V_c = I_e [\alpha r_{22} - (r_{21})_2] \]  \hspace{1cm} (9.17)

Thus, the knee of the curve occurs at a value of \( V_c = 0 \) when

\[ \alpha r_{22} - (r_{21})_2 = 0. \]  \hspace{1cm} (9.18)

Equations 9.14, 9.15 and 9.16 may be combined with equations 9.8, 9.9 and 9.10 to determine the idealized input volt-ampere characteristic of the grounded-collector point-contact transistor circuit shown in Figure 9.9. This idealized input volt-ampere characteristic is shown in Figure 9.13.

It should be emphasized, again, that the assumptions used in obtaining this idealized characteristic are so severe that they cannot ordinarily be used to obtain accurate quantitative information.

In addition to the various measurements and calculations discussed above, it is desirable to know how these parameters vary with voltage and temperature. This information can, of course, be found if the parameters themselves can be measured or calculated.
$E_{cc} = \text{Supply Voltage}$

$R_e = \text{External Emitter Resistance}$

$K = 1 + \frac{(r_{21})^3 - (r_{21})^2}{\alpha r_{22}}$

$\alpha = \frac{\frac{\partial I_c}{\partial I_e}}{V_c}$ measured in Region 2

$\frac{(r_{21})_m}{\frac{\partial V_c}{\partial I_e}} = \frac{I_c}{l_e}$ measured in Region $m$

$\frac{r_{22}}{I_c} = \frac{\frac{\partial V_c}{\partial I_c}}{l_e}$ $l_e = 0$

Figure 9.13. Idealized Input Volt-Ampere Characteristics for Circuit of Figure 9.9.
A. Summary of Results Achieved

The principal results of this investigation may be listed as follows:

1. The development and analysis of several transistor oscillator circuits which are stable against voltage variations at frequencies below the alpha cut-off frequency of the transistor.

2. The development of the Groszkowski analysis to a point where it is useful in the design of low-frequency negative-resistance oscillators.

3. The development of a design procedure for feedback oscillators which insures relatively good frequency stabilities.

4. The development of methods of compensating and stabilizing transistor oscillators.

5. The demonstration that several oscillator configurations are not suitable for use with transistors where frequency stability is important.

6. The demonstration that the characteristics of crystals and LC tuned circuits are more important in the design of transistor oscillators than in the design of vacuum-tube oscillators.

7. The investigation of transistorized versions of conventional oscillator circuits.

8. The development of circuits for operating transistors as oscillators at frequencies above alpha cut-off.

9. The development of a hypothesis that excellent frequency stability can be obtained only at frequencies well below the alpha cut-off frequency of the transistor to be used.

10. The development of a few transistor oscillator circuits that surpass the corresponding vacuum-tube circuits in many respects.

The above list is not exhaustive. At the same time, it is not implied that the investigations mentioned above are complete in all respects. This matter will be discussed further on the following pages.

The material which has been presented in this report should enable a technician experienced with transistors to choose circuits suitable for specific
applications and to adapt the circuits to more modern transistor types in such a way as to obtain acceptable stabilities. The work has not been advanced to such a point that no laboratory investigation will be necessary. However, the amount of laboratory investigation required should be minimized by carefully following the suggestions and procedures outlines in this report.

B. Remaining Problems

Many of the investigations described in this report are not yet complete. Among these are the Groszkowski analysis and the general design procedure for linear oscillators. Also, some data on many of the practical circuits presented in section VII are missing. Some of the circuits have not been investigated with a sufficient variety of transistors to determine the desired transistor characteristics accurately. Because of the large amount of time required to obtain data, accurate temperature stability data are lacking on many of the circuits. Sufficient data have not been obtained on the transistors themselves to enable accurate correlation with the performance in various circuits. This has been mainly due to the lack of high-frequency parameter measuring equipment. Data are also lacking on the performance of transistor oscillator circuits at frequencies above 5 mc. This has been due to the lack of transistors which operate satisfactorily at the higher frequencies.

From the consideration of the above-mentioned facts, the following recommendations for advancing the development of transistor oscillators are offered.

1. The circuits which have been presented should be investigated with a greater variety of transistors of known characteristics both at high and low frequencies. This would permit more accurate correlation of transistor parameters with circuit performance, and it would help to establish the maximum operating frequencies for various circuits and various degrees of stability.
2. The transistor parameter variations with respect to temperature, as well as voltage, should be investigated more thoroughly for use in the correlation mentioned in (1).

3. An attempt should be made to adapt the Groszkowski equation to feedback transistor oscillators. The validity of the results should be further investigated at higher frequencies for both negative-resistance and feedback oscillators. Capacitance nonlinearities, as well as the variation in transistor capacitance and transit times with voltage and temperature, should be investigated.

4. Other networks should be investigated for use with the general design procedure for linear oscillators. It is possible that networks can be found which will give much greater stabilities than have thus far been obtained.

5. The possibilities of the existence of a "frequency-stability product" as suggested in section VIII should be further investigated. If such a relation can be established, a goal will exist to help determine the amount of additional research required on each circuit configuration.
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XI. PERSONNEL

The key personnel who worked on this project are:

<table>
<thead>
<tr>
<th>Name</th>
<th>Position</th>
<th>Approximate Hours</th>
</tr>
</thead>
<tbody>
<tr>
<td>B. J. Dasher, Sc.D.</td>
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Dr. Dasher, Director of the School of Electrical Engineering and Research Associate, Engineering Experiment Station, Georgia Institute of Technology, joined the project at its beginning. He received the degrees of B.S. and M.S. in E.E. from Georgia Tech in 1935 and 1945, respectively, and the Sc.D. in E.E. from MIT in 1952. From 1940 to 1946 he taught electrical engineering at Georgia Tech, and from 1946 to 1951 he was first Instructor and then Assistant Professor of Electrical Communications at MIT, returning to Georgia Tech in 1951. From 1949 to 1951 he was also a Group Leader of the Telermetering and Instrumentation Group, Project Meteor, Research Laboratory of Electronics, MIT. His fields of special interest include advanced network theory, vacuum-tube electronics (circuit design and analysis), applied mathematics and communication theory.

Dr. Finn, Associate Professor of Electrical Engineering and Research Associate, Engineering Experiment Station, Georgia Institute of Technology, joined the project at its beginning. He received the B.S. and M.S. degrees from Purdue
University in 1943 and 1948, respectively, and the Ph.D. degree in E.E. also from Purdue University in 1952. From 1943 to 1946 he was a radar technician in the A.A.F., and from 1948 to 1952 he was a Teaching Assistant in the Electrical Engineering Department, Purdue University. He joined the faculty of Georgia Tech in 1952. His fields of special interest include advanced network theory, applied mathematics, electronic circuitry and communication theory.

Dr. Jones, formerly Associate Professor of Electrical Engineering and Research Associate, Engineering Experiment Station, Georgia Institute of Technology, joined the project at its beginning. He received his B.S., M.S. and Ph.D. degrees in E.E. from the Georgia Institute of Technology in 1945, 1948 and 1953 respectively. He was engaged in teaching and in research at Georgia Tech from 1948 to 1954. He has also worked on various radar projects for the Engineering Experiment Station. His special interests include oscillators, electronic circuits and digital computers. Dr. Jones left Georgia Tech to work for Hughes Aircraft in August 1954.

Mr. Lowry joined the project as a Research Assistant in July 1954. He received the B.E.E. degree from the Georgia Institute of Technology in June 1952 and the M.S. in June 1955. Mr. Lowry was a radio relay officer in the Signal Corps for two years and an application engineer (student) with Westinghouse Electric Corporation for 15 months. His special interests include digital computers and semi-conductor developments.

Mr. Syn joined the project at its beginning as a Research Assistant. He received the degrees B.E.E. and M.S. from the Georgia Institute of Technology in 1951 and 1952, respectively, and is currently continuing his graduate studies toward the Ph.D. degree. He has had experience teaching electronics and
mathematics. During 1952-1953 he worked as a Research Assistant at the Engineering Experiment Station on Project No. 171-118. His special interests include advanced network theory and advanced electromagnetic theory.

Mr. Warren joined the project as a Research Assistant in April 1954. He received the B.S.E.E. from the Georgia Institute of Technology in 1953 and the M.S. in E.E. in 1955. Mr. Warren was an electronic technician in the U.S. Navy for three years. He has been on the staff of the Engineering Experiment Station since January 1953. His special interests include communication circuitry and oscillators. He has worked on Engineering Experiment Station Projects 184, A-129 and A-109.

Mr. Witt, Research Engineer, Engineering Experiment Station, Georgia Institute of Technology, joined the project in January 1954. He received his B.S. in E.E. from Tennessee Polytechnic Institute in 1950 and completed the requirements for the M.S. in E.E. in 1953 at the Georgia Institute of Technology, where he is currently continuing his graduate studies toward the Ph.D. degree. Mr. Witt served as Electronics Instructor in the U.S. Navy from 1947 to 1948. He was first a student laboratory and teaching assistant from 1948 to 1950 and then an Instructor in Electrical Engineering from 1950 to 1951 at Tennessee Polytechnic Institute. He joined the Engineering Experiment Station in June 1951 and has worked on Projects 157, 184 and A-129 as Research
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Respectfully submitted:

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XI. BIBLIOGRAPHY


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A. Derivation of Groszkowski's Equation

Groszkowski's equation, which was first presented as equation 2.35 of section II-D, is derived as follows.

It will be assumed that the negative-resistance device (such as the point-contact transistor circuit shown in Figure 2.17) is unable to store energy. Therefore, over any closed cycle, the net energy must be zero or expressed mathematically,

\[ \int idv = 0. \]  

(13.1)

This is necessarily true if the current is a single-valued function and is true in some cases for multivalued functions.

If the voltage and current waves are assumed to be periodic, they may be expressed in Fourier series as

\[ i = \sum_{m=1}^{\infty} I_m \sin(m\omega t + \alpha_m) \]  

(13.2)

and

\[ v = \sum_{n=1}^{\infty} V_n \sin(n\omega t + \beta_n), \]  

(13.3)

where \( \omega \) is the operating frequency and \( \alpha \) and \( \beta \) represent phase displacement angles.

Equation 13.3 gives

\[ d_v = \sum_{n=1}^{\infty} n\omega V_n \cos(n\omega t + \beta_n) dt. \]  

(13.4)
Substituting 13.2 and 13.4 into 13.1 gives

\[ \int idv = 0 = \int_0^{2\pi/\omega} \sum_{m=1}^\infty \sum_{n=1}^\infty n \omega I_m V_n \sin(m \omega t + \alpha_m) \cos(n \omega t + \beta_n) dt, \quad (13.5) \]

where the double summation includes all possible products.

Since the Fourier series are absolutely convergent, the order of summation may be interchanged and the trigonometric identity

\[ \sin m \cos n = \frac{1}{2} [\sin(m + n) + \sin(m - n)] \quad (13.6) \]

may be applied to 13.5 to obtain

\[ 0 = \sum_{m=1}^\infty \sum_{n=1}^\infty \frac{1}{2} n \omega I_m V_n \int_0^{2\pi/\omega} \{ \sin[(m + n)\omega t + \alpha_m + \beta_n] + \sin[(m - n)\omega t + \alpha_m - \beta_n] \} dt, \quad (13.7) \]

but the integral in 13.7 is zero for all terms where \( m \) and \( n \) are not equal since no net area is represented. Therefore, all such terms may be rejected by letting \( m \) equal \( n \) giving

\[ 0 = \sum_{n=1}^\infty \frac{1}{2} n \omega I_n V_n \int_0^{2\pi/\omega} \{ \sin[2n \omega t + \alpha_n + \beta_n] + \sin[\alpha_n - \beta_n] \} dt \]

\[ = \sum_{n=1}^\infty \frac{1}{2} n \omega I_n V_n \left( \int_0^{2\pi/\omega} \sin[2n \omega t] + \alpha_n + \beta_n dt + \int_0^{2\pi/\omega} \sin[\alpha_n - \beta_n] dt \right). \quad (13.8) \]
The first integral of 13.8 is zero since it is an integral of a function of time over one or more complete cycles. Therefore, equation 13.8 becomes

\[ 0 = \sum_{n=1}^{\infty} \frac{1}{2} n \omega I_n V_n \frac{2\pi}{\omega} \sin(\alpha_n - \beta_n). \]  

(13.9)

The \( \omega \)'s cancel and both sides of the equation may be divided by the constants leaving

\[ 0 = \sum_{n=1}^{\infty} n I_n V_n \sin(\alpha_n - \beta_n). \]  

(13.10)

Now, let \( \theta_n \) represent the phase angle of the tuned circuit at the fundamental and harmonic frequencies. That is,

\[ \theta_n = \alpha_n - \beta_n. \]  

(13.11)

For the oscillator circuit shown in Figure 13.1,

\[ V_n = I_n Z_n \]  

(13.12)

where \( Z_n \) is the impedance of the tuned circuit. Therefore,

\[ V_n \sin \theta_n = I_n X_n \]  

(13.13)

where \( X_n \) is the reactance of the tuned circuit.
Substituting 13.11 and 13.13 in 13.10 gives,

\[ 0 = \sum_{n=1}^{\infty} n I_n^2 X_n, \]  

which is the same as equation 2.35 of section II-D.

The restrictions on the use of this equation, as noted above, are:

1. the negative-resistance device must not be capable of energy storage, and
2. the voltage and current waveforms must be periodic.

The admittance form of the Groszkowski equation may be derived in a similar manner to yield,

\[ 0 = \sum_{n=1}^{\infty} n V_n^2 B_n. \]  

(13.15)
Groszkowski also considered the possibility of applying similar procedures to feedback oscillators.

B. Evaluation of Groszkowski's Equation as Applied to Transistor Oscillators

A series of experiments was performed for the purpose of finding out whether Groszkowski's equation gives sufficiently accurate results to be useful in connection with transistor oscillators. The question was not whether the equation is correct, but rather whether or not it applies to transistor circuits which are not as simple as the theory assumes. These experiments were performed at low frequencies (approximately 1,000 cps) in order to reduce the effects of parasitic capacitances, transistor phase shifts, etc. Also, low-Q circuits were used to facilitate frequency measurements. Frequency was measured by means of a Berkeley model 554 EPUT meter. Since this instrument has a relatively low input impedance, a cathode-follower circuit employing a type 6C4 tube was used as an isolation amplifier. Voltages were measured with a General Radio type 736A wave analyzer.

It is not feasible to calculate \( \omega_0 \) directly because the element values cannot be determined with sufficient accuracy. It probably would be possible to measure the resonant frequency of the tuned circuit. However, instead of comparing measured and computed operating frequencies, it was decided to use equation 2.47 to calculate \( f_o \), the natural resonant frequency of the tank circuit, in terms of the actual frequency and the distortion voltages. Solving equation 2.47 for \( f_o \) gives

\[
f_o \approx f \left[ 1 + \frac{1}{2Q_0^2} \sum_{n=2}^{\infty} \left( \frac{V_n}{V_1} \right)^2 \frac{n^2}{n^2 - 1} \right].
\] (13.16)
Equations 2.47 and 13.16 involve the assumption that $\Delta \omega/\omega_0 \ll 1$, an assumption which is entirely justified.

Figure 13.2 shows the results obtained by varying the collector-supply voltage in the circuit of Figure 13.3.

When studying the curves in Figure 13.2, it is important to remember that the primary objective of these experiments is to determine whether the Groszkowski equation is applicable to the circuit. The significance of the curves relative to oscillator performance is of secondary interest.

At low voltages, the emitter current is positive throughout the cycle, and limiting is mainly due to collector cut-off. As the collector-supply voltage increases, the amplitude of oscillations increases until the voltage reaches 27.5 volts. At this point, the emitter current is negative during a small part of the cycle, and limiting is almost entirely due to emitter cut-off. Insofar as amplitude stability is concerned, emitter limiting appears to be more effective than collector limiting. This is to be expected because the circuit has a large positive resistance when the emitter diode is reversed—much larger than when the collector diode is reversed. This large positive resistance readily offsets the negative resistance, which is effective during most of the cycle.

For voltages below 27.5, the distortion is largely second harmonic. Above this voltage, the higher order harmonics become increasingly important. It may also be observed that both above and below the voltage at which emitter limiting begins the frequency stability is relatively good, but the frequency decreases rapidly during the transition from collector limiting to emitter limiting.

On either side of this transition, the natural frequency of the tank circuit, as computed from equation 13.16, is nearly constant, but it tends to
Figure 13.2. Harmonic Distortion and Frequency as Functions of Collector Supply Voltage.
follow the actual frequency during the transition. Of course, $f_0$ should be quite independent of the collector supply voltage. Since, at the higher voltages, the fifth harmonic is larger than the second and almost as large as the third, it appears that better results would be obtained if harmonics of higher order were included in the calculations.

Figure 13.4 shows another comparison of actual frequency and computed natural frequency. The circuit used for this data was the same as in Figure 13.2 except that a different transistor was used. Harmonics through the fifth were included in the calculations. The variation of the harmonic voltages, although not shown, was similar to that in Figure 13.2. The curves in Figure 13.5 were obtained from the same transistor as those in Figure 13.2. Harmonics through the eighth were included in these calculations. In all three of these tests, the emitter-bias current was held constant at 2 ma. The differences in the results are believed to be largely caused by differences in ambient temperature. This conclusion is supported by the fact that the collector current varied from day to day. Two successive runs made with the same setup on the same day gave substantially the same results.
Figure 13.4. Comparison of the Computed Natural Frequency and the Actual Frequency.

Figure 13.5. Frequency Comparison Curves Including the Effects of the First Eight Harmonics.
One striking discrepancy among these results is that the computed natural frequency of the tank circuit varies from run to run, even at low voltages where the distortion is small. It was suspected that this variation might be caused by reactive impedances of the electronic-regulated power supplies that were used as bias-current sources. When these power supplies were replaced by batteries, the results shown in Figure 13.6 were obtained, harmonics through the eighth being included as before. Here the computed natural frequency at low voltages is about 1,002 cps. This value represents the best agreement with the measured natural frequency which was 1,000.5 cps. Otherwise, the curves in Figure 13.6 are not much different from those in Figure 13.2. The "wiggles" in the curves representing computed natural frequency are probably caused by experimental errors, but the persistent difference of about 2 cps between the highest and lowest voltages is evidently caused by phase shifts within the transistor. The Groszkowski equation shows that most of the actual frequency variation is caused by the distortion.

During the experiments just described, a good deal of trouble was experienced because of parasitic oscillations. Although the circuit arrangement is basically a series-mode oscillator, it is possible for parallel-mode oscillations to exist for certain combinations of parameters. The presence of as much as 20 μf of capacitance in parallel with the tuned circuit would cause spurious high-frequency oscillations to appear—presumably parallel-mode oscillations. These oscillations were present during only part of a cycle of the principal mode, namely, while the emitter current was very close to zero. This difficulty was almost entirely eliminated through the use of the cathode follower mentioned earlier.
C. Oscillator Design Problem

The analysis presented in section II-D has been applied to the design of a two-terminal, parallel-mode oscillator. Information obtained from the analysis of idealized characteristics served as a guide in the selection of parameter values and prediction of resulting performance. The common-emitter configuration is best suited to parallel-mode operation with point-contact transistors. Therefore, to make use of the method described, some information is needed on the negative-resistance characteristic of a point-contact transistor operating with common emitter.

Figure 13.6. Results Obtained with Battery Operation.
Useful input characteristics for the common-emitter configuration cannot readily be obtained from the typical characteristics published by manufacturers. Such a technique requires a great deal of graphical manipulation, and the results obtained are unreliable because of discrepancies between the published data and the behavior of any particular transistor. For this design problem, input characteristics were obtained from point-by-point measurements of a WE Type 1768 point-contact transistor. Of course, such measurements may not represent a practical method for routine design problems, but the information obtained in this way may indicate the appropriate methods for later use.

The circuit for obtaining input characteristics is shown in Figure 13.7.

![Figure 13.7. Circuit for Obtaining Input Characteristics.](image)

The resulting $I_b - V_1$ characteristic exhibits a well-defined negative conductance region suitable for use in a parallel-mode oscillator. Variations in shape and orientation of the characteristic with changes in collector voltage, emitter resistance and collector resistance were shown in Figures 2.31, 2.32 and 2.33. (See section II-D.) The operating region selected corresponds to a
collector voltage of 8 volts, an external emitter resistance of 1000 ohms and an external collector resistance of zero. This combination of parameters afforded the best compromise between linearity of the negative-resistance region and stability of characteristics with changes in collector voltage among the various combinations investigated.

Figure 13.8 presents an expanded portion of the characteristics for collector voltages of six, eight and ten volts. Actual dynamic curves for the transistor would be displaced vertically by a small amount from those of Figure 2.32, because the measurements were taken with the transistor "cold" (by rapid transition from low to high currents as a means of reducing discontinuities due to heating). However, the shape of each characteristic employed is sufficiently accurate for present purposes.

A comparison of Figure 2.21(a) with Figure 13.8 indicates a similarity between the idealized type-I characteristic and the actual characteristics for $V_1$ values between plus one and minus two volts. For values of $V_1$ more negative than two volts, the actual curves depart considerably from the idealized curve, so the operating point was chosen to maintain oscillation in the more linear region. The point $P_1$ shown on Figure 13.8 corresponds to a base bias voltage of -1.0 volt and a bias current of 1.35 milliamperes. These bias conditions were obtained by the use of a voltage divider across the collector supply, as shown in Figure 13.9.
Figure 13.8. Characteristics Employed for Oscillator Design.
The variation in bias voltage due to changes in supply voltage depends jointly on the values of $R_1$ and $R_2$ and on the displacement of transistor characteristics. The dashed lines of Figure 13.8 represent bias characteristics, for which the slope is given by:

$$G_{\text{source}} = -\frac{R_1 + R_2}{R_1 R_2}$$

and the voltage intercepts for zero current are given by:

$$V_{\text{intercept}} = \frac{R_1 E_{cc}}{R_1 + R_2}$$

For reasons discussed below, a value of 100 ohms was selected for $R_1$. $R_2$ is determined by:
which for the selected operating point becomes:

$$R_2 = \frac{|E_{cc} - V_1| R_1}{|V_1| - I_b R_1}$$

$$R_2 = \frac{(8 - 1) 100}{1 - (1.35 \times 10^{-3}) 100} = 809 \text{ ohms.}$$

The slope of the bias characteristic is, therefore:

$$G_{source} = -\frac{100 + 809}{100 \times 809} = -11.24 \text{ millimhos (89 ohms).}$$

The intercepts corresponding to the three collector voltages are:

$$V_{\text{intercept}} \text{ for 6 volts} = \frac{100}{809} \times -6 = -0.66v,$$

$$V_{\text{intercept}} \text{ for 8 volts} = \frac{100}{809} \times -8 = -0.88v \text{ and}$$

$$V_{\text{intercept}} \text{ for 10 volts} = \frac{100}{809} \times -10 = -1.10v.$$

Intersections of the bias and transistor characteristics define points $P_2$ and $P_3$ of Figure 13.8, which represent the quiescent operating points for supply-voltage variations of ±25 per cent from the normal eight volts.

Graphical measurements from Figure 13.8 of the transistor characteristic slopes at points $P_1$, $P_2$ and $P_3$ indicate incremental negative conductances as follows:
$G_{o1} = -1.29$ millimhos (775 ohms),

$G_{o2} = -1.38$ millimhos (725 ohms) and

$G_{o3} = -1.21$ millimhos (826 ohms).

The slope of the positive conductance region of each characteristic (for positive values of $V_1$) is approximately 0.05 millimho, corresponding to a positive resistance of 20 kilohms. The ratio of positive and negative slopes is therefore 0.04, and $M$ may be considered zero for this case.

Predicted behavior of the oscillator for different values of load conductance may be observed in Figure 2.27, by regarding the ratios $-R_1/R_0$ as equivalent to $-G_1/G_o$ in the parallel-mode case. Each intersection of a $-R_1/R_0$ curve with the $M = 0$ curve represents a possible operating condition, with corresponding values of $P^2$ and $K$. Since the extreme values for $G_o$ are already known ($-1.38$ and $-1.21$ millimhos), the selection of any conductance will establish extreme values of $P^2$ and $K$, thereby providing information about frequency and amplitude stability. A load conductance of 1.0 millimho (1000 ohms) was selected, and the following values of $P^2$ and $K$ were obtained from Figure 2.27:

<table>
<thead>
<tr>
<th>Supply Voltage (volts)</th>
<th>$G_o$ (Figure 12.8) (mho)</th>
<th>$-G_1/G_o$</th>
<th>$P^2$ (Figure 5)</th>
<th>$K$ (Figure 5)</th>
</tr>
</thead>
<tbody>
<tr>
<td>6</td>
<td>-1.38</td>
<td>0.725</td>
<td>0.09</td>
<td>2.80</td>
</tr>
<tr>
<td>8</td>
<td>-1.29</td>
<td>0.775</td>
<td>0.06</td>
<td>2.15</td>
</tr>
<tr>
<td>10</td>
<td>-1.21</td>
<td>0.826</td>
<td>0.04</td>
<td>1.85</td>
</tr>
</tbody>
</table>

The frequency of operation was set at 1 kc to minimize reactive effects of the transistor. The resonant circuit was constructed with a large air-core...
inductor of approximately 3.6 millihenries, tuned to 1000 cycles and shunted by a resistance to yield a 1000-ohm impedance at resonance. The effective Q of the completed resonator was determined by bandwidth measurements to be 37. Several performance tests have been run on the oscillator, in which frequency was measured with a Berkeley Model 5570 Frequency Meter and amplitude was measured with a calibrated oscilloscope. Results of a typical test are shown below:

<table>
<thead>
<tr>
<th>Supply Voltage (volts)</th>
<th>Peak Voltage Amplitude (volts)</th>
<th>Frequency (200 second counting periods) (cycles)</th>
</tr>
</thead>
<tbody>
<tr>
<td>6.0</td>
<td>0.6</td>
<td>1000.18</td>
</tr>
<tr>
<td>6.0</td>
<td>0.6</td>
<td>1000.17</td>
</tr>
<tr>
<td>6.0</td>
<td>0.6</td>
<td>1000.17</td>
</tr>
<tr>
<td>6.0</td>
<td>0.6</td>
<td>1000.18</td>
</tr>
<tr>
<td>8.0</td>
<td>1.0</td>
<td>1000.23</td>
</tr>
<tr>
<td>8.0</td>
<td>1.0</td>
<td>1000.24</td>
</tr>
<tr>
<td>8.0</td>
<td>1.0</td>
<td>1000.24</td>
</tr>
<tr>
<td>8.0</td>
<td>1.0</td>
<td>1000.24</td>
</tr>
<tr>
<td>10.0</td>
<td>1.2</td>
<td>1000.24</td>
</tr>
<tr>
<td>10.0</td>
<td>1.2</td>
<td>1000.24</td>
</tr>
<tr>
<td>10.0</td>
<td>1.2</td>
<td>1000.24</td>
</tr>
</tbody>
</table>

For the range of supply voltages chosen, measurements show an amplitude variation of 2 to 1 and a frequency variation of 0.07 cycle or 70 ppm.

The relation of frequency to \( P^2 \) and \( Q \) was given by equation 2.61 as:

\[
f = f_0 \left[ 1 - \frac{P^2}{2Q_o^2} \right].
\]

From this relation, the stability corresponding to any pair of values for \( P^2 \) is:

\[
\frac{\Delta f}{f} = \frac{\Delta P^2}{2Q_o^2}.
\]
Using the values for $P^2$ obtained from the analysis, predicted frequency stability is:

$$\Delta f = \frac{(0.09 - 0.04)}{2 \times 37^2} = 18 \text{ ppm}.$$ 

To predict amplitude stability, graphical measurements of $V_o$ (analogous to $I_o$ previously discussed) must be obtained from Figure 13.8. By approximating a "break point" for each characteristic, the following values are obtained:

$V_o$ for 6 volts = 0.3 volt and
$V_o$ for 10 volts = 0.9 volt.

Predicted amplitudes corresponding to supply voltages of 6 volts and 10 volts are therefore:

- Voltage amplitude for 6 volts supply = $KV_o$
  $= 2.80 \times 0.3$
  $= 0.84$ volt and

- Voltage amplitude for 10 volts supply = $KV_o$
  $= 1.85 \times 0.9$
  $= 1.67$ volts.

Therefore, the predicted amplitude variation is 1.99 (compared with a measured variation of 2.0), and the ratio of predicted amplitudes to measured amplitudes is approximately 1.4.

The reason for choosing a voltage divider as a bias source will be apparent from a study of Figure 13.8. If bias were obtained from a single resistor in

-243-
series with the tank circuit, such a resistor would have a value of 741 ohms. The slope of the resulting bias characteristic would be almost the same as the slope of the transistor characteristic, thus affording very poor definition of the operating point and correspondingly poor temperature stability. To provide bias characteristics which intersect the transistor characteristics in well-defined points, the bias source must have an internal conductance which is large compared with the effective negative conductances of the transistor. This consideration was the basis for selecting 100 ohms as the value for $R_1$. The chief disadvantage of such an arrangement is the large power dissipated in the voltage divider, which in this case averaged 76 per cent of the total power consumed.

The value of load conductance described above was selected to provide a particular performance. The original goal was 10 ppm, which would have required a $\Delta P^2$ of 0.05 with a Q of 50. From Figure 2.27, a value of $P^2 = 0.05$ corresponds to a $-R_i/R_o$ ratio of 0.8. Assuming that $\Delta P^2$ would be approximately equal to $P^2$ for the nominal supply voltage (8 volts), this ratio of 0.8 indicated a desired load conductance of 1.03 millimho, which was taken as 1.0 millimho for convenience. The Q of 50 was obtained with an inductance employing a powdered-iron toroidal core, which presented the desired load conductance at resonance. However, nonlinearity of the core caused variations of inductance with amplitude, resulting in frequency variations much larger than those caused by the transistor alone. Within the accuracy of the method used to measure the inductance variation, the core variation accounted for all of the frequency variation. This result led to the construction of an air-core inductance as a means of obtaining linearity.
The results described above represent an accuracy of prediction by analysis which approaches the limits of available information on the components of the resonant circuit. To refine the results at this frequency would require an investigation of the linearity of the capacitance employed, which the objectives did not justify. It can reasonably be concluded that the Groszkowski analysis is effective at audio frequencies.

D. High-Frequency Oscillator Design Problem

High frequency here means frequencies higher than approximately 10 per cent of the alpha cut-off frequency. Thus, for the WE 1768 transistor, frequencies above about 30 kc are considered as high frequencies. The circuit of Figure 13.9 has been operated at 32 kc. A parallel-resonant circuit was constructed which exhibited the same resonant impedance (1000 ohms) and Q (40) as the 1-kc circuit originally employed. From the analysis based on the assumption of a resistive negative impedance, the oscillator performance expected would be the same at 32 kc as it was at 1 kc.

Performance of the circuit at 32 kc did not compare favorably with that at 1 kc when the original WE Type 1768 transistor was employed. The frequency variation for 1 per cent voltage change increased from 1.5 parts in $10^6$ at 1 kc to 45 parts in $10^6$ at 32 kc, and oscillation could be maintained only over a voltage range of ±10 per cent of the nominal supply voltage. Substitution of WE Type 1698 and WE Type 1729 transistors in the oscillator of Figure 13.9 with the 32-kc resonant circuit improved the frequency stability by a factor of 10 and permitted voltage variations of ±25 per cent of nominal supply voltage. With no alteration of the oscillator other than adjustment of the bias network, stabilities between 4 and 5 parts in $10^6$ for 1 per cent voltage change...
were obtained. These results are largely what would be expected from a consideration of the alpha cut-off frequencies of the transistor types employed, as indicated below:

<table>
<thead>
<tr>
<th>Transistor Type</th>
<th>Alpha Cut-off Frequency (mc)</th>
<th>Frequency Variation †</th>
</tr>
</thead>
<tbody>
<tr>
<td>WE 1768</td>
<td>0.35</td>
<td>45</td>
</tr>
<tr>
<td>WE 1698</td>
<td>2.0</td>
<td>4.5</td>
</tr>
<tr>
<td>WE 1729</td>
<td>3.7</td>
<td>4.2</td>
</tr>
</tbody>
</table>

†Parts in $10^6$ for one per cent supply voltage change at 32 kc.

Oscillographic studies of the volt-ampere characteristic at the transistor-input terminals provide a means of determining the conditions which produce multivalued relations between voltage and current. At 32 kc, for example, the input characteristic of a WE Type 1768 transistor produces an unmistakably open figure, indicating a reactance which varies considerably with the operating point and the amplitude of the applied signal. Under the same conditions, the input characteristics of WE Type 1698 and WE Type 1729 transistors are single-valued and produce single lines rather than open figures.
<table>
<thead>
<tr>
<th>Abstract</th>
<th>xix</th>
</tr>
</thead>
<tbody>
<tr>
<td>Alpha cut-off frequency</td>
<td>124, 185</td>
</tr>
<tr>
<td>Amplitude of oscillation</td>
<td>124, 243</td>
</tr>
<tr>
<td>Amplitude stability</td>
<td>33, 230, 243</td>
</tr>
<tr>
<td>Amplitude stabilized oscillator</td>
<td>110</td>
</tr>
<tr>
<td>Analogies, vacuum-tube</td>
<td>77</td>
</tr>
<tr>
<td>Analysis, graphical</td>
<td>26</td>
</tr>
<tr>
<td>Analysis, nonlinear</td>
<td>26, 31</td>
</tr>
<tr>
<td>Appendix</td>
<td>225</td>
</tr>
<tr>
<td>Base-tuned oscillator</td>
<td>130, 131, 137, 140, 141</td>
</tr>
<tr>
<td>Bias compensation</td>
<td>103</td>
</tr>
<tr>
<td>Bibliography</td>
<td>221</td>
</tr>
<tr>
<td>Bridge stabilization</td>
<td>113</td>
</tr>
<tr>
<td>Butler oscillator</td>
<td>184</td>
</tr>
<tr>
<td>Capacitance losses</td>
<td>85</td>
</tr>
<tr>
<td>Capacitance nonlineari-</td>
<td>215, 245</td>
</tr>
<tr>
<td>ties</td>
<td></td>
</tr>
<tr>
<td>Capacitors</td>
<td>85, 86, 106, 108</td>
</tr>
<tr>
<td>Cathode follower</td>
<td>229, 234</td>
</tr>
<tr>
<td>Characteristic types I and II</td>
<td>44</td>
</tr>
<tr>
<td>Collector-tuned oscillator</td>
<td>21</td>
</tr>
<tr>
<td>Compensation, bias</td>
<td>103</td>
</tr>
<tr>
<td>Components, selection of</td>
<td>105</td>
</tr>
<tr>
<td>Construction, mechanical</td>
<td>108</td>
</tr>
<tr>
<td>Core nonlinearity</td>
<td>85, 87, 107, 244</td>
</tr>
<tr>
<td>Crystals</td>
<td>81, 107</td>
</tr>
<tr>
<td>Design examples</td>
<td>235, 245</td>
</tr>
<tr>
<td>Diode Characteristics</td>
<td>111</td>
</tr>
<tr>
<td>Diode stabilization</td>
<td>110</td>
</tr>
<tr>
<td>Distortion, measurement of</td>
<td>198-202</td>
</tr>
<tr>
<td>Emitter tuned oscillator</td>
<td>17, 128, 129</td>
</tr>
<tr>
<td>Empirical design</td>
<td>76</td>
</tr>
<tr>
<td>Energy storage</td>
<td>41, 225, 228</td>
</tr>
<tr>
<td>Feedback, negative</td>
<td>65</td>
</tr>
<tr>
<td>Feedback loops</td>
<td>65</td>
</tr>
<tr>
<td>Figure of merit</td>
<td>70</td>
</tr>
<tr>
<td>Figures, list of</td>
<td>v</td>
</tr>
<tr>
<td>Fourier series representation</td>
<td>41, 45</td>
</tr>
<tr>
<td>Frequency-controlling network</td>
<td>68, 73</td>
</tr>
<tr>
<td>Frequency measurements</td>
<td>189-191</td>
</tr>
<tr>
<td>Frequency pulling</td>
<td>37, 39, 48</td>
</tr>
<tr>
<td>Frequency-stability product</td>
<td>186, 215</td>
</tr>
<tr>
<td>Graphical analysis</td>
<td>26, 198, 202-206</td>
</tr>
<tr>
<td>Groszkowski analysis</td>
<td>31</td>
</tr>
<tr>
<td>Groszkowski equation, derivation of</td>
<td>225-229</td>
</tr>
<tr>
<td>Groszkowski equation, evaluation of</td>
<td>229</td>
</tr>
<tr>
<td>h-parameters</td>
<td>193-196</td>
</tr>
<tr>
<td>Harmonic currents</td>
<td>37</td>
</tr>
<tr>
<td>Harmonic frequency pulling</td>
<td>37, 39, 48</td>
</tr>
<tr>
<td>Harmonic voltages</td>
<td>37</td>
</tr>
<tr>
<td>Hartley oscillator</td>
<td>147, 148</td>
</tr>
<tr>
<td>Heating of components</td>
<td>108</td>
</tr>
<tr>
<td>High frequency</td>
<td>185, 245</td>
</tr>
<tr>
<td>High-frequency oscillators</td>
<td>136, 137, 146</td>
</tr>
<tr>
<td>History of the contract</td>
<td>1</td>
</tr>
</tbody>
</table>
Ideal transformer.............. 187
Impedance, inversion of........ 87
Impedance inversion oscillator.... 135
Impedance matched oscillator..... 152
Impedance matching............. 188
Impedance of power supplies..... 234
Impedance, resonant............ 86
Inductances.................... 106, 107
Inductances, magnetic core...... 85, 87, 107
Introduction.................... 1
Isocline diagram................ 26
Lamp-stabilized oscillator...... 182
Lamp used as a limiter........... 176, 179, 182, 199
Linear circuit design........... 7
Linearization................... 198, 209-211
Load lines....................... 204-206
Magnetic cores................... 85, 87, 107
Matrix equations............... 193-196
Meacham bridge oscillator...... 176-181, 190
Measurements, distortion........ 198-202
Measurements, frequency........ 189-191
Measurements, parameter......... 191
Mechanical construction......... 108
Microphonics.................... 108
Negative feedback.............. 65
Negative feedback network....... 73
Negative feedback oscillator.... 156
Network, frequency controlling.. 68, 73
Network, negative feedback...... 73
Nonlinear analysis............. 31
Oscillation, condition for....... 27
Oscillation, parasitic.......... 234
Oscillators, amplitude
stabilized..................... 135, 145
Butler......................... 184
class-A......................... 78
crystal-controlled............. 94
feedback....................... 79, 92, 94
free-running................... 92
Hartley......................... 147, 148
high frequency.................. 136, 137, 146
impedance inversion............ 135
impedance matched.............. 152
lamp-stabilized................ 182
Meacham bridge.................. 176-181, 190
negative-feedback.............. 156
negative-resistance............. 79, 91
overtone....................... 175
Pierce......................... 173, 174, 183
transformer coupled............. 138, 139, 142
tuned-base...................... 130, 131, 137, 140, 141
tuned-collector.................. 126, 127
tuned-emitter.................... 128, 129
zero-base-current.............. 158, 159, 161, 162, 164
Overtone oscillator.............. 175
Parameter, non-variance of....... 197
Parameters, small-signal........ 192, 197
Parameter test set............... 197
Parasitic oscillation........... 234
Parasitic reactances............ 40
Performance index . . . . . 82
Personnel . . . . . . . . . . . 217
Pi-equivalent circuit . . . . . 192
Pierce oscillator . . . . . . . . 173, 174, 183
Pi-section network . . . . . . 89
Power series representation . 40
Power supply impedance . . . 234
Purpose of contract . . . . . xvii

Q, multiplication of . . . . . 76, 186
Q of resonators . . . . . . . 86

Reactances, parasitic . . . . 40
Recommendations . . . . . . 119, 214
Resistors . . . . . . . . . . . 106
Resonant elements . . . . . . 81
Resonant impedance . . . . . 86

Solder connections . . . . . 108
Stability, amplitude . . . . . . 33, 230, 243
Stability, definition of . . . 5
Stabilization, bridge . . . . . 113
Stabilization by distortion . . 110
Stabilization, diode . . . . . 110
Stabilization of oscillators . . 103
Static curves . . . . . . . . . 192
Symbols, list of . . . . . . . xiii

T-equivalent circuit . . . . . 192
Tables, list of . . . . . . . . xi
Temperature compensation . 85, 109
Test set, parameter . . . . . . 197
Thermistors . . . . . . . . . . 115
Thevenin generator . . . . . 65, 68
Transformer coupled oscillator . . . . . . 138, 139, 142
Transformer design . . . . . 187
Transformer, ideal . . . . . . 187
Transformer, phase shift in . . 188
Transistors, selection of . . . 97
Transistors, table of . . . . . 99
Transit time . . . . . . . . . 64
Transmission line . . . . . . . 88
Tungsten lamps . . . . . . . 107, 108, 199

Vacuum-tube analogies . . . . 77
Varistors . . . . . . . . . . . 107, 115
Voltage divider . . . . . . . . 243

WWV radio station . . . . . . 189, 190

y-parameters . . . . . . . . . 192-196
z-parameters . . . . . . . . . 192-196

-249-