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A TRANSISTORIZED CURRENT-CONTROLLED OSCILLATOR FOR DC-EXCITED STRAIN GAGE APPLICATION

by

C. E. LAND

OCTOBER 1958
WORK PERFORMED UNDER AEC CONTRACT AT-(29-1)-789

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LIST OF ILLUSTRATIONS

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INTRODUCTION

This paper describes the transistorized current-controlled oscillator (abbreviated TCCO). The TCCO was designed primarily to operate with a DC excited, resistive bridge, strain gage transducer to provide a modulating signal for an FM telemetering transmitter. A change in input current produces a proportional change in output frequency of the TCCO. The principal features of this unit are: its compatibility with DC excited, resistive bridge transducers; its size - 2.8 cubic inches; its inherent high reliability due to the simplicity of the circuit and complete transistorization; and its low cost. Its compact and rugged design and its low power requirement make it especially suitable for airborne telemetry applications.

Summary

The purpose of this paper is to present a functional description of the TCCO. This description makes use of the fundamental concepts of the theory associated with each circuit component; however, a rigorous circuit analysis is not intended. On the contrary, the scope of the circuit description is limited to intuitive reasoning verified by experimental results.

A description of the operation of the TCCO is presented in terms of typical experimental transfer characteristics related to a block diagram.

Each major component of the TCCO (i.e., the DC difference amplifier, the variable inductor, and the oscillator) is described in detail. The description of each major component includes the development of a nonrigorous mathematical expression relating the output to the input and a discussion of temperature stability of the component.

Using the mathematical input-output relationships developed as part of the description of each component, an over-all input-output equation is developed which relates the output frequency to the magnitude of the input current.

Circuit performance and proposed applications are discussed briefly preceding a general summary and statement of conclusions.

Description of Operation

General

The TCCO is shown in block diagram form in Figure 1. The TCCO consists of three basic functional components; a DC amplifier, a variable inductor; and a transistor oscillator. A resistive bridge transducer is shown connected across the input terminals of the DC amplifier.

The key component in the conversion of a low-level change in input current to a change in output frequency is the variable inductor. The variable inductor makes use of the principle that the effective permeability of a ferromagnetic core material subjected to the influence of superimposed AC and DC magnetic fields is a function of the magnitude of the DC field.

Since the inductance of a coil wound on a specimen of ferromagnetic material is a function of the permeability of the material, the inductance is also a function of the magnitude of the DC field or of the DC current producing the field. The DC biasing field of the variable inductor is developed by current flowing in two separate control windings. A third winding of the inductor is connected in the tank circuit of an oscillator. Changes in magnitudes of the DC currents in the control windings produce proportional changes in inductance of the tank circuit winding.

The DC difference amplifier appears as a constant resistive impedance to the bridge transducer. When the resistance bridge is balanced, the current is zero in the input circuit of the amplifier. An unbalance in one or more arms of the bridge produces a current in the amplifier input circuit which is proportional to the bridge unbalance and the magnitude of the DC excitation voltage. A change in DC amplifier input current is amplified and appears as a change in DC current in the control windings of the variable inductor. This, in turn, produces a change in inductance of the tank circuit winding of the variable inductor. The oscillator frequency is a function of the inductance in the tank circuit; therefore, a change in inductance results in a change in oscillator frequency.

Strain Gage Bridge

It is a well known principle that strain is proportional to stress in the region below the yield point of most materials. The strain gage bridge makes use of this principle for measuring stresses in a material. The function of the strain gage bridge is to convert stress variations (e.g., in the skin or a structural member of an aircraft or missile) to voltage level changes at the output of the bridge. This is effected by making one or more arms of the resistance bridge active in the sense that their resistances are proportional to the strain in the material to which they are bonded. Therefore, if the bridge is balanced with normal, steady-state stress in the material, stress variations in flight will produce proportional bridge unbalance. The bridge output voltage is zero for the balanced bridge condition. Figure 2 is a typical bridge characteristic giving output voltage as a function of DC excitation voltage and resistance variation in one active arm.

Balanced Bridge Operation

When the strain gage bridge is balanced, $e_s = o$ and $i_s = o$ (Figure 8A). The oscillator output remains at center frequency fo as indicated on the input-output characteristic, Figure 6. With $e_s = o$, the DC amplifier output currents i_{c1} and i_{c2} are determined by the fixed bias on transistors Q_1 and Q_2 and the characteristics of these transistors. A typical transfer characteristic for the DC amplifier is shown in Figure 3. With i_{c1} and i_{c2} fixed at values corresponding to $e_s = o$ as shown on Figure 3, the inductor bias current I_B is adjusted to obtain the desired operating point on the transfer characteristic of the variable inductor. The inductance corresponding to this operating point, Figure 4, is the value required to produce oscillation at center frequency fo as shown on the output frequency versus inductance curve for the oscillator, Figure 5.

Unbalanced Bridge Operation

When the strain gage bridge is unbalanced, a voltage e_s is impressed between the base of Q_1 and the base of Q_2 (Figure 8A) which produces a current i_s. If the polarity of e_s is positive at the base of Q_1 with respect to the base of Q_2, the base current i_{b1} of Q_1 is reduced by approximately the magnitude of i_s, while the base current i_{b2} of Q_2 is increased by approximately the magnitude of i_s. This results in a decrease in Q_1 collector current i_{c1}, and a corresponding increase in Q_2 collector current i_{c2}, as shown in Figure 3. If, on the other hand, the polarity of e_s is positive at the base of Q_2 with respect to the base of Q_1, the base current i_{b2} is reduced, while base current i_{b1} is increased by approximately the magnitude of i_s. This results in a reduction of i_{c2} and a corresponding increase in i_{c1}.

In addition to the collector currents of the DC amplifier, a fixed bias current I_B is used to establish a DC magnetizing force, which, in turn, produces a DC flux density in the cores of the variable inductor. The polarities of the control windings of the variable inductor are opposite; therefore, the DC current relationship which controls the inductance (assuming a constant AC flux density) is the difference between the currents in the control windings, or: $(I_B + i_{c2}) - i_{c1}$. When e_s is positive at the base of Q_1 with respect to the base of Q_2, i_{c2} is increased and i_{c1} is decreased as indicated above. From the relationship $(I_B + i_{c2}) - i_{c1}$, it is seen that the DC flux density is increased for this condition, resulting in a reduction in inductance of the tank circuit winding of the variable inductor (refer to Figure 4), which produces an increase in oscillator frequency (refer to Figure 5). When e_s is positive at the base of Q_2 with respect to the base of Q_1, reasoning similar to that above can be used to show that the inductance of the tank circuit winding will be increased and the oscillator frequency reduced. Figure 6 shows the change in output frequency produced by a change in input current.

DC Difference Amplifier

Transfer Function

The schematic diagram of the DC Difference Amplifier is shown in Figure 7. Figure 8A is an equivalent circuit representation of the amplifier, and Figure 8B is an equivalent circuit representation of transistor Q_1 and its associated circuit components.

In Figure 8A, R_{L1} and R_{L2} represent the resistances of the DC control windings of the variable inductor. Since the two windings have equal numbers of turns, these resistances are essentially equal. Bias resistors R_1 and R_2 are equal; also, R_4 and R_6 are equal. The characteristics of transistors Q_1 and Q_2 are matched so that $\beta_1 = \beta_2$ and $I_{co1} = I_{co2}$.*

The approximate relationships between input and output currents (neglecting I_{co}) are:

$$i_{c1} \doteq \beta_1(i_{b1}) \tag{1}$$

$$i_{c2} \doteq \beta_2(i_{b2}) \tag{2}$$

---

* β is the transistor current feedback factor ($\beta = \frac{a}{1-a}$); I_{co} is the base-collector current when the emitter current is zero.

If the β values of Q_1 and Q_2 are carefully matched so that $\beta_1 = \beta_2 = \beta$ then:

$$i_{c2} - i_{c1} \doteq \beta \, (i_{b2} - i_{b1}) \qquad (3)$$

Equation (3) states that the difference between output currents is proportional to the difference between input currents. This is the fundamental equation for the difference amplifier.[1]

When the strain gage bridge is balanced (i.e., $i_s = o$), Equation (4) expresses the relationship between transistors Q_1 and Q_2 collector currents for the quiescent condition of the amplifier,

$$(i_{c2})_o - (i_{c1})_o \doteq \beta \left[(i_{b2})_o - (i_{b1})_o \right] \qquad (4)$$

where $(i_{c1})o$, $(i_{c2})o$ represent the values of i_{c1} and i_{c2}, respectively, when $i_s = o$; $(i_{b1})_o$, $(i_{b2})_o$ represent the values of i_{b1} and i_{b2}, respectively, when $i_s = o$.

If the strain gage bridge is unbalanced so that i_s flows in the positive direction, indicated in Figure 8A, i_{b1} is decreased and i_{b2} is increased:

$$\begin{aligned}
(i_{b1})_h &= (i_{b1})_o - \Delta\, i_{b1} \doteq (i_{b1})_o - i_s \\
(i_{b2})_h &= (i_{b2})_o + \Delta\, i_{b2} \doteq (i_{b2})_o + i_s \\
(i_{c1})_h &= (i_{c1})_o - \Delta\, i_{c1} \\
(i_{c2})_h &= (i_{c2})_o + \Delta\, i_{c2}
\end{aligned} \qquad (5)$$

where $(i_{b1})_h$, $(i_{b2})_h$, $(i_{c1})_h$ and $(i_{c2})_h$ represent the values of i_{b1}, i_{b2}, i_{c1}, i_{c2} respectively, for positive i_s as indicated in Figure 8A.

If the strain gage bridge is unbalanced in the opposite direction, i.e., i_s flows in the negative direction indicated on Figure 8A, i_{b1} is increased and i_{b2} is decreased:

$$\begin{aligned}
(i_{b1}) &= (i_{b1})_o + \Delta\, i_{b1} \doteq (i_{b1})_o + i_s \\
(i_{b2}) &= (i_{b2})_o - \Delta\, i_{b2} \doteq (i_{b2})_o - i_s \\
(i_{c1}) &= (i_{c1})_o + \Delta\, i_{c1} \\
(i_{c2}) &= (i_{c2})_o - \Delta\, i_{c2}
\end{aligned} \qquad (6)$$

where (i_{b1}), (i_{b2}), (i_{c1}), and (i_{c2}) represent the values of i_{b1}, i_{b2}, i_{c1}, and i_{c2} respectively, for negative i_s as indicated in Figure 8A.

Since Equation (3) holds for all values of i_{b1} and i_{b2} under consideration:

$$(i_{c1})_h - (i_{c1})_h \doteq \beta \left[(i_{b2})_h - (i_{b1})_h \right] \qquad (7)$$

$$(i_{c2})_\ell = (i_{c1})_\ell \doteq \beta \left[(i_{b2})_\ell - (i_{b1})_\ell \right] \qquad (8)$$

Expanding Equation (7) using Equation (5), the following relationship is obtained:

$$(i_{c2})_o + \Delta\, i_{c2} - \left[(i_{c1})_o - \Delta i_{c1} \right] \doteq \beta \left[(i_{b2})_o + i_s \right] - \beta \left[(i_{b1})_o - i_s \right] \qquad (9)$$

Rearranging terms:

$$(i_{c2})_o - (i_{c1})_o + \Delta\, i_{c2} + \Delta\, i_{c1} \doteq \beta \left[(i_{b2})_o - (i_{b1})_o + 2\beta\, i_s \right] \qquad (10)$$

Subtracting Equation (4) from Equation (10):

$$\Delta i_{c2} + \Delta i_{c1} \doteq 2\,\beta\, i_s \tag{11}$$

Equation (11) states that the sum of the changes in collector currents is essentially equal to the strain gage current multiplied by the factor 2β

Equation (8) can be expanded in the same manner as Equation (9) by using Equation (6), to yield the relationship given in Equation (11). Equation (11) is, therefore, a statement of the transfer function for the DC amplifier. The current gain of the DC amplifier is approximately 2β when $\beta_1 = \beta_2 = \beta$.

Temperature Stability[1]

The stability of the DC amplifier depends almost entirely upon two factors:

a. Selection of an optimum operating point for each of the transistors.

b. Careful matching of the transistors with respect to β and I_{co}.

The first of these factors minimizes the change in collector current of each transistor for a given change in temperature; the second factor tends to equalize the changes occurring with temperature in the two transistors (Q_1 and Q_2).

Selection of an optimum operating point for each transistor involved defining a current stability factor, the magnitude of which depended upon the incremental change in emitter current for a given incremental change in I_{co}. The base biasing resistors of Q_1 and Q_2 were chosen to minimize the current stability factor and, thus, maximize the temperature stability of the DC amplifier. In addition to determining the optimum bias point for the transistors, the change in collector currents with a given change in temperature was further reduced by connecting a thermistor, having the proper resistance-temperature characteristic, in parallel with the two transistors. The resistance of the thermistor decreases with an increase in temperature which tends to stabilize the collector currents.

The variable inductor is designed so that equal changes in Q_1 and Q_2 collector currents, within certain bounds, produce no effective change in the DC magnetizing force of the inductor. Therefore, if the characteristics of Q_1 and Q_2 (especially β and I_{co}) are carefully matched, a change in temperature produces approximately equal changes in the two collector currents. Since the DC magnetizing force remains essentially constant, the inductance of the tank circuit winding of the variable inductor also remains essentially constant.

Variable Inductor

General

The variable inductor is a toroidal saturable reactor. The inductor has four windings and two matched toroidal cores, as shown in Figure 9A. Each core is 0.500 inch OD, 0.312 inch ID, and 0.093 inch high. The core material is a ferrite manufactured by General Ceramics Corporation designated "Ferramic 0-2."

Referring to Figure 9B, winding 1-2 is wound on one core; winding 3-4 is wound on the second core. Windings 1-2 and 3-4 are matched, i.e., they each have the same number of turns. These two windings are connected in series with opposite polarities, as shown in Figure 9A. The series-connected windings are used in the tank circuit of the oscillator. Current flowing in one direction through these windings produces flux in opposing directions in the two cores. After windings 1-2 and 3-4 are on the cores, the cores are placed one on top of the other and control windings 5-6 and 7-8 are wound on both cores as a unit. The two control windings are also matched and their polarities are opposite.

The inductance which windings 1-2, 3-4 exhibit in the tank circuit of the oscillator is the incremental inductance. Incremental inductance is the inductance offered to AC by a coil wound on a specimen of ferromagnetic core material when the magnetic flux in the core is produced by an AC field superimposed on a DC field. The incremental inductance of the tank circuit winding of the variable inductor is directly proportional to the incremental permeability(2) of the core material. The incremental permeability is expressed as follows:

$$\mu_{\Delta} = \frac{\Delta B}{\Delta H} \tag{12}$$

where $\mu\Delta$ is the incremental permeability; ΔB is the change in flux density produced by the AC field; and ΔH is the change in magnetizing force, or the amplitude of the AC field.

The incremental permeability of the core material of the variable inductor in the TCCO depends upon the following factors in addition to the magnitudes of the AC and DC magnetizing forces:

a. The core material.

b. Stresses in the core material resulting from external forces.

c. The temperature.

The core material is fixed, so this factor becomes a constant. The material and techniques used in the encapsulation of the inductor were carefully selected in order to isolate the core from external forces; hence, this factor is not considered in the following analysis. Therefore, if the temperature is assumed to be constant, an expression can be developed which gives the inductance of the tank circuit winding as a function of the magnitudes of the applied AC and DC fields.

Transfer Function

Although the term "transfer function" may be a misnomer in view of its usual application, the term is here used to designate the relationship between the inductance of the tank circuit winding of the variable inductor and collector currents of the DC amplifier. The inductance L of the tank circuit winding can be expressed as follows:

$$L = \mu\Delta\, N^2\left(\frac{A}{\ell}\right)\ (4\pi \times 10^{-9})\ \text{Hy}$$

$$L \doteq \mu\Delta\, N^2\left(h \ln \frac{d_2}{d_1}\right)\ (2 \times 10^{-9})\ \text{Hy} \qquad (13)$$

where $\mu\Delta$ is the incremental permeability as previously defined.

N is the total number of turns in the tank circuit winding (i.e., the sum of the turns of windings 1-2 and 3-4).

A is the total cross-sectional area of core material (i.e., twice the area of one core) in square centimeters.

ℓ is the mean length of flux path in centimeters

h is the sum of the heights of the two cores or twice the height of one core in centimeters.

d_2 is the outside diameter of the core.

d_1 is the inside diameter of the core.

(NOTE: d_1 and d_2 must be expressed in the same units of measure.)

For a given inductor, all the factors in Equation (12) are fixed by the physical characteristics of the inductor except L and $\mu\Delta$. The inductance can, therefore, be expressed as a function of incremental permeability alone.

$$L = K_1 \mu \Delta \qquad (K_1 = \text{constant}) \tag{14}$$

The incremental permeability can be expressed as a function of the applied AC and DC fields. The magnitudes of these fields are functions of the currents producing the fields. The general equation for the magnetizing force developed by a given winding is:

$$H = \frac{F}{\ell} = 0.4\,\pi\,\left(\frac{N\,I}{\ell}\right) \qquad (15)$$

where H is the magnetizing force in oersteds (gilberts per centimeter).

N is the number of turns in the winding.

I is the current in the winding in amperes.

ℓ is the mean length of flux path in centimeters.

For a particular inductor, ℓ and N are fixed, so H can be expressed as a function of the current alone:

$$H = K_2\,I \qquad (K_2 = \text{constant}) \qquad (16)$$

The magnetizing force H, producing flux in the cores of the specific inductor in the TCCO, consists of the following components:

a. The DC magnetizing force developed by Q_1 collector current in control winding 7-8, and the combination of Q_2 collector current and the fixed bias current I_B in control winding 5-6.

b. The AC magnetizing force developed by the AC current in the tank circuit winding.

The component of the DC magnetizing force produced in control winding 7-8 opposes that produced in winding 5-6 since the two windings are wound with opposing polarities. Since the number of turns in each of these windings is equal, the effective DC magnetizing force is a function of the difference of the current magnitudes in the two windings. Referring to Equation (15) the DC magnetizing force is expressed as follows:

$$H_{DC} = H_b = K_2 (I_B + i_{c2} - i_{c1}) \tag{17}$$

The AC magnetizing force is given by the following expression:

$$H_{AC} = \Delta H = K_2 \, i_{AC} \tag{18}$$

where i_{AC} is the rms value of the AC current in the tank circuit winding.

From the definition of incremental permeability as expressed by Equation (12) (i.e., $\mu\Delta = \frac{\Delta B}{\Delta H}$), it appears that $\mu\Delta$ is a function of the AC magnetizing force ΔH , but independent of the DC magnetizing force H_b . It can be shown, however, that the magnitude of ΔB is dependent upon the DC magnetizing force H_b, which, in turn, makes $\mu\Delta$ a function of both ΔH and H_b .

The DC magnetizing force H_b produces a DC flux, which can be described as a biasing flux B_b, in the core of the inductor. When the AC magnetizing force is superimposed upon H_b, the result is effectively an increase in the biasing flux B_b. If the initial value of B_b is sufficiently near the flux density at saturation B_s, the superimposed AC magnetizing force ΔH produces relatively little increase in B_b; whereas, if the initial value of B_b is well below saturation, ΔH produces a relatively large increase in B_b. It follows logically, then, that the incremental flux density ΔB produced by a given AC magnetizing force ΔH is a function of the initial biasing flux density B_b.

From the definition of incremental permeability $\mu\Delta$, it can be seen that both a DC magnetizing force H_b and a superimposed AC magnetizing force ΔH must be present before the phenomena expressed as $\mu\Delta$ can exist. Assuming then, that both H_b and ΔH are applied initially, the limiting value of $\mu\Delta$ as ΔH is decreased to zero can be expressed as follows:

$$\lim_{\Delta H \to 0} \mu\Delta = \mu r \tag{19}$$

where μr is, by definition, the reversible permeability.[2]

Refer to Figures 12 and 13.

The reversible permeability is a finite limiting value of the incremental permeability as ΔH is reduced to zero. For low values of H_b, μr is a function of both H_b and the maximum value of ΔH applied. As H_b is increased so that B_b approaches saturation, μr is almost entirely a function of H_b, and it can be expressed as:

$$\mu r \doteq \frac{B_b}{H_b} \tag{20}$$

In view of the relationship between μr and $\mu\Delta$ given in (19) above, it appears logical that $\mu\Delta$ can be expressed in terms of μ_r . This expression is given below:

$$\mu\Delta = \mu_r + \nu\Delta\, H\Delta \tag{21}$$

where $\nu\Delta$ is $\dfrac{d\,\mu\Delta}{d\,H\Delta}$

$H\Delta$ is $\dfrac{\Delta H}{2}$

Substituting Equation (21) into Equation (14)

$$L = K_1 \ (\mu r + \nu\Delta \ H\Delta) \tag{22}$$

When H_b is made large enough so that μr is a function of H_b alone, then Equation (22) becomes:

$$L = K_1 \left(\frac{B_b}{H_b} + \nu\Delta \ H\Delta \right) \tag{23}$$

or, substituting Equation (17) into Equation (23):

$$L = K_1 \left[\frac{B_b}{K_2 \ (I_B + i_{c2} - i_{c1})} + \nu\Delta \ H\Delta \right] \tag{24}$$

Equation (24) accurately describes the relationship, which exists in the TCCO, between the inductance of the tank circuit winding of the variable inductor and the magnitudes of the DC control currents.

Temperature Stability

A characteristic of the core material is that there exists a limited region of incremental permeability of the material where the temperature coefficient of change of permeability is approximately zero in the temperature range between 25°C and 75°C. The limits of this region are determined experimentally as a function of the DC bias current I_B. I_B is then fixed at a value within this region which results in excellent temperature stability for the variable inductor.

As previously mentioned, stress due to external forces on the core affects the permeability of the material. For this reason, the encapsulating process used in packaging the variable inductor isolates the core from stresses which can result from volumetric changes of the encapsulating material with temperature.

Oscillator

Circuit Operation

The transistor Q_3 exhibits alternately high and low impedances during successive half cycles of oscillator operation. During the part of each cycle in which the transistor impedance is high, the battery supplies current to the series LC tank circuit. While the transistor impedance is low, it provides a conduction path for current from the tank circuit and the battery. By this switching action, the transistor controls the tank circuit excitation and thereby sustains oscillation.

Refer to oscillator circuit diagram, Figure 10A, and oscillator current and (measured) voltage wave shapes, Figure 11.

Switch S is closed at $t = 0$.

Due to inductance L, the initial tank circuit current is zero at $t = 0$. The voltage V_X at X (across capacitor C) with respect to ground is zero at $t = 0$.

The voltage V_B at B with respect to ground is zero at $t = 0$.

When switch S is closed, the battery current is determined, for a short but finite interval of time, by the resistance of R_9 and emitter-collector impedance of transistor Q_3. Since the voltage, V_B, is initially zero with respect to ground, the transistor starts conducting when a positive voltage is impressed at E. With the transistor conducting, the voltage V_E drops almost instantaneously to a low positive value with respect to ground.

At this point in time, the circuit enters operation Region I as indicated on Figure 11. Although the voltage at E at this instant is small, it is positive with respect to V_B, V_X, and ground. With transistor Q_3 conducting, base current starts to flow through R_8 producing a voltage across C_2 and C_1 in series, and starts charging capacitor C_2. Since C_2 is small compared to C_1, the voltage V_B starts increasing in a positive direction. For a finite period of time the voltage V_B is positive with respect to V_X and negative with respect to V_E. Simultaneously the current is increasing in L, causing the voltage V_X to increase in a positive direction. As the voltage V_X approaches V_B, current flow through C_2 reverses direction and drives B further positive. When the voltage V_B becomes equal to the voltage V_E, the circuit operation leaves Region I and enters Region II.

As the voltage V_B goes positive with respect to the voltage V_E, reverse bias is applied to the emitter-base junction of the transistor causing the impedance between E and B to increase to a high value. While B is positive with respect to E, the voltage V_E depends almost entirely upon the voltage drop across R_9 and the battery voltage. As the tank current swings through its maximum positive value, the voltage at E with respect to B reaches its maximum negative value. As the tank current decreases toward zero, the current through R_9, and hence the voltage-drop across R_9, decreases causing V_E to increase toward V_B.

While V_X is positive and increasing with respect to V_B, C_2 is charging through R_8 and the relatively high base-collector junction impedance of the transistor.

When the voltage V_E equals the voltage V_B, the circuit operation goes into Region III.

As V_B becomes negative with respect to V_E, the transistor base-collector junction impedance decreases rapidly, causing V_B to go further negative with respect to V_E until near maximum current is flowing through the transistor.

When the transistor emitter-collector impedance reaches its minimum value, the circuit operation goes into Region IV.

The battery current through R_9 and the transistor reaches a maximum value as the emitter-collector impedance reaches a minimum value. The low impedance between E and ground also allows the tank current to flow in a negative direction through the transistor. Since the transistor impedance is a low resistance, the battery current is practically limited by the value of R_9. The transistor current, therefore, reaches its maximum value at the same time the tank current reaches its maximum negative value.

While the circuit is operating in Region IV, V_E remains positive with respect to V_B; V_B remains positive with respect to ground, and V_X remains positive with respect to V_B. In this region, the voltage V_X goes from near its maximum positive value to a value at which the voltages V_X and

V_B are equal. When the voltage V_X equals the voltage V_B, the current in C_2 is zero. As V_X goes negative with respect to V_B, the circuit operation goes into Region I and the cycle through Regions I, II, III and IV repeats itself.

Transfer Function

Again, it is necessary to qualify the use of the term "transfer function." It is defined in this instance to be that function which states the relationship between the inductance of the tank circuit winding of the variable inductor and the frequency of the oscillator. Since the oscillator is operated Class B or Class C, a rigorous analysis of its operation is beyond the scope of this paper. Instead, an intuitive development of the transfer function follows.

Refer to the equivalent circuit, Figure 10B. The switching action of the transistor Q_3 is controlled by the base current, which is a function of the voltage V_B. A condition which is essential to Class B or Class C operation is that the voltage V_B must be positive with respect to V_E during a portion of each cycle. During this portion of the cycle, base current i_b is flowing in a negative direction, as indicated on Figure 10B, and the transistor emitter-collector impedance is quite high. Similarly, during another portion of each cycle V_B must be negative with respect to V_E. During this portion of the cycle, the base current i_b is flowing in a positive direction, as indicated in Figure 10B, and the transistor emitter-collector impedance is quite low. It has been determined experimentally that the optimum relationship between C_2 and R_8 at a given frequency fo is: $X_{c2} = R_8$. The ratio

C_2/C_1 is fixed at each frequency fo to produce minimum distortion. The maximum value of C_2/C_1 is 0.5. Resistor R_9 limits the transistor emitter-collector current during that portion of the cycle when V_B is positive with respect to V_E; therefore, R_9 is selected to obtain the desired operating point during this portion of the cycle.

An approximate equivalent circuit of the oscillator with the transistor replaced by a switch and two resistors is shown in Figure 10C. Class C operation is approximated when the switch is closed during half of each cycle and open during the remaining half cycle. The resistance R_A represents the emitter-collector resistance of the transistor when V_B is positive with respect to V_E. The resistor R_B represents the emitter-collector resistance when V_B is negative with respect to V_E. Capacitor C' represents the effective capacitance paralleling C_1 due to C_2.

The transfer function, as previously defined, must be valid for the range of values of L necessary to produce ±7.5% change in frequency from some center value. The form of the transfer function is:

$$f = F(L) \tag{25}$$

where f is the frequency of the oscillator.
F(L) is a function of L which states the relationship between f and L over the range given above.

Let:
$$C_1 + C' = C_T \tag{26}$$

Then:
$$X_L \doteq X_{CT} \tag{27}$$

Equation (27) must be true in the range of operation stated above.

It follows then:

$$\omega L \doteq \frac{1}{\omega C_T} \tag{28}$$

$$\omega \doteq \frac{1}{\sqrt{L C_T}} \tag{29}$$

$$f \doteq \frac{1}{2\pi \sqrt{L C_T}} = \left(\frac{1}{2\pi \sqrt{C_T}}\right) \left(\frac{1}{\sqrt{L}}\right) \tag{30}$$

which is the transfer function given in Equation (25).

The relationship between C_2 and C' depends upon the class of operation, the Q associated with L, and the effective Q of the $C_1 + C'$ combination. This relationship can be evaluated experimentally for a given set of circuit parameters, and in the range of operation required, can be stated as follows:

$$C' = K_3 C_2 \quad (K_3 = \text{constant}) \tag{31}$$

then $$C_1 + K_3 C_2 = C_T \tag{32}$$

and $$f = \left(\frac{1}{2\pi \sqrt{C_1 + K_3 C_2}}\right) \left(\frac{1}{\sqrt{L}}\right) \tag{33}$$

The constant K_3 is determined by the conditions stated above.

Temperature Stability

This oscillator exhibits excellent temperature stability when it is operated Class B or, preferably, Class C. The transistor functions as a switch, as explained above, so that if the magnitude of the driving signal remains above the value required for Class B operation, changes in transistor characteristics with temperature have little effect on the operating frequency.

Over-all Input-Output Characteristic

The relationship between oscillator frequency and magnitude and direction of the resistive bridge transducer current can be expressed, by combining the previously developed transfer functions for the DC amplifier, the variable inductor, and the oscillator.

Beginning with Equation (33) which gives oscillator frequency as a function of the inductance of the tank circuit winding of the variable inductor:

$$f \doteq \left(\frac{1}{2\pi\sqrt{C_1 + K_3 C_2}} \right) \left(\frac{1}{\sqrt{L}} \right) \tag{33}$$

For a particular design of the TCCO, the factor $\left(\frac{1}{2\pi\sqrt{C_1 + K_3 C_2}} \right)$ is fixed and can be represented as a constant K:

$$f \doteq \frac{K}{\sqrt{L}} \tag{34}$$

Substituting for L in Equation (34), the equation which expresses L as a function of the DC control currents i_{c1}, i_{c2} and I_B (i.e., Equation 24), the following expression is obtained:

$$f = \frac{K}{\sqrt{K_1 \left[\frac{B_b}{K_2 (I_B + i_{c2} - i_{c1})} + \nu\Delta\, H\Delta \right]}} \tag{35}$$

By setting i_{c1} and i_{c2} equal to their respective values when the DC amplifier input current is zero (i.e., $i_{c1} = (i_{c1})_o$; $i_{c2} = (i_{c2})_o$), Equation (35) becomes an expression giving the oscillator center frequency fo as a function of $(i_{c1})_o$ and $(i_{c2})_o$:

$$fo = \frac{K}{\sqrt{K_1 \left[\frac{B_b}{K_2 \left[I_B + (i_{c2})_o - (i_{c1})_o \right]} + \nu\Delta\, H\Delta \right]}} \tag{36}$$

Using Equations (10) and (11) to express DC amplifier collector currents as a function of the resistive bridge transducer current:

$$fo \pm \Delta f = \frac{K}{\sqrt{K_1 \left[\frac{B_b}{K_2 \left[I_B + (i_{c2})_o - (i_{c1})_o \pm 2\beta\, i_s \right]} + \nu\Delta\, H\Delta \right]}} \qquad (37)$$

The ± sign for 2B i_s gives the direction for i_s as indicated on Figure 8A. When the plus (+) sign is used for $2\beta\, i_s$, the + sign applies for Δf; the opposite is true for the negative (-) signs.

Redefining the constants in Equation (37):

$K = \left(\frac{1}{2\pi\sqrt{C_1 + K_3\, C_2}} \right)$ where K_3 depends upon the class of operation of the transistor in the oscillator. As the transistor operation approaches Class C, K_3 approaches 1.

$K_1 \doteq N^2 \left(h\, \chi_n \frac{d_2}{d_1} \right) (2 \times 10^{-9})$ Refer to Equation (13) for definition of symbols.

$K_2 = \frac{0.4\pi N}{\ell}$ Refer to Equation (15) for definition of symbols.

Figure 6 is a typical input-output characteristic for the TCCO with both experimental and calculated points for one prototype unit plotted. Figure 6 shows reasonably close agreement between calculated and measured points on the curve.

Performance

Laboratory and production prototypes of the TCCO have been constructed. Production prototypes are presently being subjected to extensive environmental testing. Tests which have been performed on laboratory prototypes indicate that the TCCO will be particularly suitable for airborne FM-FM telemetry applications. Although the circuit is simple from the standpoint of the number of components required, somewhat rigid production control is required in the selection of matched transistors for the DC amplifier and in the fabrication of the variable inductor.

Certain specifications and nominal performance data are summarized in Figure 14. Final performance specifications will be issued as soon as complete environmental test data are available.

Applications

It has been previously stated that the TCCO will find major applications in the field of airborne FM-FM telemetry. A Six Channel Airborne Strain Gage Calibrator and Excitation Unit has been designed to provide inflight calibration of six TCCO units and to provide isolated DC excitation for the associated resistance bridge transducers. The TCCO used in conjunction with the Calibrator and Excitation Unit provides a simple means of measurement of any physical phenomena for which resistance bridge transducer principles can be employed. These include, in addition to strain measurements, the measurement of pressure and acceleration phenomena. Certain types of temperature measurements can be made utilizing thermistors or other temperature sensitive elements. The TCCO can also be used with suitable adapters for low-level voltage measurements.

Summary and Conclusions

The TCCO is a practical device for providing a frequency modulated signal directly proportional to the appropriate excitation phenomena such as stress, pressure, acceleration, and, in some instances, temperature. The unit was designed primarily for airborne FM-FM telemetry applications. The TCCO is simple from the standpoint of number of parts. It consists of only three functional components: the DC amplifier, the variable inductor, and the oscillator. Each of the functional components was designed to provide maximum reliability and stability of operation. The use of transistors, a subminiature toroidal inductor, and printed wiring have made possible a compact, rugged, low-cost unit with extremely low power requirements. The TCCO provides a unique method of converting the output of a resistance bridge transducer to a proportional frequency modulated signal with the required accuracy and stability.

APPENDIX

Temperature Stability of the DC Amplifier

The temperature stability of the DC amplifier depends, in part, upon the selection and stabilization of the operating point of transistors Q_1 and Q_2. The operating point of each transistor was selected as follows:

1. Relatively low β transistors were selected for Q_1 and Q_2. The total current gain required for most applications of the TCCO-SG does not exceed 30. From Equation (11) in the text, the required β of each transistor is then approximately 15. By using transistors with β values in the range from 15 to 30, the stability problem was minimized.
2. Extensive measurements using Type 2N413 transistors showed that with an emitter current of approximately 2 milliamperes, the transistor input and output resistances remained essentially constant over the expected range of operation. For a fixed value of emitter current, the collector current is expressed as follows:

$$i_c + I_{co} = \alpha i_e = \left(\frac{\beta}{\beta + 1}\right) i_e \qquad (1)$$

Therefore, with i_e = 2 milliamperes and β in the range between 15 and 30, the range of collector currents is 1.88 to 1.94 milliamperes. Reference to the transistor manufacturer's data revealed that with this range of collector current, stable operation could be obtained by using a collector-to-base voltage V_c of approximately -4 volts.

3. R_{L1} and R_{L2} were fixed by the core size, the number of turns, and the wire size used for the DC control windings of the variable inductor. These factors were, in turn, fixed by the DC magnetizing force required for the variable inductor. $R_{L1} \doteq R_{L2} \doteq 300$ ohms. The voltage drop in each DC control winding due to the collector currents in paragraph 2 above lies in the range of 0.565 to 0.583 volts.

4. Neglecting the voltage drop across the emitter-base diode, the voltage drop across resistor R_{10} is then expressed as follows:

$$V_{R10} = V_T - (i_{c1}R_{L1} + |V_c|) \tag{2}$$

Where: V_T is the supply voltage

V_c is the collector-to-base voltage

Using the values from paragraphs (2) and (3) above, and a supply voltage of 12 volts:

$$V_{R10} = 12 - (0.575 + 4) = 7.425 \text{ volts} \qquad (3)$$

The current producing this voltage drop is the sum of the two emitter currents or approximately 4 milliamperes. The value of R_{10} was then obtained:

$$V_{R10}/2i_e = 7.425/0.004 = 2965 \text{ ohms} \qquad (4)$$

A value of 3000 ohms was used.

5. With the emitter current, emitter resistor, collector current, collector voltage and load resistance fixed, the ratio of R_1/R_4 (and R_2/R_6) was reasonably well established by the voltage required at the base of each transistor. A restriction in the selection of the value of R_1 (= R_2) was that the resistance of R_1 must be at least several times the value of input resistance of Q_1 in order to prevent shunting the input signal e_s.
6. It was assumed that a current stability factor could be expressed as the ratio of change in emitter current to the change in I_{co} for the temperature range of the amplifier.

$$S_I = \frac{\Delta i_e}{\Delta I_{co}} \doteq \frac{di_e}{dI_{co}} \tag{5}$$

By minimizing S_I, the current stability of the amplifier is maximized. In order to obtain a minimum value of S_I for the DC amplifier, S_I was expressed as a function of R_1 (= R_2) and R_4 (= R_6). (Refer to Figure 8B which is the equivalent circuit for Q_1 and its associated external resistances.) Again it was assumed that the voltage drop across the emitter-base junction was zero. Let $R = 2R_{10}$:

$$I_c = i_c + I_{co} \tag{6}$$

$$i_c = \alpha i_e \tag{7}$$

$$i_b = (1 - \alpha)\, i_e - I_{co} \tag{8}$$

$$I_b = i_e \left(\frac{R}{R_4}\right) \tag{9}$$

$$I_b + i_b = i_e \left[\frac{R}{R_4} + (1 - \alpha)\right] - I_{co} \tag{10}$$

The following relationships are obtained from the equivalent circuit, Figure 8B:

$$V_T = i_e \left(\frac{R}{R_4}\right) R_4 + \left\{ i_e \left[\frac{R}{R_4} + (1 - \alpha)\right] - I_{co} \right\} R_1 \tag{11}$$

Rearranging terms in (11):

$$V_T = i_e \left[\frac{RR_4 + RR_1 + (1 - \alpha)R_1R_4}{R_4} \right] - I_{co}R_1 \tag{12}$$

If I_{co} increases by a factor ΔI_{co}, the following holds:

$$V_T = (i_e + \Delta i_e) \left[\frac{RR_4 + RR_1 + (1 - \alpha)R_1R_4}{R_4} \right] - (I_{co} + \Delta I_{co})R_1 \tag{13}$$

Subtracting (12) from (13):

$$0 = \Delta i_e \left[\frac{RR_4 + RR_1 + (1 - \alpha)R_1R_4}{R_4} \right] - \Delta I_{co}R_1 \tag{14}$$

$$S_I = \frac{\Delta i_e}{\Delta I_{co}} = \frac{R_1R_4}{RR_4 + RR_1 + (1 - \alpha)R_1R_4} \tag{15}$$

For β values in the range 15 to 30, the factor $(1 - \alpha)$ assumes values in the range 0.063 to 0.032. Since negligible error was introduced by discarding the factor $(1 - \alpha)\ R_1R_4$:

$$S_I = \frac{\Delta i_e}{\Delta I_{co}} \doteq \frac{R_1R_4}{R(R_1 + R_4)} = \frac{R_1R_4}{2R_{10}(R_1 + R_4)} \tag{16}$$

From Equation (16), for a fixed value of R_{10}, S_I was minimized by selecting R_1 and R_4 to be the smallest values which satisfied the previously stated criteria for these resistors. The minimum value of R_1 was obtained experimentally; the corresponding value for R_4 was selected to obtain the desired collector current. Since $R_1 = R_2$ and $R_4 = R_6$, all values of bias resistors were determined by the above analysis.

7. The minimum value of S_I obtainable with linear bias circuit elements was achieved as indicated above. A means was then sought for further reducing the current stability factor. A satisfactory method was found which required the use of a nonlinear compensating element shunting transistors Q_1 and Q_2. Thermistor RT 1 was selected so that:

$$i_t \doteq K + (\Delta i_{c1} + \Delta i_{c2}) \tag{17}$$

Where: i_t is the current through the thermistor RT 1.

Δi_{c1}, Δi_{c2} represent the change in i_{c1} and i_{c2}, respectively, for a given increase in temperature.

If the resistance-versus-temperature characteristic of the thermistor is selected so that the relationship in Equation (17) is satisfied over the temperature range of operation, the current stability of the amplifier is appreciably improved. The change in resistance required for the thermistor from some initial value at an initial temperature To was expressed as a function of change in I_{co} of the transistors. Since change in I_{co} is, in turn, a function of temperature, the resistance-versus-temperature characteristic required of the thermistor was obtained.

The change in collector current with a change in temperature ΔT was expressed as follows:

$$\Delta i_c = \alpha \Delta i_e + \Delta I_{co}$$
$$\Delta i_c = \left(\alpha \frac{\Delta i_e}{\Delta I_{co}}\right) + 1 = \alpha\, S_I + 1 \qquad (18)$$

Substituting Equation (18) in Equation (17):

$$i_t = i_{to} + (\alpha S_I + 1)(\Delta I_{co1} + \Delta I_{co2}) \qquad (19)$$

Where: i_{to} is the value of thermistor current at the lowest temperature in the range under consideration (i.e., To).

The resistance of RT 1 at To is:

$$R_{RTo} = \frac{V_c + i_c R_L}{i_t} = \frac{V_T - 2 i_e R_{10}}{i_t} \qquad (20)$$

The required change in resistance of RT 1 (for a temperature change ΔT) expressed as a function of change in I_{co} is as follows:

$$\Delta R_{RT} = \frac{V_T - 2i_e R_{10}}{(\alpha S_I + 1)(\Delta I_{co1} + \Delta I_{co2})} \tag{21}$$

When $I_{co1} \doteq I_{co2}$, Equation (30) simplifies to:

$$\Delta R_{RT} \doteq \frac{V_T - 2i_e R_{10}}{2(\alpha S_I + 1)(\Delta I_{co})} \tag{22}$$

Using the following relationship:

$$I_{co} \doteq I_o \left(e \frac{K'V_c}{T} - 1 \right) \tag{23}$$

Where: I_o is the intrinsic collector-base current flowing with reverse bias on the collector-base junction; V_c is the collector-base voltage; T is temperature in ^{o}Kelvin-absolute.

Then:

$$\Delta I_{co} = (I_{co})_{T1} - (I_{co})_{To}$$

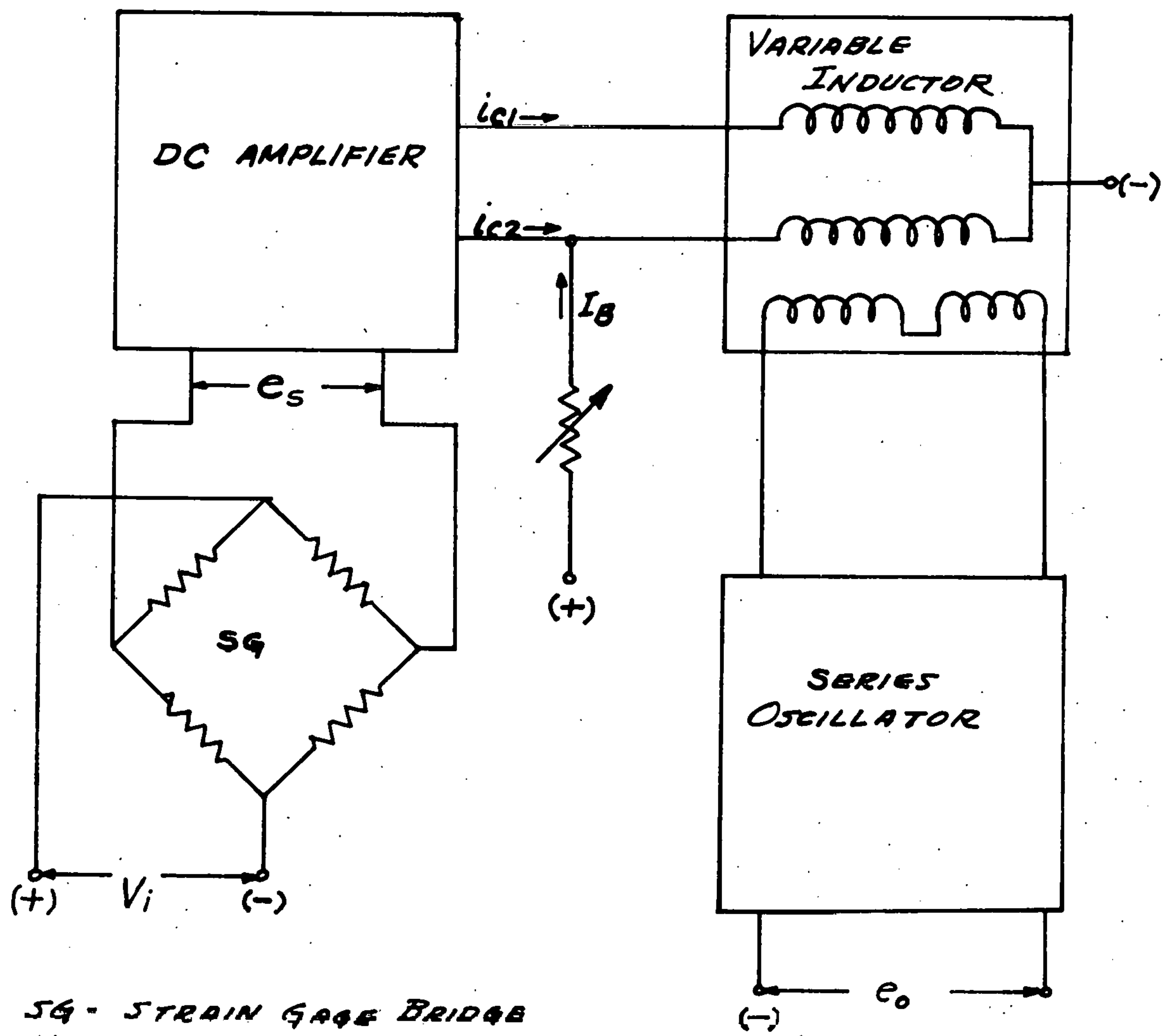
$$\Delta I_{co} = I_o \exp K'V_c \left(\exp \frac{1}{T_1} - \exp \frac{1}{To} \right) \tag{24}$$

$$\Delta I_{co} = K \exp \frac{1}{T_1} - \left(\exp \frac{1}{To} \right)$$

Substituting (24) in (22):

$$\Delta R_{RT} \doteq \frac{V_T - 2i_e R_{10}}{2K(\alpha S_I + 1)(\exp \frac{1}{T_1} - \exp \frac{1}{To}} \tag{25}$$

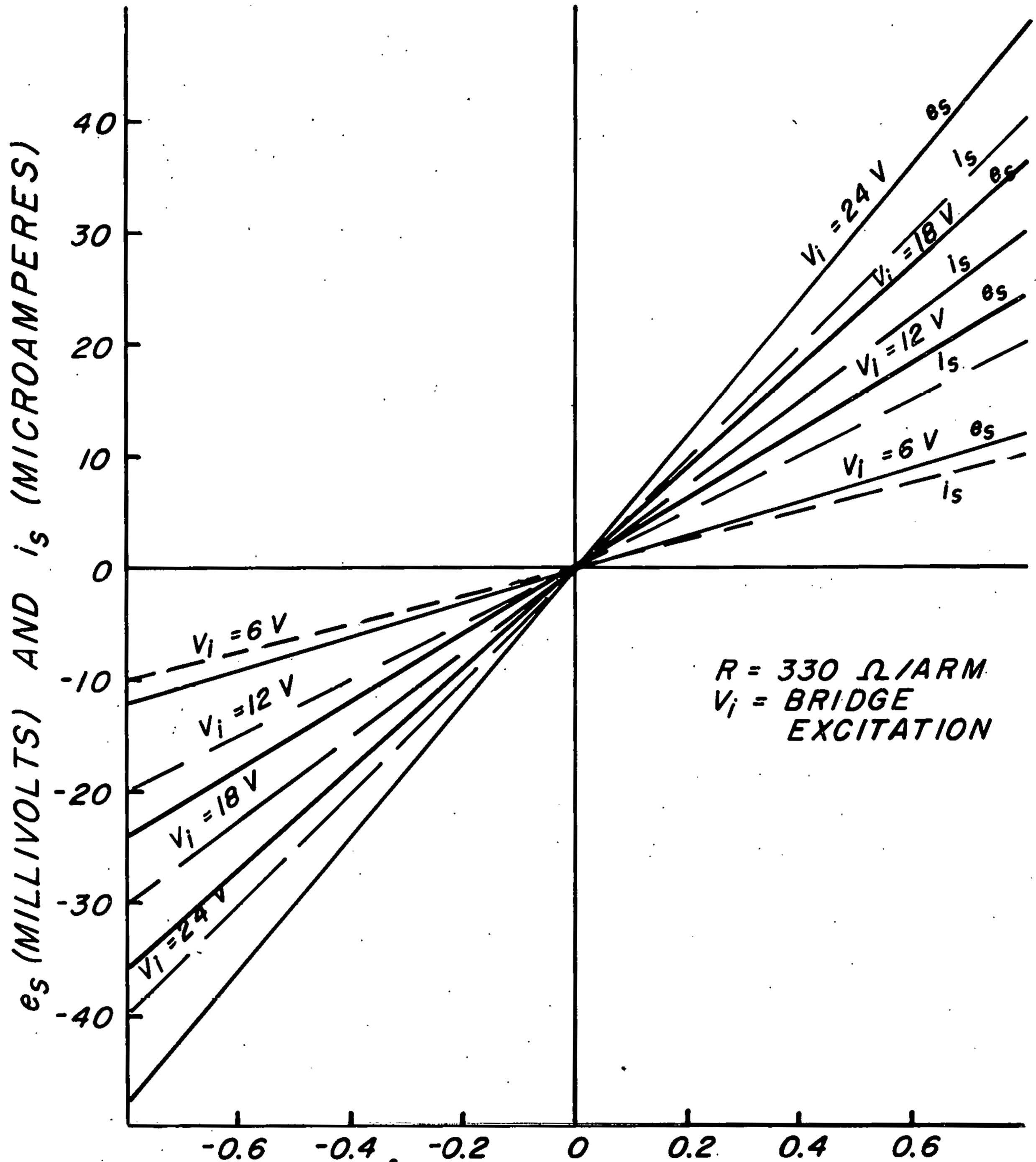
From Equation (25) a characteristic curve can be constructed allowing R_{RT} to vary as a function of the parameter T. This can be expanded into a family of curves by assigning different values to K.



SG - STRAIN GAGE BRIDGE
V_i - STRAIN GAGE EXCITATION VOLTAGE
e_S - STRAIN GAGE OUTPUT VOLTAGE
e_o - OSCILLATOR OUTPUT VOLTAGE

FIGURE 1
BLOCK DIAGRAM - STRAIN GAGE AND TCCO-SG

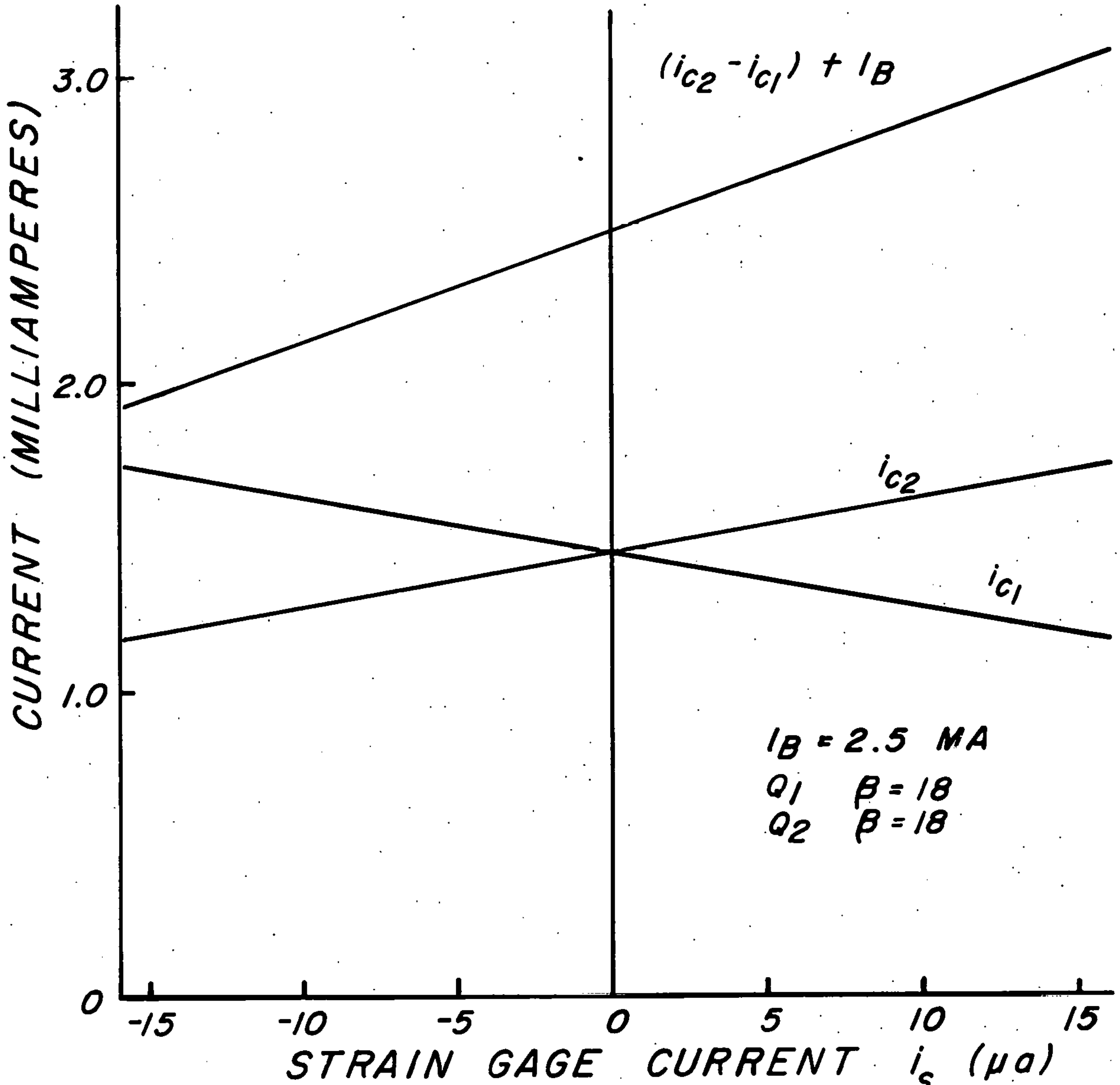
890 039



$\left(\frac{\Delta R}{R}\right)$ IN PERCENT

ONE ACTIVE ARM
TYPICAL CHARACTERISTIC STRAIN GAGE BRIDGE
FIGURE 2

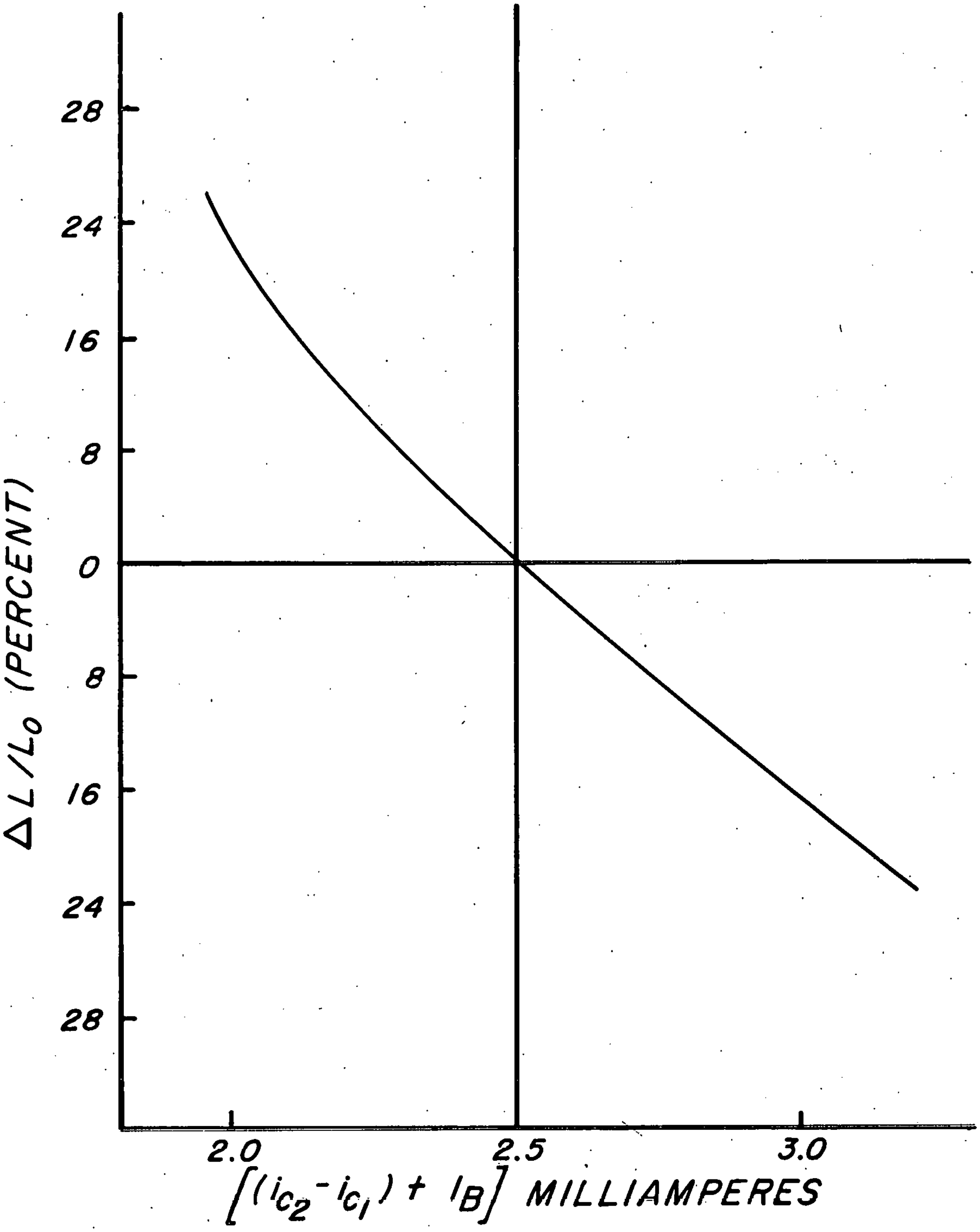
890 040



DC DIFFERENCE AMPLIFIER
TRANSFER CHARACTERISTIC

FIGURE 3

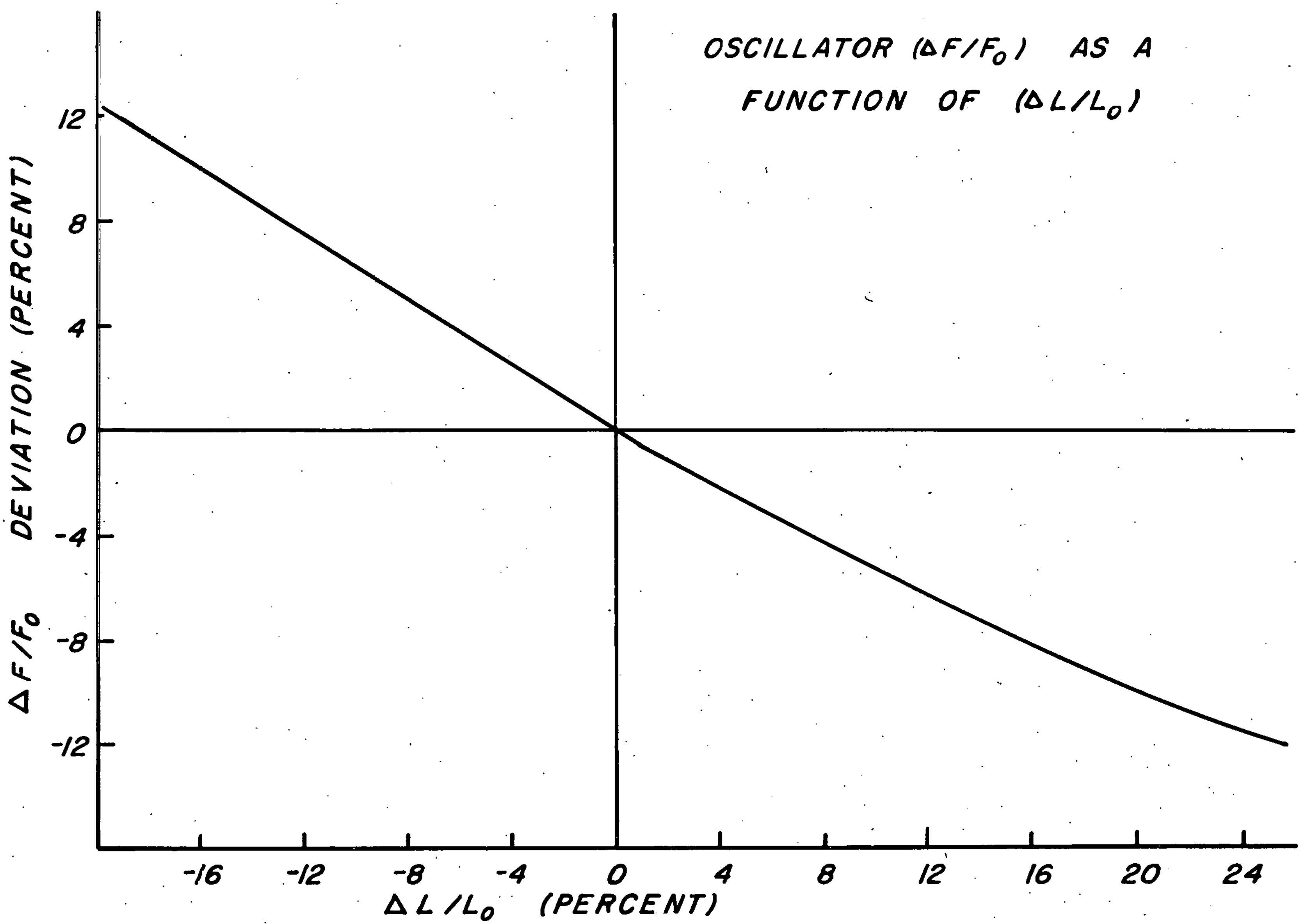
890 041



TYPICAL CHARACTERISTIC
VARIABLE INDUCTOR

FIGURE 4

890 042



OSCILLATOR ($\Delta F/F_o$) AS A FUNCTION OF ($\Delta L/L_o$)
$\Delta F/F_o$ DEVIATION (PERCENT)
12
8
4
0
-4
-8
-12
-16
-12
-8
-4
0
4
8
12
16
20
24
$\Delta L/L_o$ (PERCENT)

FIGURE 5

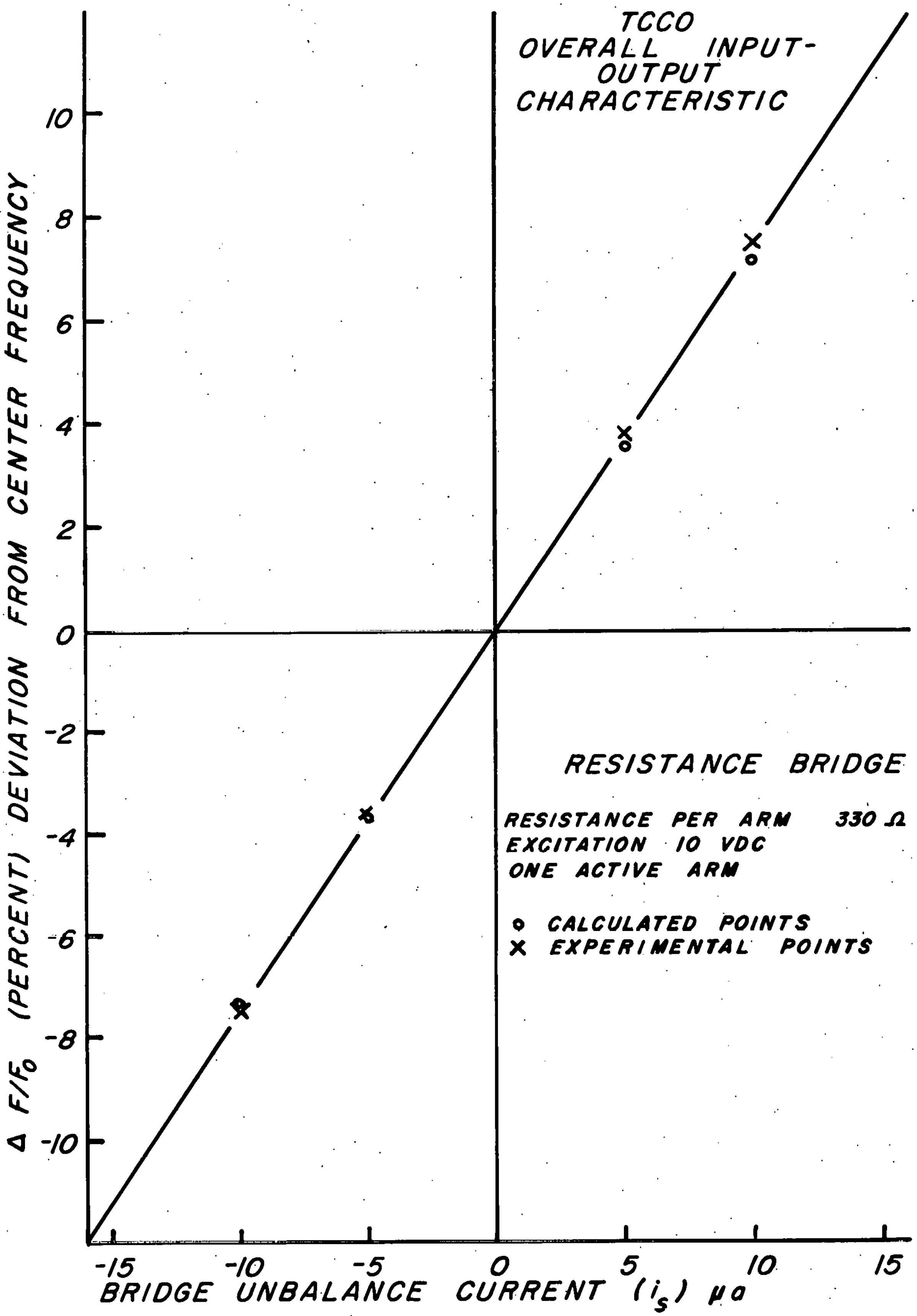
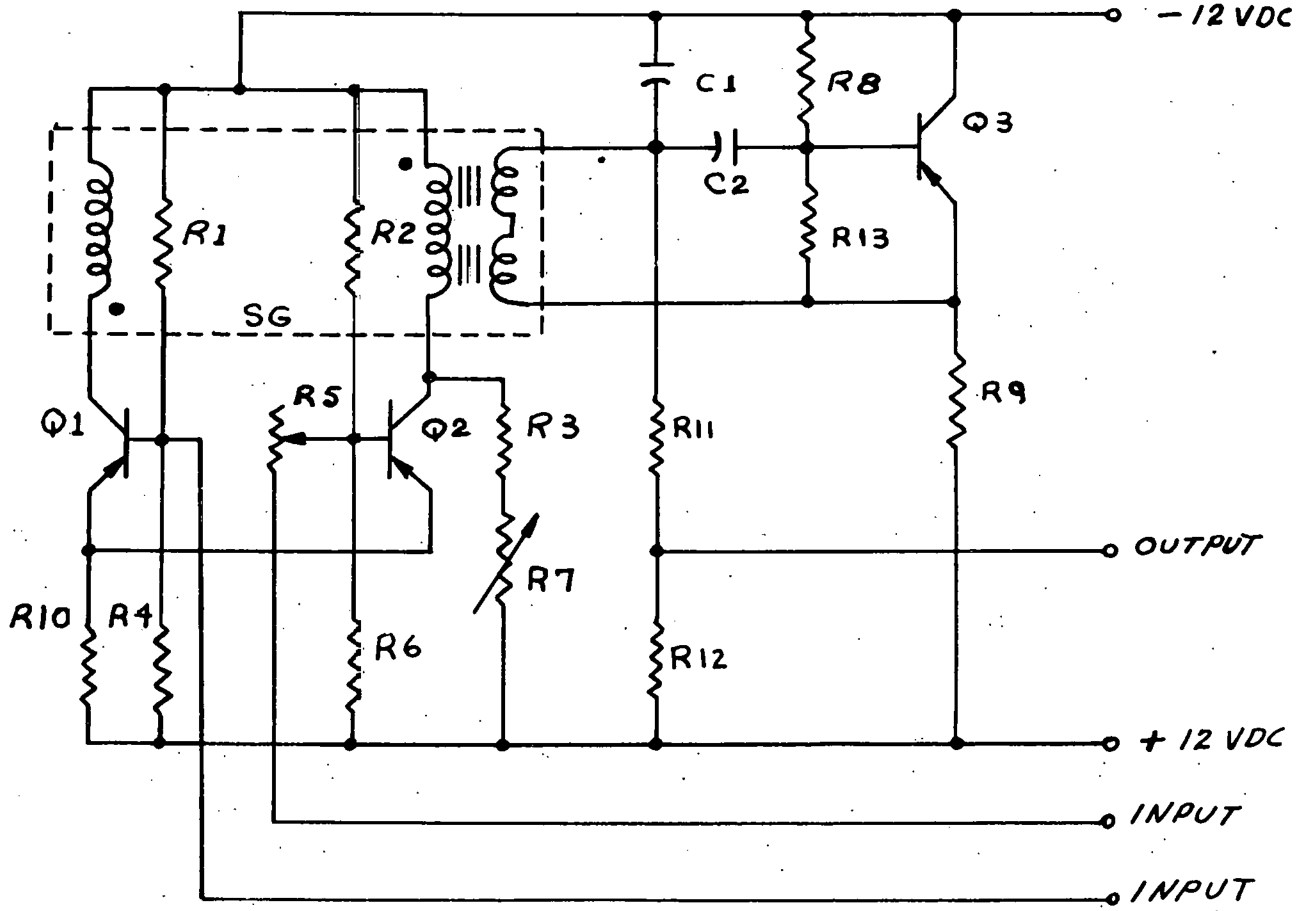

TCCO
OVERALL INPUT-OUTPUT
CHARACTERISTIC
RESISTANCE BRIDGE
RESISTANCE PER ARM 330 Ω
EXCITATION 10 VDC
ONE ACTIVE ARM
o CALCULATED POINTS
X EXPERIMENTAL POINTS
Δ F/F₀ (PERCENT) DEVIATION FROM CENTER FREQUENCY
10
8
6
4
2
0
-2
-4
-6
-8
-10
-15
-10
-5
0
5
10
15
BRIDGE UNBALANCE CURRENT (i_s) μa

FIGURE 6



SCHEMATIC DIAGRAM
TCCO-SG-MOD. 4 STRAIN
GAGE OSCILLATOR
FIGURE 7

890 045

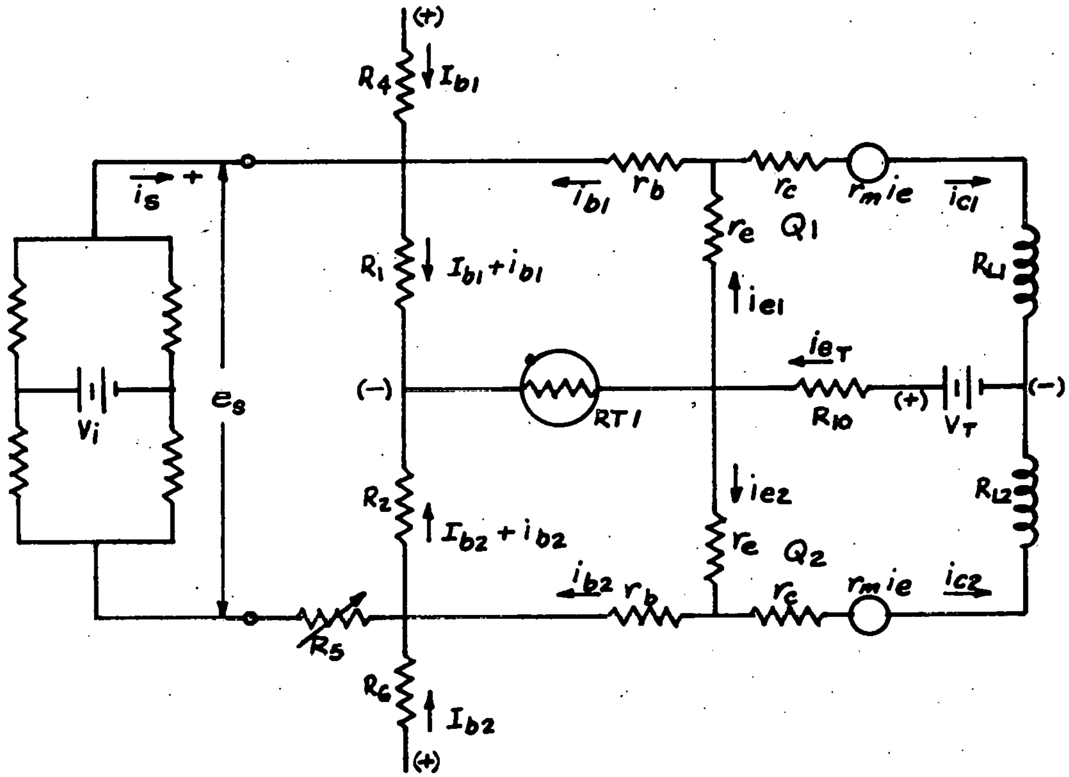
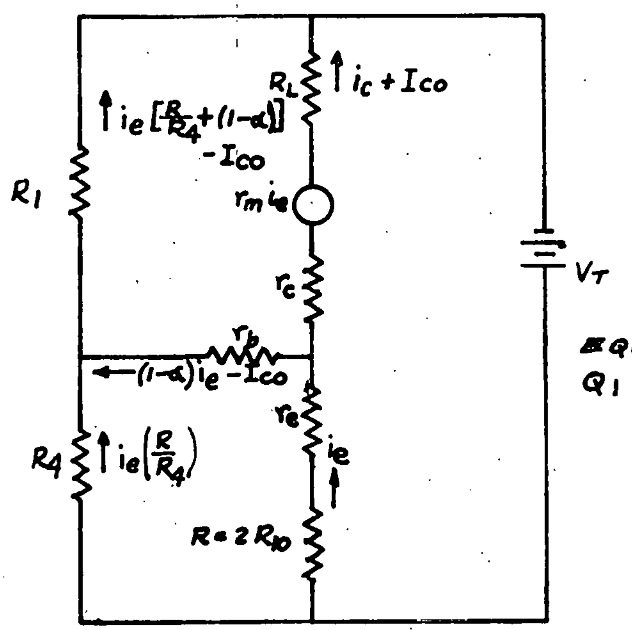


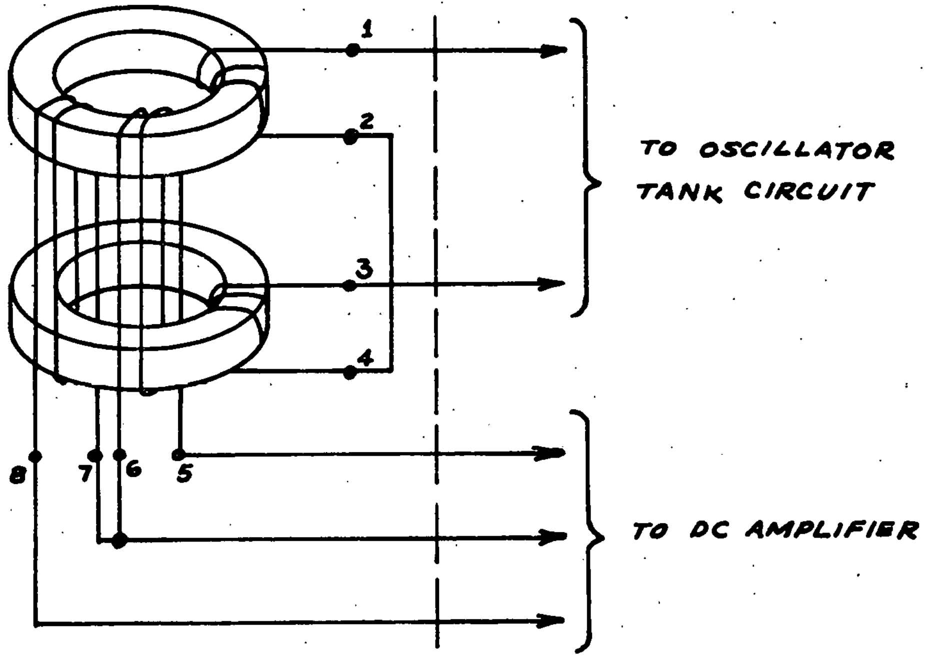
FIGURE 8A

EQUIVALENT CIRCUIT - STRAINGAGE AND DC AMPLIFIER

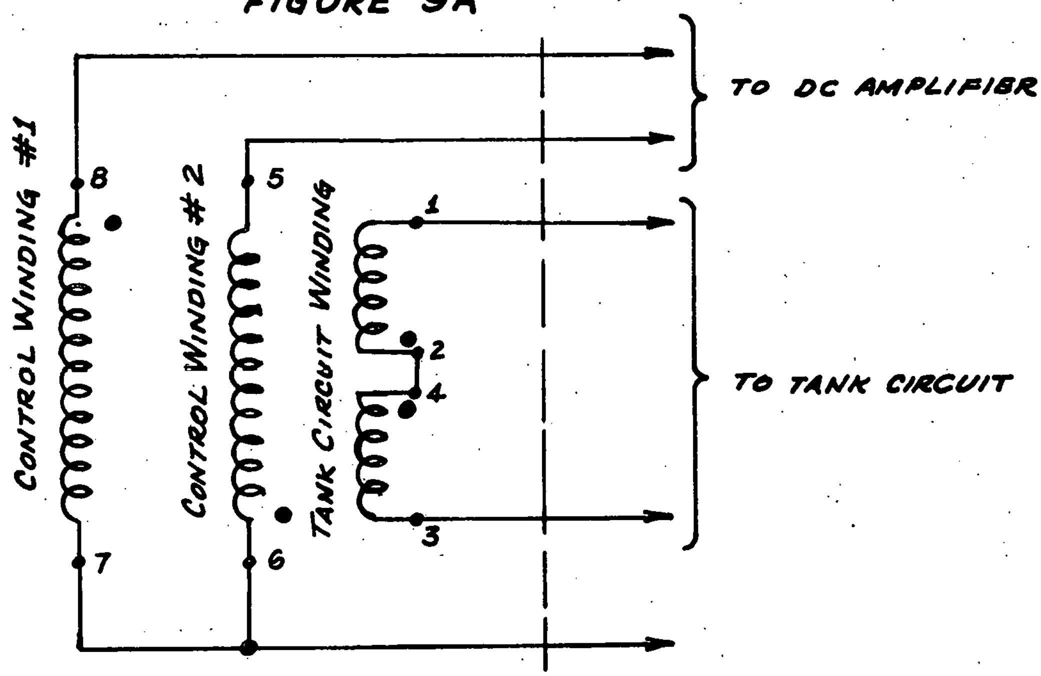


EQUIVALENT CIRCUIT
Q1 TRANSISTOR

FIGURE 8B



CONSTRUCTION DETAIL
FIGURE 9A



INDUCTOR
SCHEMATIC DIAGRAM
FIGURE 9B

890 047

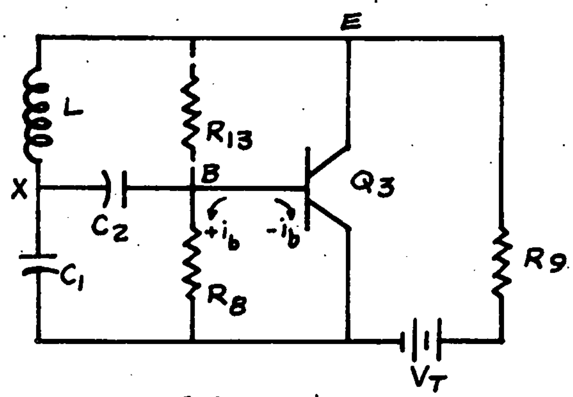


FIGURE 10A
OSCILLATOR SCHEMATIC

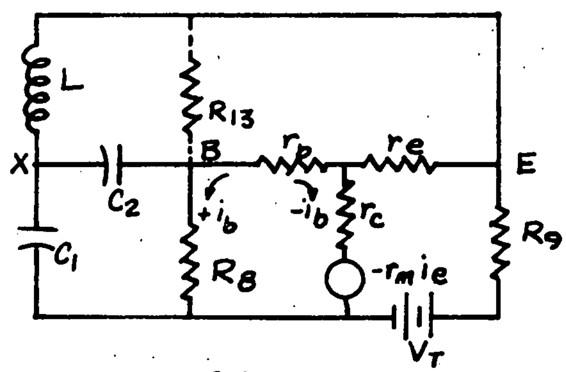


FIGURE 10B
OSCILLATOR EQUIVALENT CIRCUIT

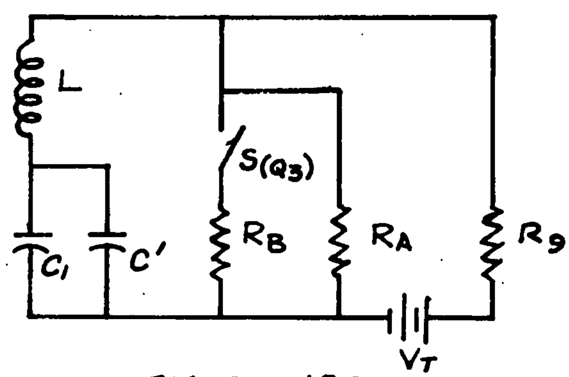


FIGURE 10C
APPROXIMATE EQUIVALENT CIRCUIT

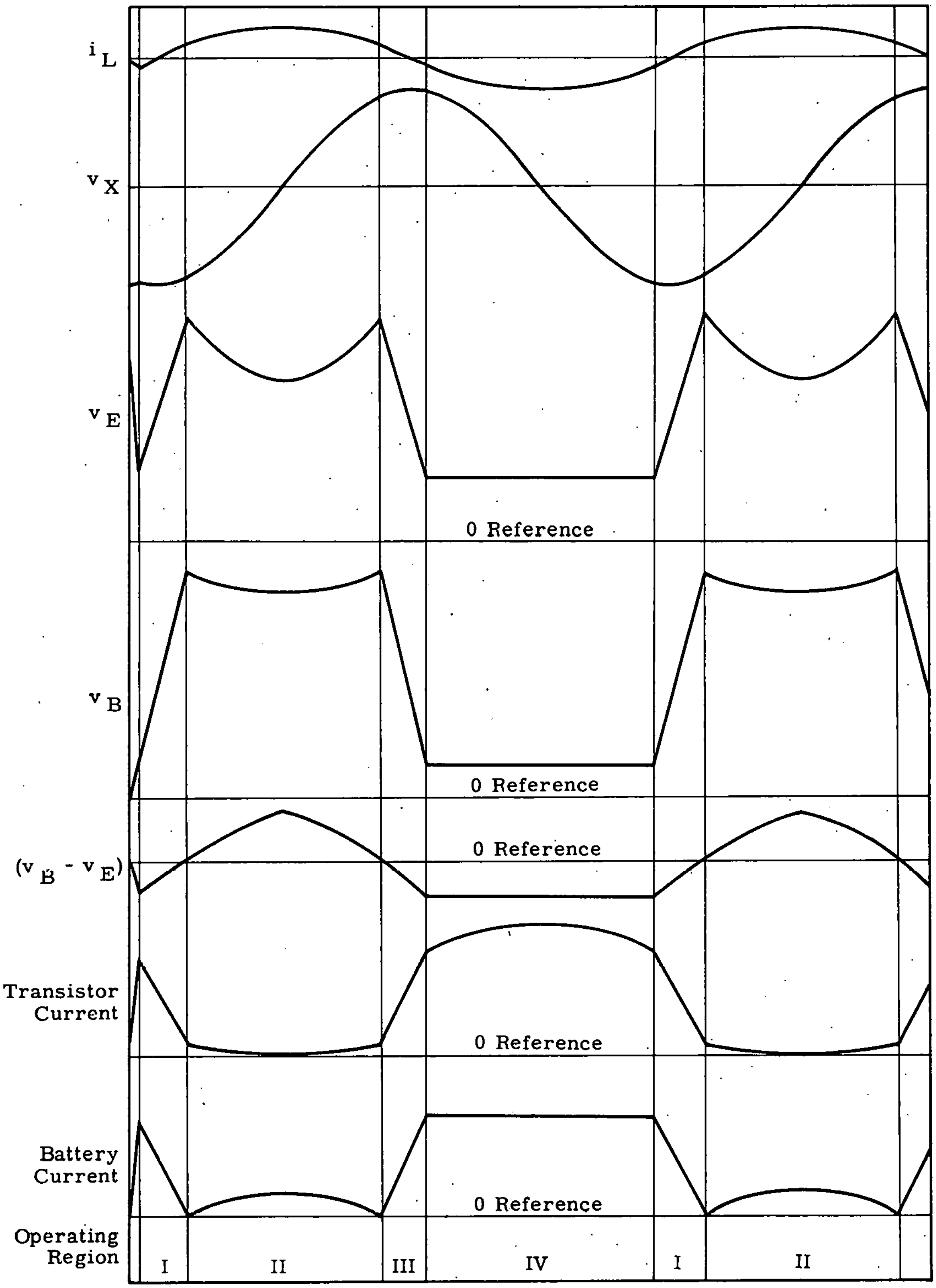


Figure 11 Wave Forms, Oscillator

890 049

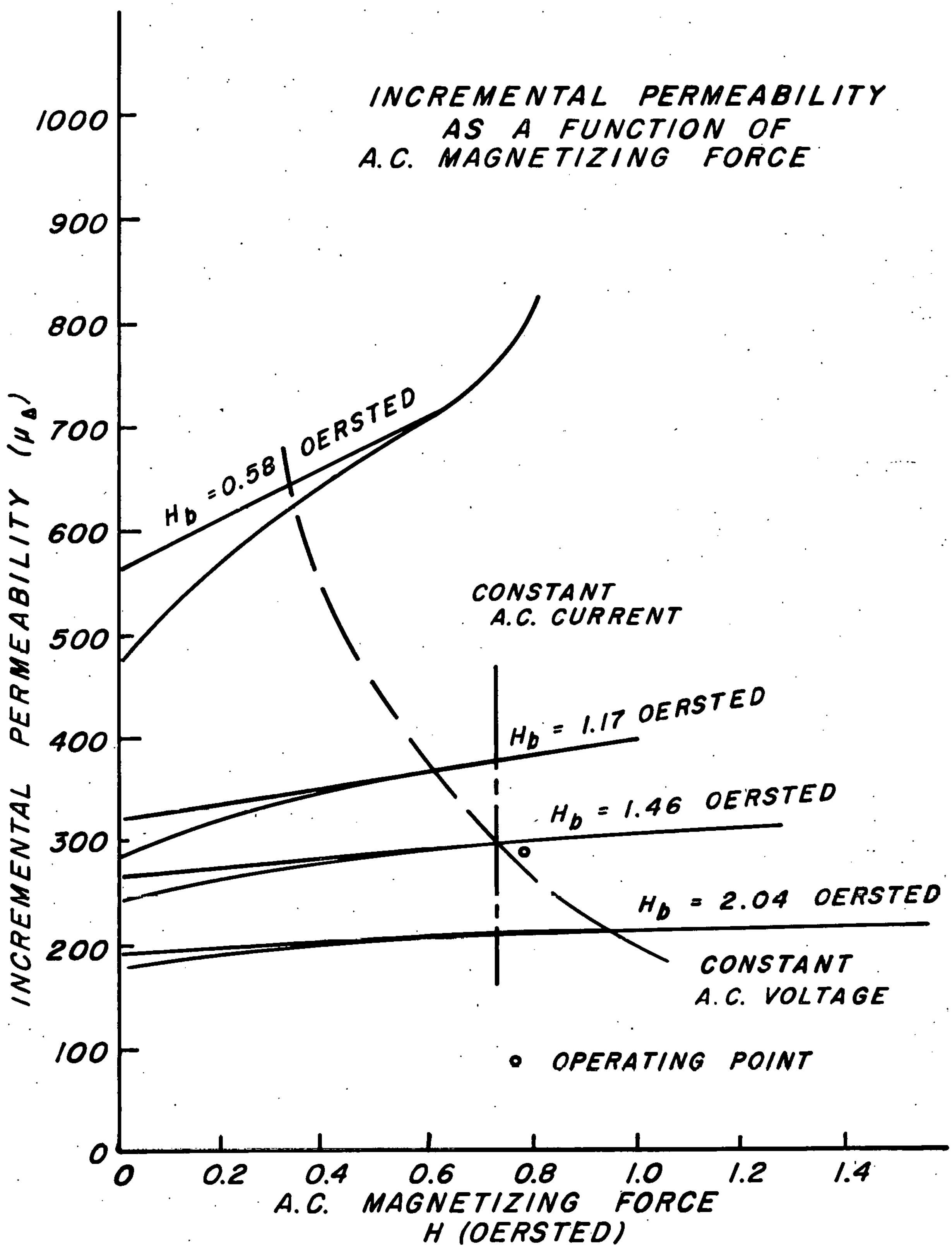

INCREMENTAL PERMEABILITY
AS A FUNCTION OF
A.C. MAGNETIZING FORCE
1000
900
800
700
600
500
400
300
200
100
0
INCREMENTAL PERMEABILITY (μ_Δ)
H_b = 0.58 OERSTED
CONSTANT
A.C. CURRENT
H_b = 1.17 OERSTED
H_b = 1.46 OERSTED
H_b = 2.04 OERSTED
CONSTANT
A.C. VOLTAGE
OPERATING POINT
0
0.2
0.4
0.6
0.8
1.0
1.2
1.4
A.C. MAGNETIZING FORCE
H (OERSTED)

FIGURE 12

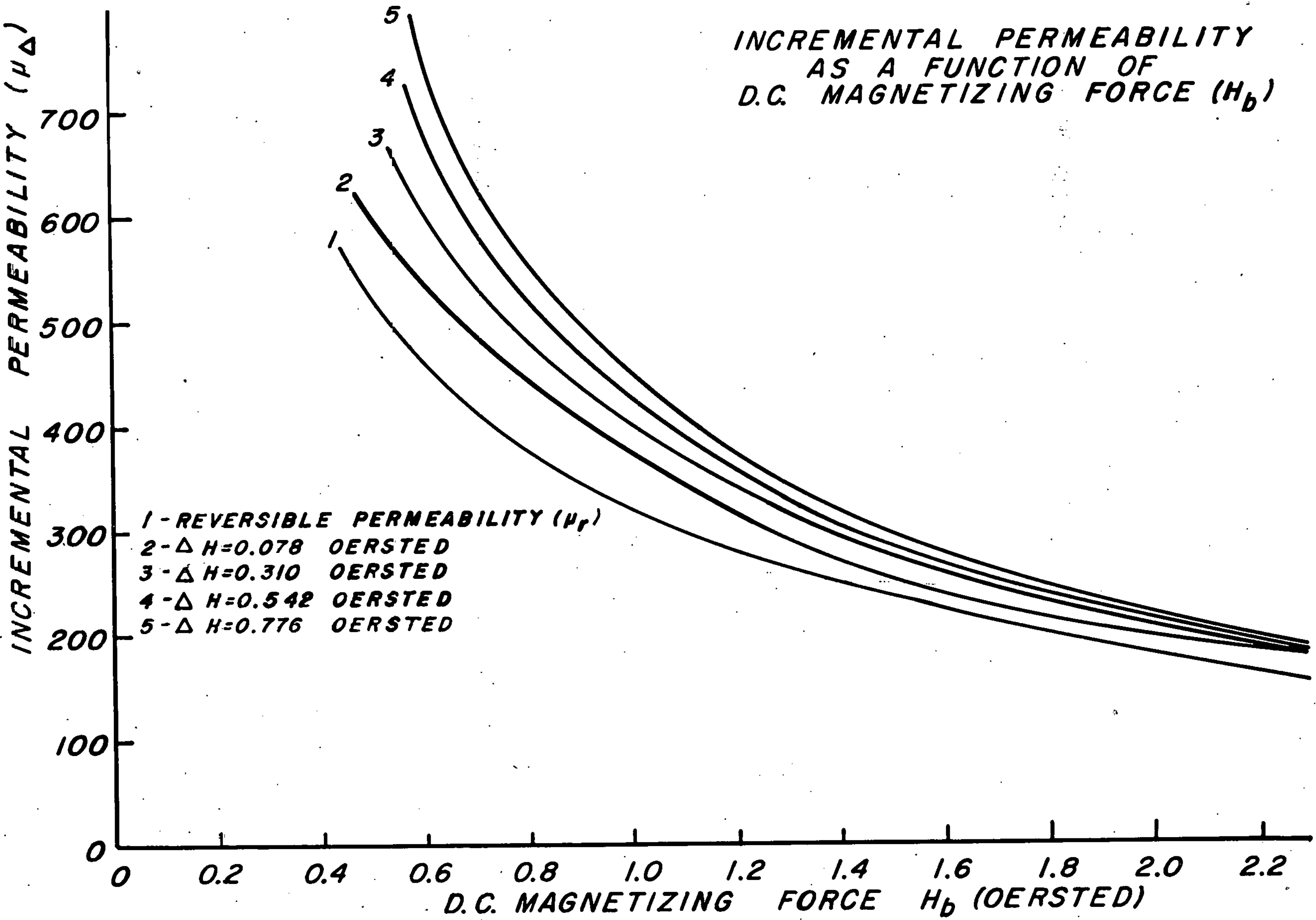

INCREMENTAL PERMEABILITY
AS A FUNCTION OF
D.C. MAGNETIZING FORCE (H_b)
INCREMENTAL PERMEABILITY (μ_Δ)
700
600
500
400
300
200
100
0
1 - REVERSIBLE PERMEABILITY (μ_r)
2 - Δ H=0.078 OERSTED
3 - Δ H=0.310 OERSTED
4 - Δ H=0.542 OERSTED
5 - Δ H=0.776 OERSTED
1
2
3
4
5
0
0.2
0.4
0.6
0.8
1.0
1.2
1.4
1.6
1.8
2.0
2.2
D.C. MAGNETIZING FORCE H_b (OERSTED)

FIGURE 13

FIGURE 14

PERFORMANCE DATA AND SPECIFICATIONS TCCO

PERFORMANCE DATA

Center Frequencies Available:	IRIG Band No. 4 through Band No. 18 (0.96 KC to 70 KC)
Power Required:	10 ma nominal at 12 volts ±0.5%.
Input:	DC excited, four-arm resistance bridge. Bridge excitation must be isolated from TCCO power supply.
Sensitivity:	±10 microamperes produces ±7.5% deviation from center frequency. Refer to Figure 2. With 330 ohm bridge, one active arm, 12 volts excitation: ±0.4% strain produces ±7.5% frequency deviation.
Output Signal Amplitude:	2.5 volt nominal across 70K load.
Output Signal Distortion:	Less than 1.5% total harmonic distortion over bandwidth.
Frequency Response:	Flat within 0.5 db to twice IRIG recommended frequency response (modulation index of 5).
Output Signal Linearity:	Within ±1% of bandwidth from best straight line through ±7.5% bandwidth points.
Amplitude Modulation of Output Signal:	Less than 10% over bandwidth when calculated as follows:

$$AM = \frac{Emax - Emin}{Emax + Emin} X\ 100$$

Nominal Operating Temperature Range:	70°F to 150°F
Frequency Drift:	Less than ±1% of center frequency over a two hour period after 15 minutes warm-up within nominal operating temperature range.

NOTE: The data listed above is average for all protypes tested.

PHYSICAL CHARACTERISTICS

Overall Length:	2-13/16 inches	
Overall Width:	1-11/16 inches	
Overall Height:	3/4 inch	
Overall Weight:	3.1 ounces	
Vibration:	To 10 g at 500 cps	
Shock:	500 g all planes, both directions)	Causes less than 2% of B. W. frequency shift.
Acceleration:	100 g all planes, both directions) 20 seconds duration)	

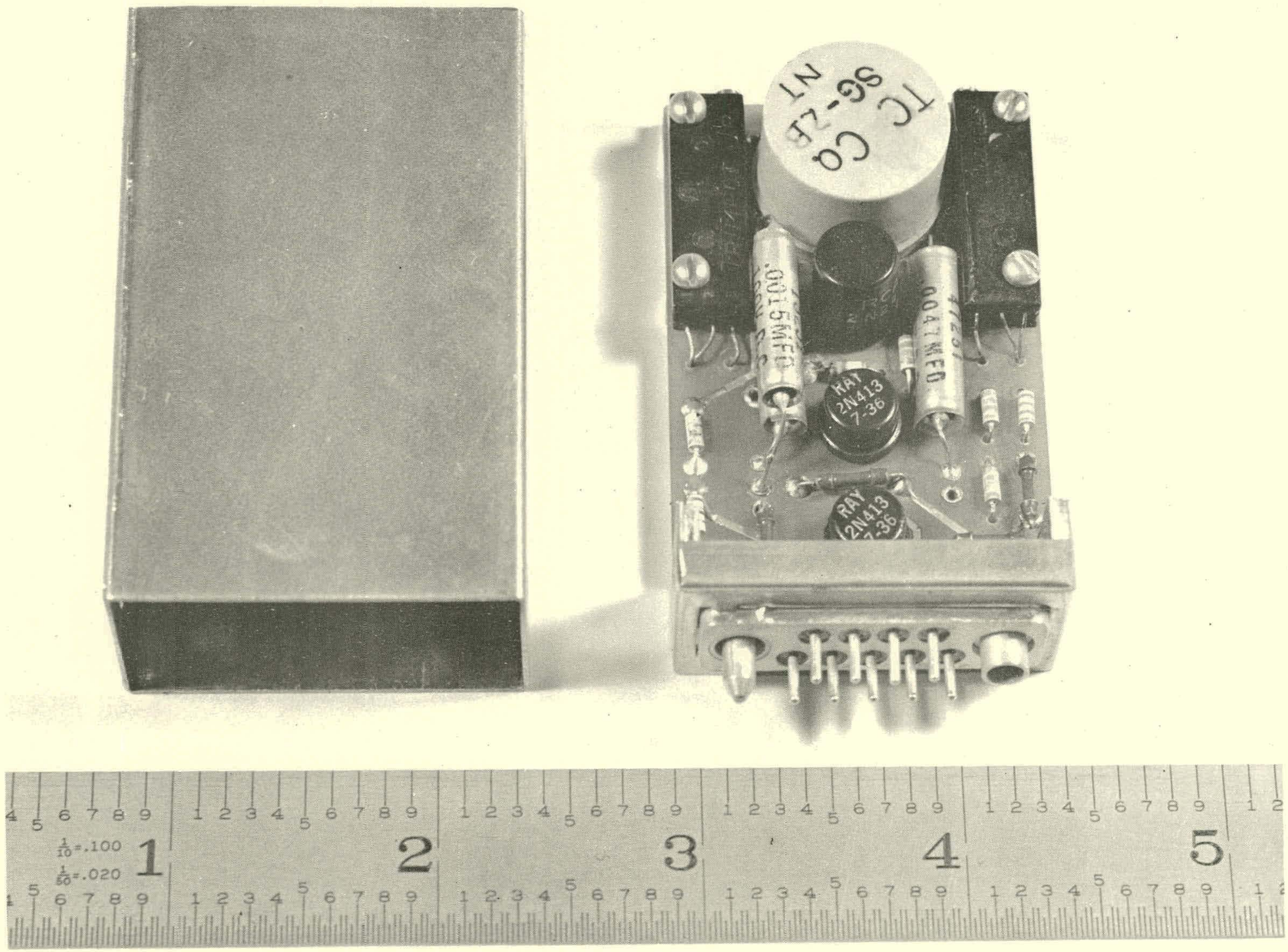


Figure 15

890 053

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