TRANSISTOR PHASE SHIFT OSCILLATORS

by

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Chapter 1.

A TRANSISTOR PHASE SHIFT OSCILLATOR

(1.1) INTRODUCTION.

This is an era of scientific development. There is little doubt about that. Great volumes of literature on new scientific findings are being published daily. The periodicals devote much space to detailing the newer discoveries of man. One can hardly be unaware of the forward strides of present day scientists, nor can he fail to be impressed with the magnitude of the findings of these men of science.

So enormous are the obstacles encountered and overcome by the scientist that one is impelled to think only in terms of general problems. Thus he tends to forget the multitude of details which must be overcome in the solution of the more general problems.

This paper is devoted to a detail problem, the design of a transistor phase shift oscillator.

It is inevitable that with the development of the transistor most of the standard vacuum tube circuits will be developed for transistors. Among the most common circuit types seen in electronic equipment is the oscillator. There are many types of oscillators, all having certain desirable characteristics. For a specific application one may be more desirable than another for one or more reasons. For low frequency applications the phase shift oscillator affords a simple, inexpensive design.

This paper is concerned with a transistor version of the phase shift oscillator. Particular emphasis will be placed on the low frequency and audio frequency ranges. Two basic types of oscillators, both of the phase shift variety, are discussed. They will be referred to as the "shunt R" and "shunt C" oscillators, respectively. They are shown in figure (1.1) and figure (1.2). The shunt R oscillator is particularly useful for the very low frequency regions while the shunt C oscillator is more useful in the high audio region. A more thorough treatment of the shunt R oscillator will be made. In the higher frequency areas the shunt C oscillator may not be as desirable as one of other types such as the "Wein Bridge", the "Twin T", or even the more common inductance-capacitance type oscillators.

(1.2) THE PROBLEM.

In designing the transistor phase shift oscillator for low frequencies the first problem that one encounters is that of the relatively (to the vacuum tube) low impedances associated with the transistor. The frequency of the phase shift oscillator is inversely proportional to the RC combination and since the impedance levels must be kept low

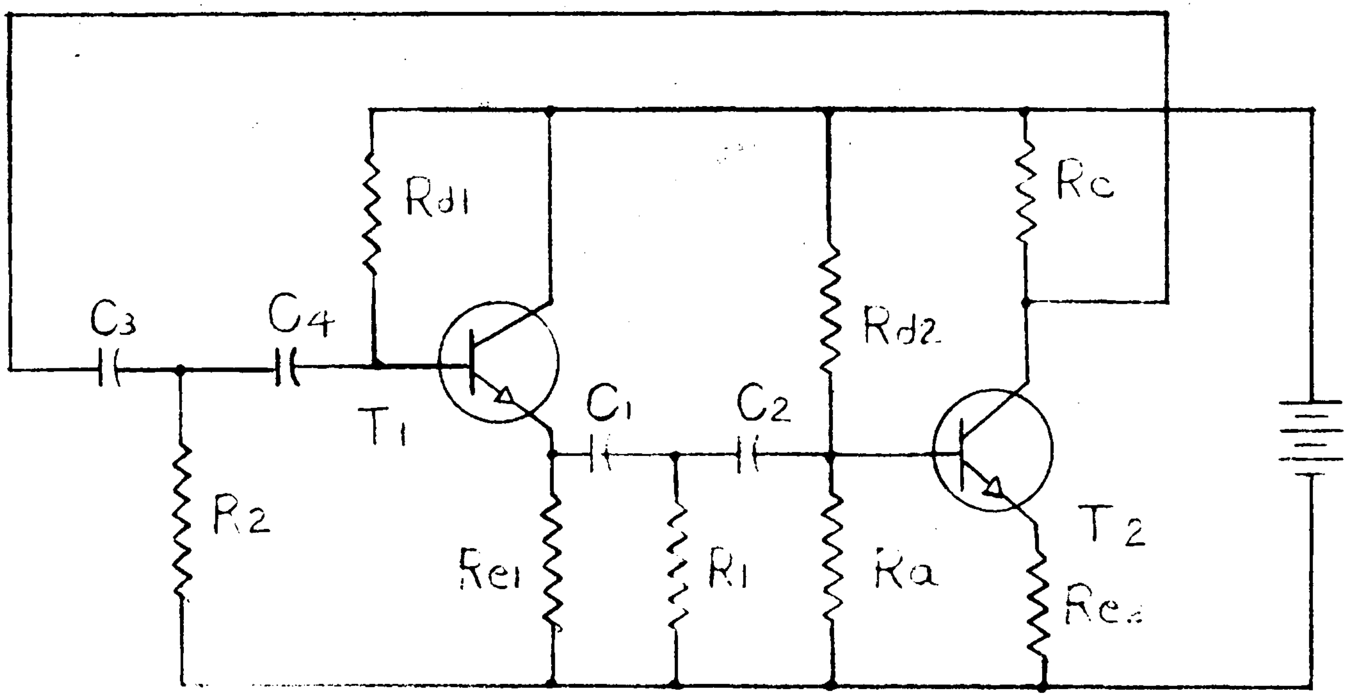


Fig. 1.1. Schematic diagram of the shunt R phase shift oscillator described in this thesis.

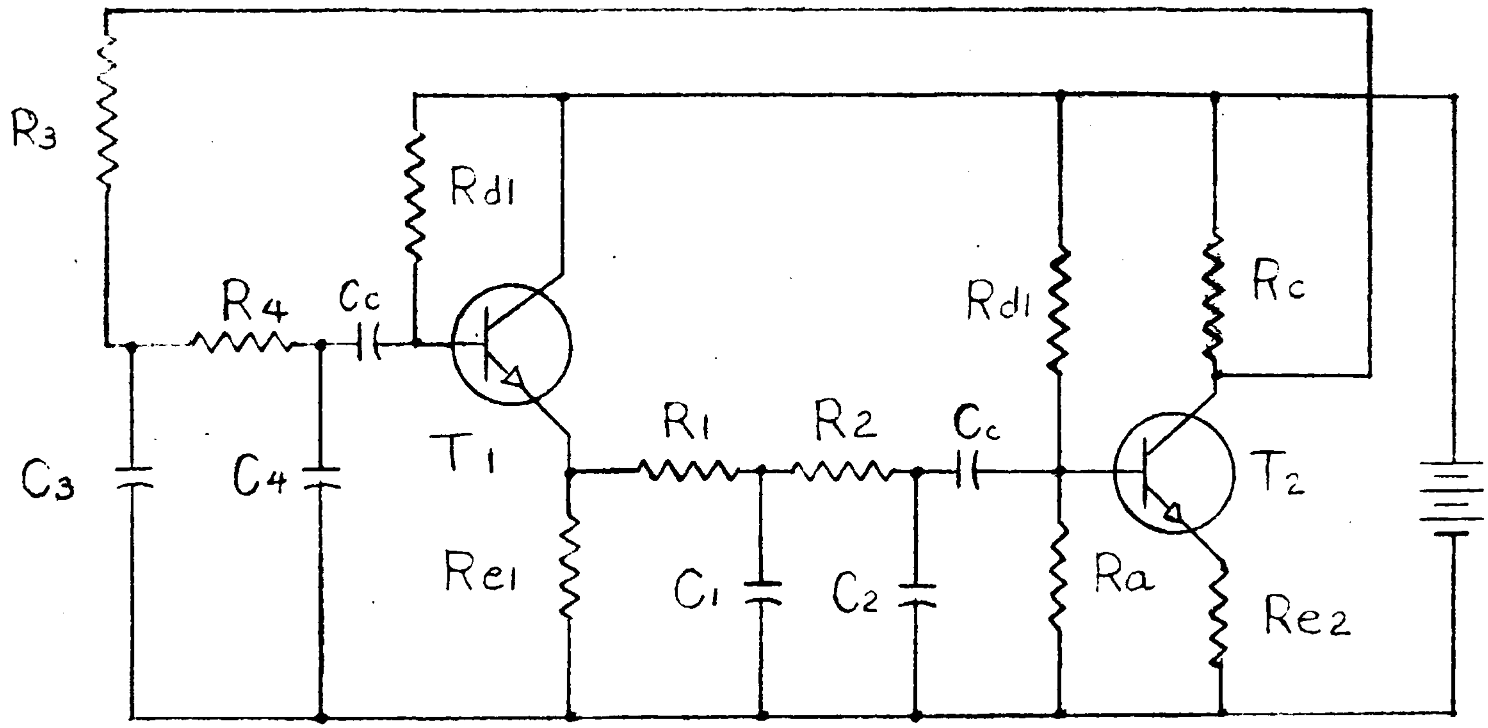


Fig. 1.2. Schematic diagram of the shunt C phase shift oscillator described in this thesis.

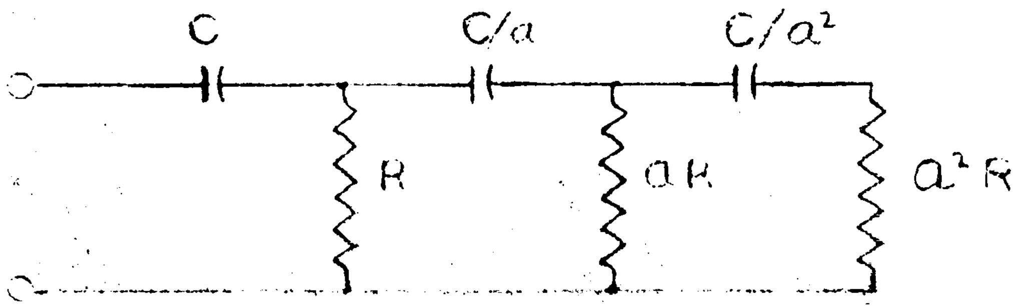
there is a definite limit on the R values that can be used. Another serious problem is that the input impedance of a transistor may be so low as to affect the frequency of the oscillator. In this respect the design problems are similar to the design of a radio frequency amplifier wherein the input capacitance of the vacuum tube must be taken into account.

In the design criteria developed in this paper the input and output impedances of the transistor amplifiers are included as a part of the phase shift circuits and are included in frequency calculations.

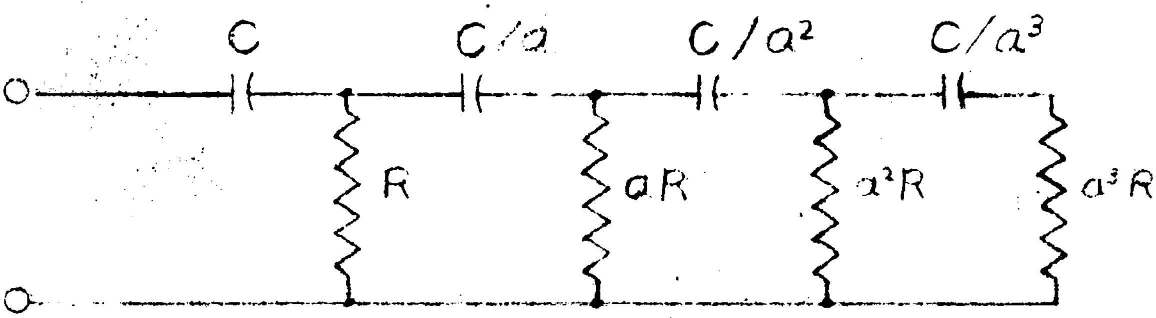
(1.3) PHASE SHIFT OSCILLATOR LITERATURE.

Phase shift oscillators have just about become the standard low frequency oscillator. Many good papers have been written on the vacuum tube phase shift oscillator. In an article in the Proceedings of the Institute of Radio Engineers[1] Peter G. Sulzer derives the gain and frequency equations for impedance tapered phase shift oscillators. By tapering the impedances; that is, to increase the impedance level of successive sections of a ladder network, as in figure (1.3), the voltage attenuation can be reduced while still obtaining the required 180° phase shift. This enables the

[1] Peter G. Sulzer, "Tapered Phase Shift Oscillator" Proc. IRE., vol. 36, Oct. 1948, pages 1302-06.



(a)



(b)

Fig. 1.3. Tapered phase shift ladder networks showing (a) a three section and (b) a four section ladder.

designer to make an oscillator out of a low mu triode vacuum tube.

In this article Mr. Sulzer presents curves showing the voltage attenuation plotted against the tapering factor. He also shows the advantages of using a four section network over a three section network.

An article which is almost the dual of the one above appeared in the August 1956 issue of Electronic Engineering[2]. The authors derive the gain and frequency expressions of a transistor phase shift oscillator from current considerations, rather than from the normal voltage approach. An impedance taper is presented for current which is just the inverse of the voltage taper. That is, successive sections are lower in impedance than were the preceeding ones.

The authors designed two basic oscillator types. One was the single transistor phase shift type, similar to the single stage vacuum tube phase shift oscillator. The other was a two transistor oscillator using the 0° phase shift philosophy.

The article stated that the maximum frequency for the single stage oscillator was in the neighborhood of two kilocycles per second.

There is one other good article on vacuum tube phase shift oscillators in Electronic Equipment[3]. This article

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[2]D. E. Hooper and A. E. Jackets, "Current Derived Resistance-Capacitance Oscillators using Junction Transistors, Electronic Engineering, Aug. 1956, page 333-337.

[3]D. L. Wardelich, "Optimum Phase Shift Oscillator Design" Electronic Equipment, Vol. 4, April 1956, page 38-42.

covers a cathode follower type single stage phase shift oscillator as well as the more common phase shift oscillators. An oscillator is constructed using a cathode follower by inverting the phase shift circuit giving an effective 360° phase shift at the oscillator frequency. This circuit is shown in figure (1.4).

R. P. Turner[4] constructed the circuit of figure (1.5). This is a simple single stage transistor oscillator very similar to that described above. It was constructed in the laboratory and worked very well over a small range of frequencies. It required a transistor with a high current gain in order to maintain oscillations. This circuit had a bad frequency drift with temperature change. In one instance it stopped oscillating when heated, requiring a readjustment of the emitter resister to restart oscillation.

None of the above articles on transistor oscillators ended with useable design equations. This thesis develops two particular types of phase shift oscillators of the 180° phase shift type. The design equations and procedure are developed. It is shown that with the two types of oscillators a frequency range of about 10 to 500,000 cycles per second is obtainable. It is shown that for the low and medium frequencies the oscillators described have good frequency stability and low distortion, primarily because the gain requirements

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[4] R. P. Turner, "Transistor Phase-Shift Oscillator", Radio and T. V. News, Vol. 4, April 1956, page 108.

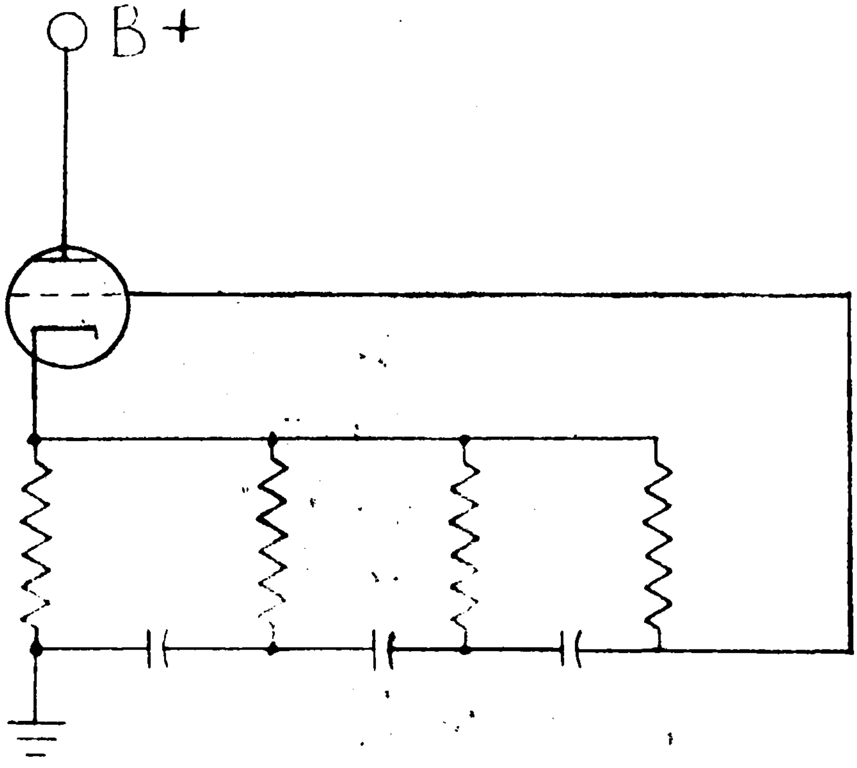

Fig. 1.4. Schematic diagram of a cathode follower phase shift oscillator.

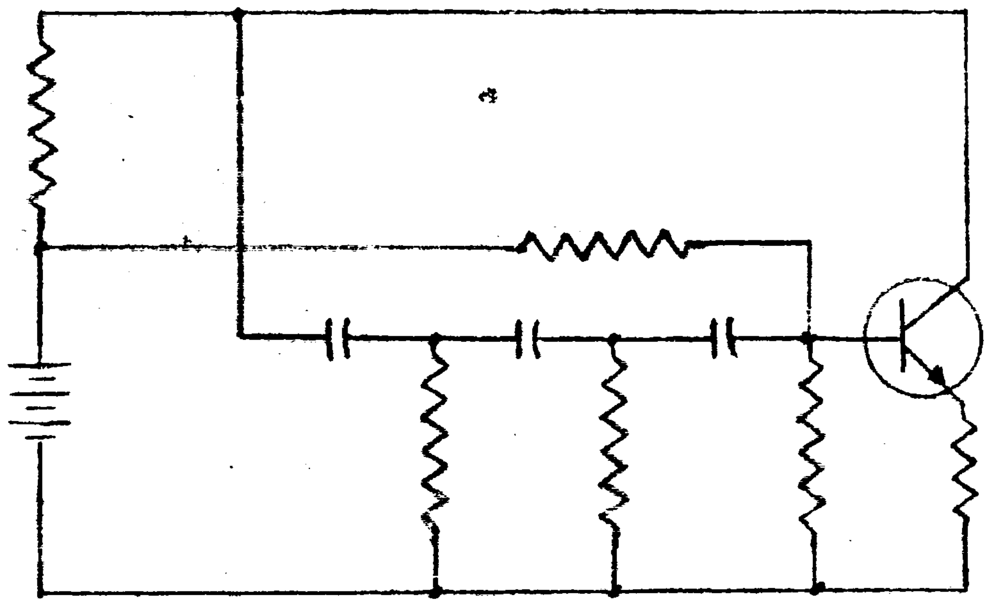

Fig. 1.5. Schematic diagram of a single transistor phase shift oscillator.

are not great and as a result current feedback can be used. This tends to maintain the same quiescent operating level even with temperature changes.

In chapter two the equivalent circuits for the oscillators are drawn and the equations for gain and frequency derived.

Chapter three describes the tests made on the two oscillators. The test results are given and compared with the calculated values.

Chapter four outlines a complete design procedure based on techniques discovered in testing the design equations.

Chapter five presents a summary of the thesis and an outline of the design procedure.

Chapter 2.

DERIVATION OF EQUATIONS FOR THE FREQUENCY AND GAIN REQUIREMENTS OF THE OSCILLATORS

(2.1) INTRODUCTION.

This chapter is devoted to the derivation of the equations of gain and frequency of the two oscillators.

In deriving these equations conditions are imposed for simplifying the expressions which have become a part of the design procedure. These conditions are necessary in order to make the design equations useable.

(2.2) ANALYSIS OF AN OSCILLATOR.

The mathematics of an oscillator can be summarized by the following statement: for a sine wave oscillation to occur, at some frequency, f, there must be a closed loop feedback circuit in which the total loop gain is one and the phase shift is equal to 360°.

This statement can be represented by a block diagram such as figure (2.1), where $KG(jw) = 1/\underline{360^\circ}$. The phase shift requirements in $G(jw)$ can be somewhat eased by making K equal to a minus gain. That is $K = -K_1 = K_1/\underline{180^\circ}$. Then $G(jw) = K_2/\underline{180^\circ}$ and $KG(jw) = K_1K_2\,/\underline{360^\circ}$. Thus $K_1 = \frac{1}{K_2}$.

In general terms this describes mathematically the requirements for any oscillator. Many different schemes for

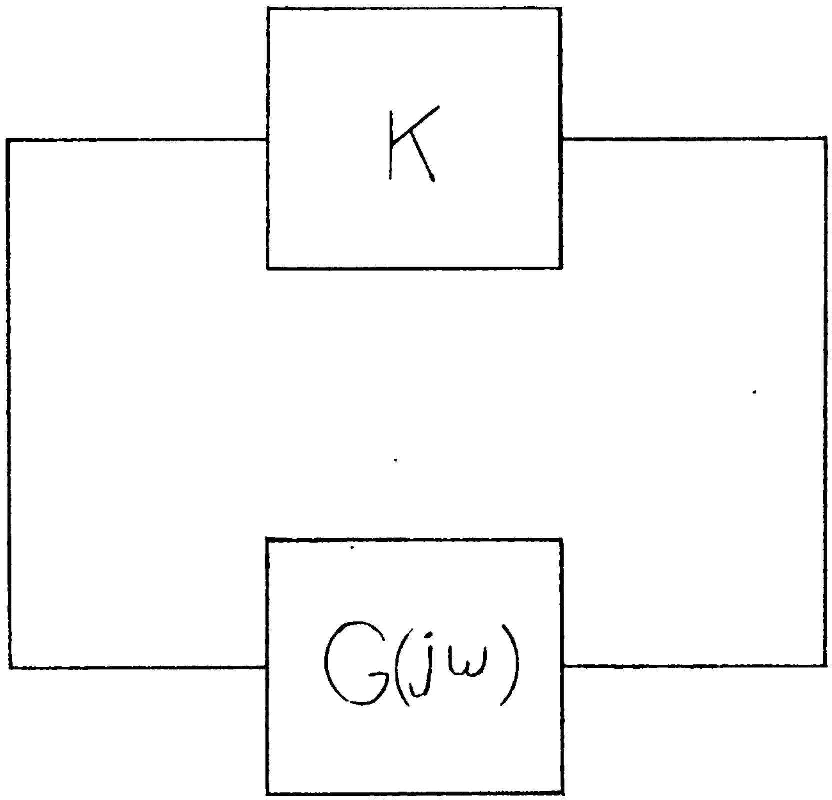

Fig. 2.I. Simple block diagram of an oscillator.

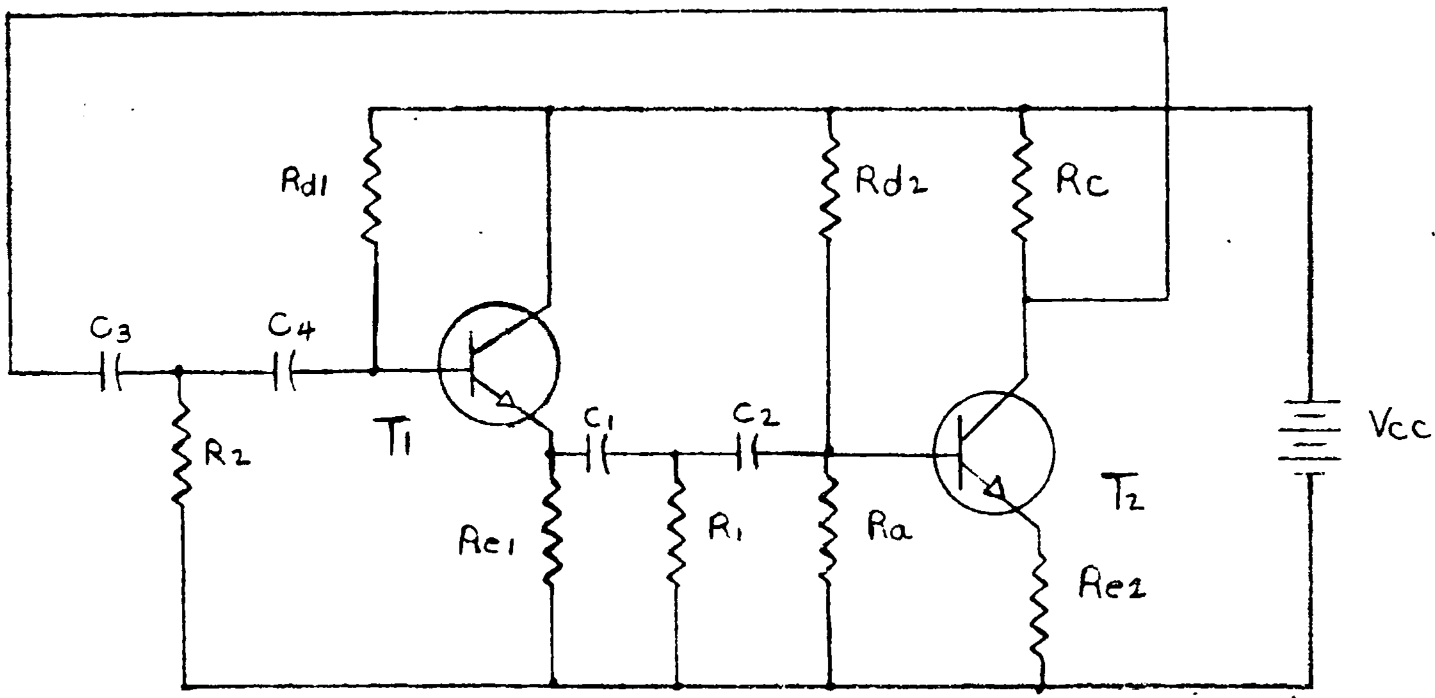

Fig. 2.2. Schematic diagram of transistor shunt R phase shift oscillator.

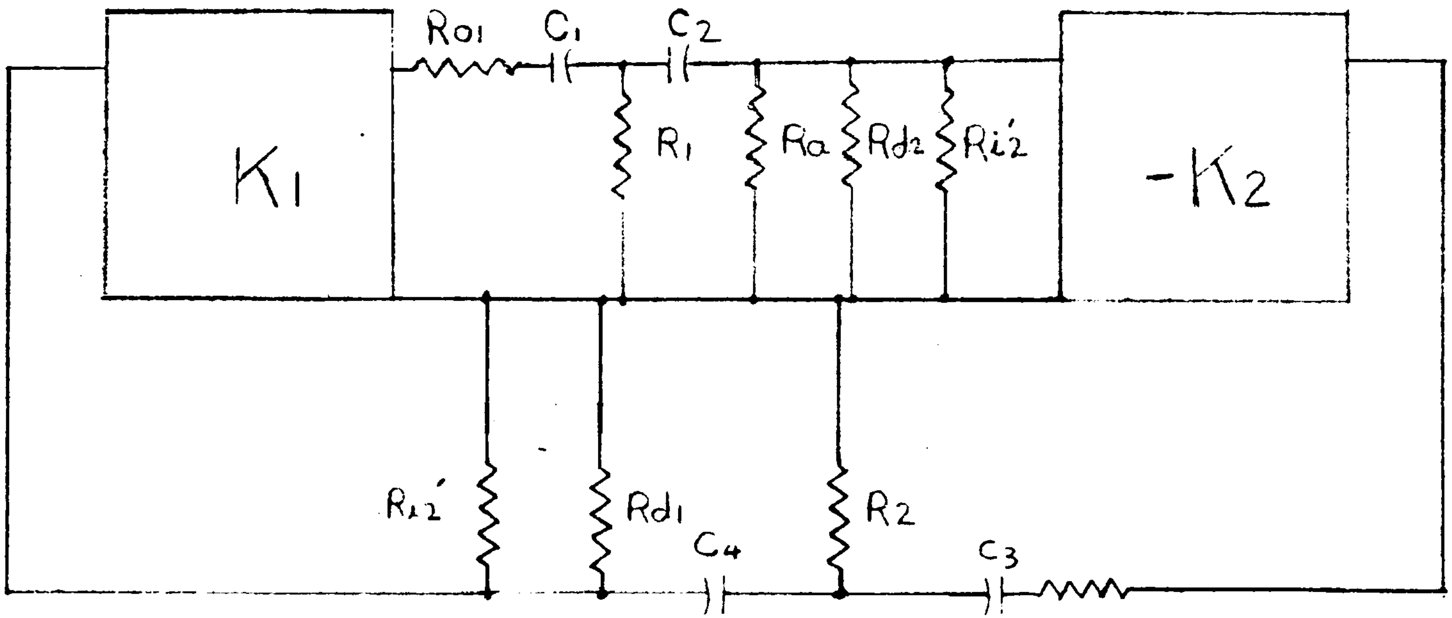

Fig. 2.3. The general equivalent circuit of the shunt R phase shift oscillator.

obtaining the required phase are used. In the oscillators with which this paper is concerned simple RC networks are used.

(2.3) THE SHUNT R OSCILLATOR.

The shunt R oscillator designed as part of this investigation is shown in figure (2.2). An equivalent circuit for this oscillator can be drawn as in figure (2.3), wherein R_{i2}' is the input impedance of T_2, R_{i1}' is the input impedance of T_1, R_{o1} is the output impedance of T_1, R_{o2} is the output impedance of T_2. K_1 is the gain of T_1 and $-K_2$ is the gain of T_2.

The circuit can be further simplified by letting R_{i2} represent the parallel combination of R_a, R_{d2}, and R_{i2}' and by letting R_{i1} be the parallel impedance of R_{d1} and R_{i1}'. Then the equivalent circuit reduces to figure (2.4).

This circuit can be further generalized and represented by figure (2.5). The function $G_1(jw)$ is the transfer function of the circuit of figure (2.6) and $G_2(jw)$ is the transfer function of the circuit in figure (2.7).

In Section (2.2) the basic requirements for an oscillator were written. Applying this criteria to the circuit shown in figure (2.5), $K_1K_2G_1(jw)G_2(jw)$ must equal one at an angle of 360°. Thus the magnitude of K_1K_2 must equal the reciprocal of $G_1(jw)G_2(jw)$. Further if K_2 equals $|K_2| \angle 180^\circ$ then

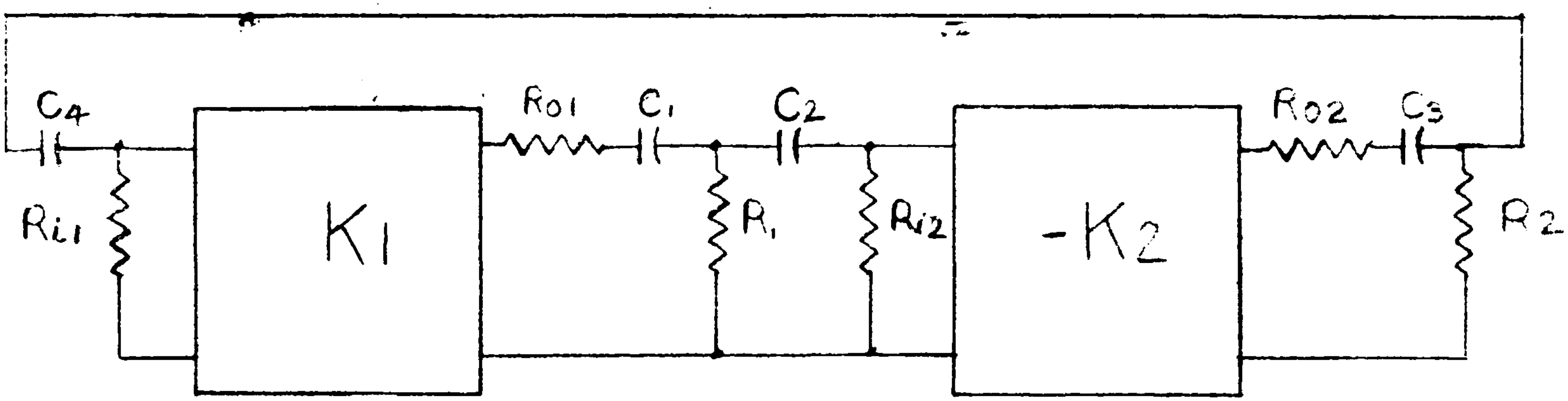

Fig. 2.4. Simplified equivalent circuit of the shunt R phase shift oscillator.

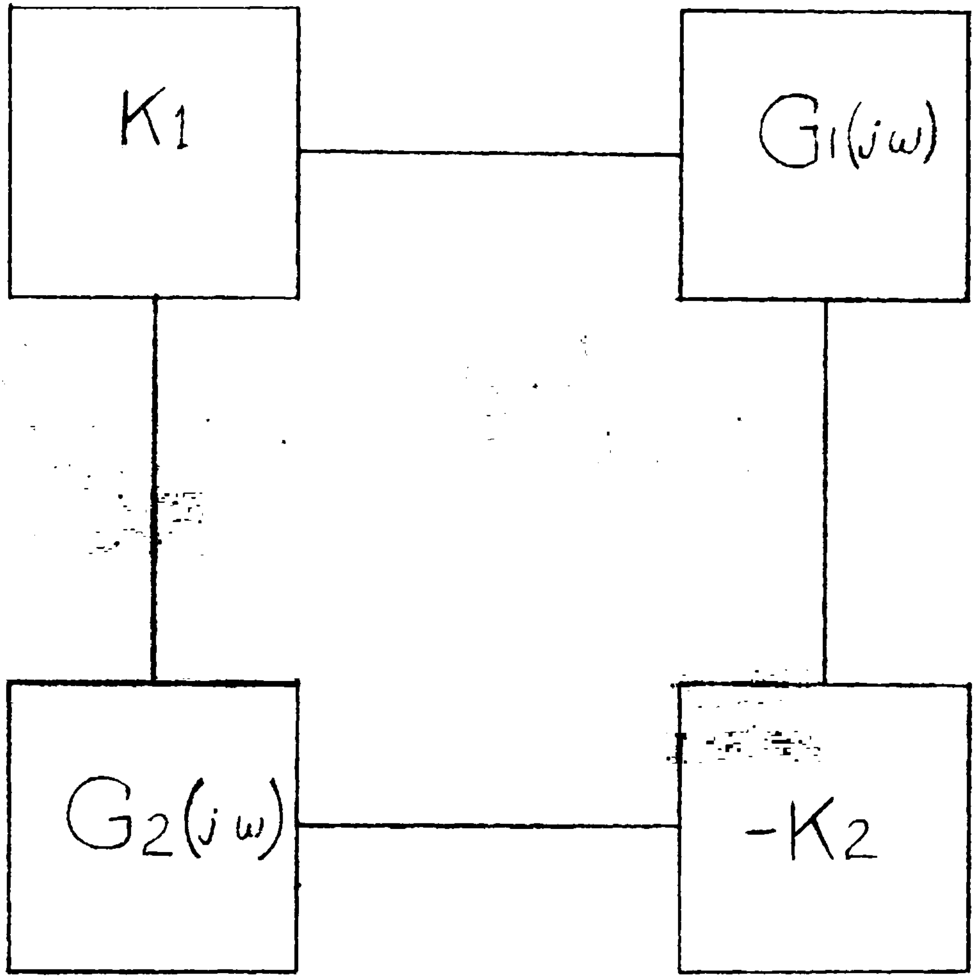

Fig. 2.5. Block diagram of the phase shift oscillator.

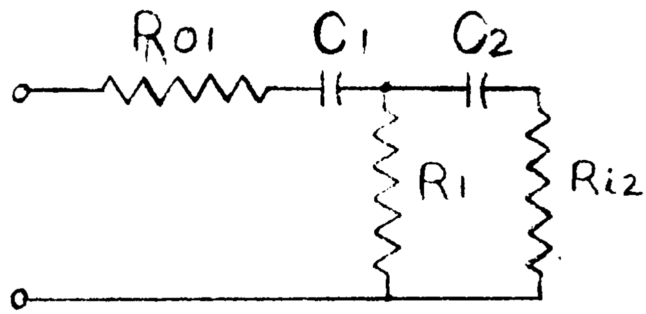

Fig. 2.6. That portion of the phase shift circuit represented by $G_1(jw)$.

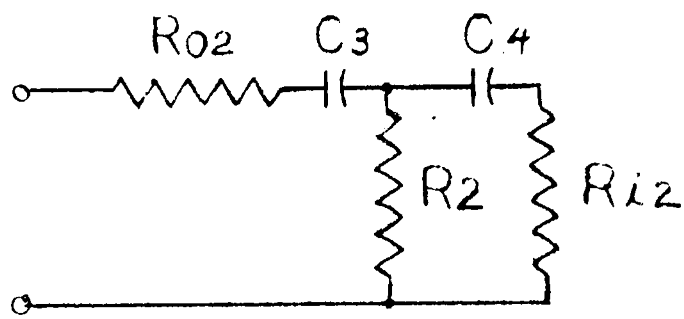

Fig. 2.7. That portion of the phase shift circuit represented by $G_2(jw)$.

$$G_1 G_2 = \left| \frac{1}{K_1 K_2} \right| \angle 180^\circ .$$

To solve the networks G_1 and G_2 including the transistor input and output impedances in general terms results in a long and (as a consequence) almost useless expression. The various transistor parameters appear many times in the expression. Since the parameters themselves must be measured if any degree of accuracy is to be attained, it is just as easy and probably more accurate to measure the input and output impedances of the transistor directly. This saves calculating the impedances from the transistor parameters.

In order to simplify the final equations of frequency and gain the assumptions given below were made.

$$C_1 = C_2, \quad C_3 = C_4, \quad R_2 = aR_1, \quad C_1 = aC_3 .$$

Then

$$R_1 C_1 = R_2 C_3 .$$

These assumptions make the equations less general but they do not impose severe limitations on the equations as far as their usefulness is concerned. In Chapter four these assumptions will be added to the design procedure.

Also, let

$$R_{12} = m R_1,$$

$$R_{01} = n R_1,$$

$$R_{i1} = h R_3,$$

and

$$R_{02} = k R_3 .$$

Then solving the expression $G_1(jw)G_2(jw)$ and setting the imaginary terms equal to zero an expression for w can be found.

$$\omega^2 = \frac{4 + h + k + m + n}{R_1^2 C_1^2 \left[(2+h+k)(m+n+mn) + (2+m+n)(h+k+hk)\right]} \quad (2.1)$$

In order to write the equation in more compact form let:

$$A = h + k + hk, \quad B = m + n + mn,$$
$$D = 2 + h + k, \quad F = 2 + m + n.$$

Then,

$$\omega^2 = \frac{D + F}{R_1^2 C_1^2 (DB + FA)} . \quad (2.2)$$

Substituting equation (2.2) into the expression for $G_1(jw)G_2(jw)$ equation (2.3) results.

$$\left| G_1(j\omega) G_2(j\omega) \right| =$$

$$\frac{hm(D+F)^2}{2AB(D^2+3DF+F^2) + BD^3F + AD^2F^2 + BD^3F^2 + ADF^3 + A^2DF + 2A^2F^2 + 2B^2D^2 + B^2DF} \quad (2.3)$$

Even though several simplifying assumptions have been made, equations (2.2) and (2.3) are still both cumbersome to use. In order to affect equations which would give a little more insight into the design problem the assumptions listed below will be made.

$$R_{\lambda 2} = R_1, \quad R_{\lambda 1} = R_2,$$

$$R_{o1} = 0.1 R_1, \quad R_{o2} = 0.1 R_2$$

(These assumptions are made valid by the design procedure). When these values are substituted into equations (2.2) and (2.3) equations (2.4) and (2.6 are obtained.

$$W^2 = \frac{1}{1.2 \, R_1^2 \, C_1^2},$$

$$f = \frac{1}{2\pi \sqrt{1.2} \, R_1 C_1}, \tag{2.4}$$

and

$$G_1 G_2 = \frac{1}{21}, \tag{2.6}$$

Although equations (2.4) and (2.6) have a number of assumptions restricting them they are reasonable equations for obtaining results that closely approach the desired result.

(2.4) THE SHUNT C OSCILLATOR.

The shunt C oscillator equivalent circuit can be drawn in a fashion similar to that of the shunt R oscillator. The equivalent circuit used in this analysis is shown in figure

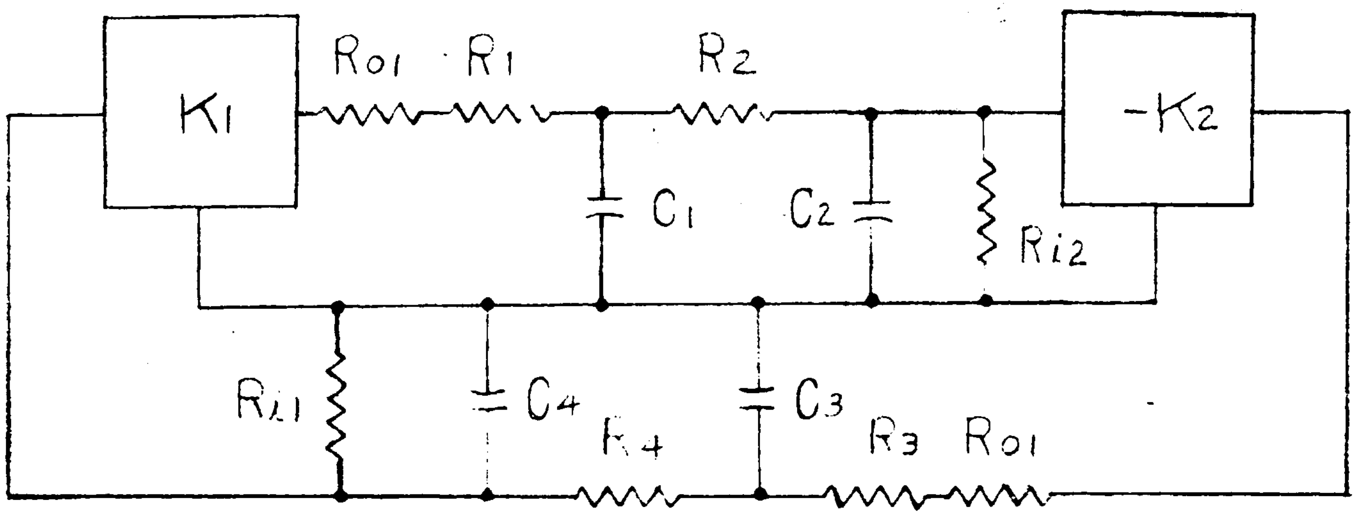

Fig. 2.8. Equivalent circuit of shunt C phase shift oscillator.

(2.8). Again, this circuit can be analyzed using the general equivalent circuit of figure (2.5). In this case $G_1(jw)$ is the transfer function of the circuit shown in figure (2.9). $G_2(jw)$ is the transfer function of the circuit of figure (2.10). In order to not overcomplicate the calculations and because it does not necessarily lessen the value of the calculations, the assumptions given below will be made.

$$C_1 = C_2, \quad C_3 = C_4,$$

$$C_1 = aC_3, \quad R_1 = R_2$$

$$R_3 = R_4, \text{ and}$$

$$R_1 = R_3/a.$$

Then by again solving for $G_1(jw)G_2(jw)$ and setting the imaginary terms equal to zero a solution for w can be found.

$$\omega^2 = \frac{BD_1 + B_1 D}{R_1^2 C_1^2 (B_1 A + BA_1)}, \qquad (2.7)$$

where;

$$B = \frac{1 + k + 3h + 2hk}{h},$$

$$A = 1 + k,$$

$$D = \frac{2 + h + k}{h},$$

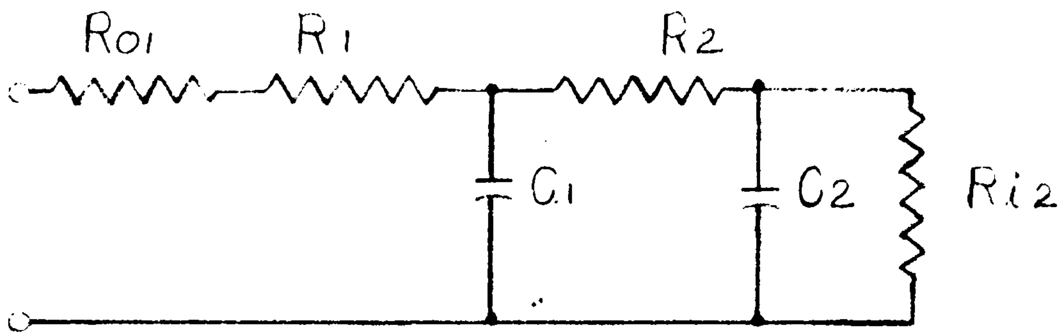

Fig. 2.9. The portion of the phase shift circuit represented by $G_1(jw)$.

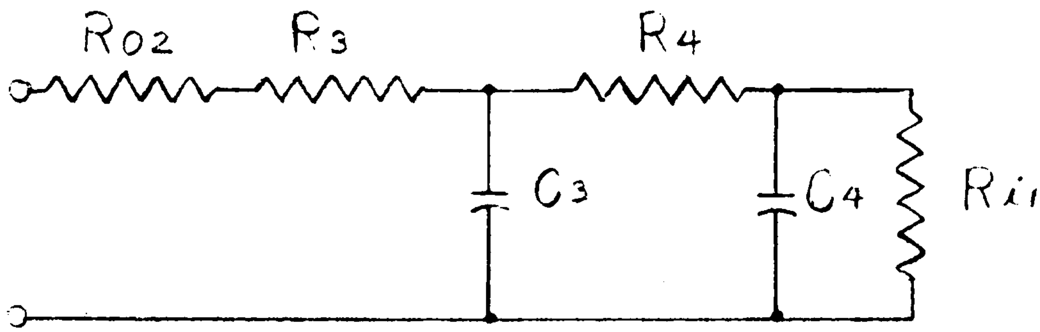

Fig. 2.10. The portion of the phase shift circuit represented by $G_2(jw)$.

$$D_1 = \frac{2 + m + n}{m},$$

$$A_1 = 1 + n,$$

$$B_1 = \frac{1 + n + 3m + 2mn}{m}$$

$$m = \frac{R_{i2}}{R_1}, \quad n = \frac{R_{o1}}{R_1},$$

$$h = \frac{R_{i1}}{R_3}, \quad k = \frac{R_{o2}}{R_3}$$

Substituting equation (2.7) into the general expression for G_1G_2 the attenuation of the two networks is found.

$$|G_1G_2| = \frac{1}{\dfrac{A_1A(BD_1 + B_1D)^2}{(AB_1 + A_1B)^2} - \dfrac{BB_1(BD_1 + B_1D)}{AB_1 + A_1B} + DD_1} \qquad (2.8)$$

If the assumptions are made that were made in section (2.4) for the shunt R oscillator, then $m \approx h \approx 1$ and $n = k = 0.1$.

When these substitutions are made in equations (2.7) and (2.8), equations (2.9) and (2.10) are obtained.

$$\omega^2 = \frac{2.1}{1.1\, R_1^2 C_1^2},$$

thus,
$$f = \frac{1.68}{2\pi R_1 C_1}, \qquad (2.9)$$

and

$$G_1 G_2 = \frac{1}{26.5} \qquad (2.10)$$

(2.5) SUMMARY.

In this chapter the calculations for finding the frequency and required gain of the two types of oscillators have been made. In the following chapter these results are compared with test results obtained in the laboratory. From these results a design procedure is formulated in Chapter 4.

A summary of this chapter and an outline of the design procedure is presented in Chapter 5.

CHAPTER 3

TEST DATA AND RESULTS

(3.1) INTRODUCTION.

In this chapter the equations derived in Chapter two are compared to test data taken in the laboratory. The discrepancies between calculated and test data are discussed. Possible sources of error are presented.

(3.2) SHUNT R OSCILLATOR TESTS.

In Chapter two the equations for the gain and frequency of the oscillators were derived. Equation (2.4) is the equation for the frequency of the shunt R oscillator under the stated conditions. Those were that the output impedance of the two transistors were equal to one-tenth of the shunt resistors and the input impedance was equal to the resistors. More simply

$$n = k = 0.1$$

and

$$m = h = 1.0.$$

In order to test the validity of equation (2.4) an oscillator of the type shown in figure (1.1) was carefully constructed using all precision* components. The input and

* one per cent tolerance.

output impedances of the transistors were measured. The input impedance of T_1 was measured and plotted as a function of R_{d1}. The input impedance of T_2 was measured and plotted as a function of R_{e2} and R_a. These plots are shown in figures (3.1) and (3.2).

The output impedances were assumed to be constant.

From the curves of input impedance the values of R_{d1} and R_{e2} required to give the desired input impedance could be selected. R_{o1} was found to be approximately 1000 ohms. R_{o2} was very nearly equal to R_c which in this case was 10,000 ohms. Thus in order to use equation (2.4) R_1 was made 10,000 ohms and R_2 was 100,000 ohms.

From figures (3.1) and (3.2) R_{d1} was selected as 200,000 ohms and R_{e2} was chosen as 350 ohms. These corresponded to input impedances of 100,000 ohms and 10,000 ohms respectively. It was found that when the oscillator was connected R_{e2} had to be increased to keep the transistor out of saturation. This changed the input impedance of T_2 to 11,000 ohms.

Using these values the calculated frequency was 1420 cycles per second. The test oscillator frequency was measured as 1320 cps. This amounts to an error of seven per cen based on the calculated frequency. For a carefully controlled test this does not seem exceptionally close. Several possible explanations for the discrepancy can be given. The best explanation seems to be that the equivalent circuit used to derive the equations is strictly a low frequency equivalent circuit of the transistor. None of the transistor capacitances were

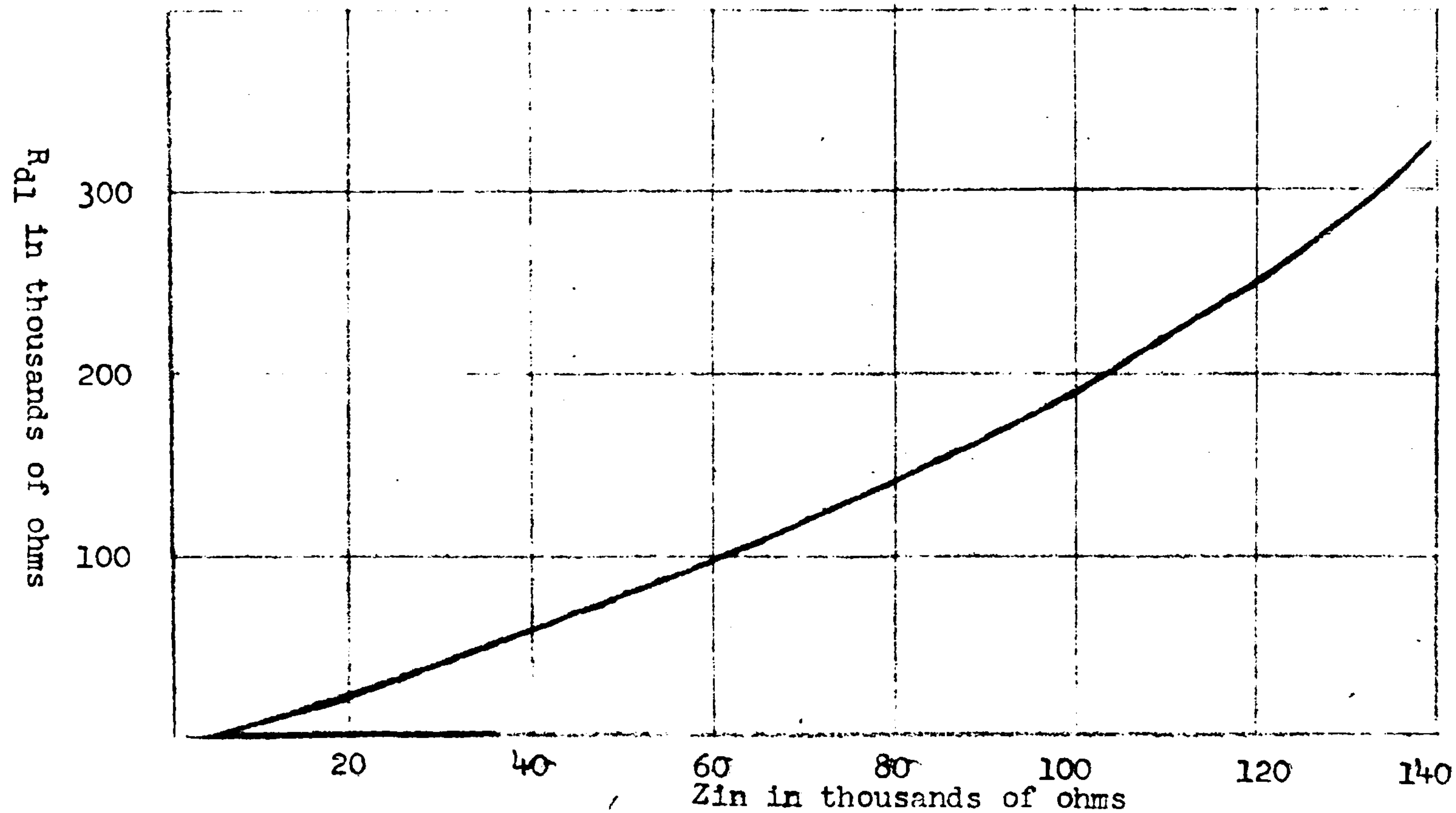

Fig. 3.1. Curve of input impedance as a function of R_{dl}.

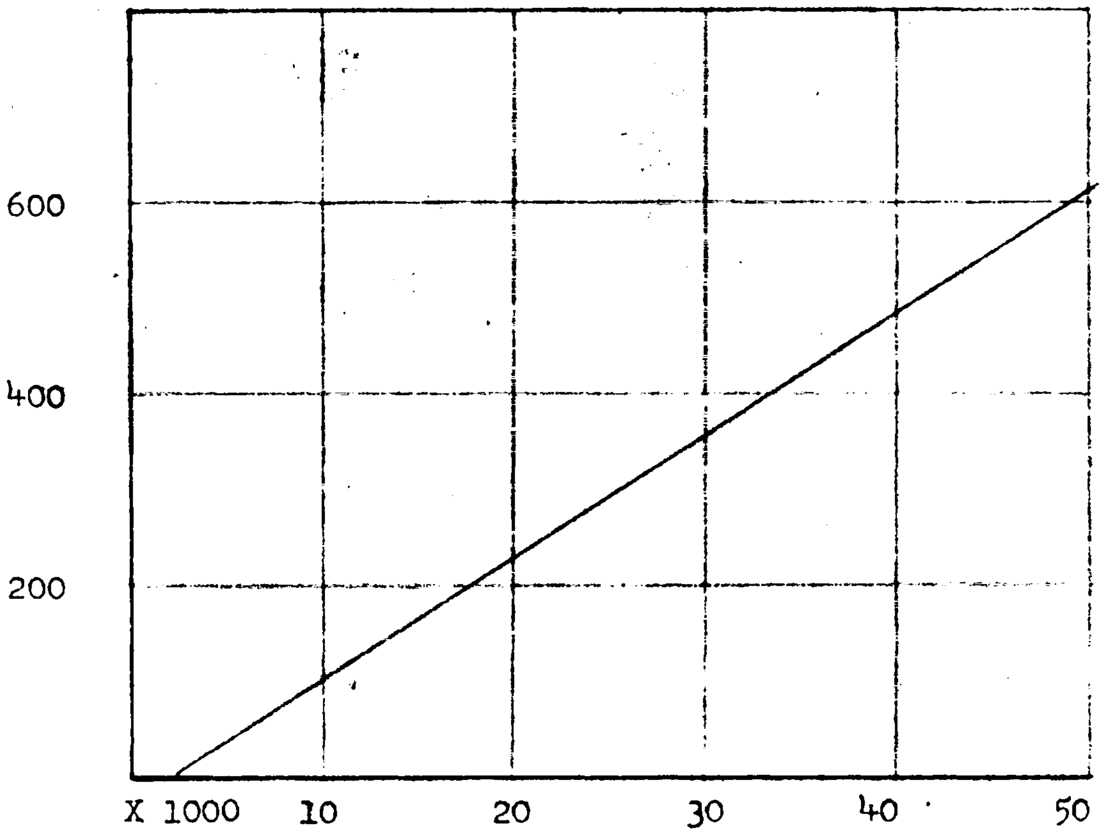

Fig. 3.2. Input impedance of the grounded emitter amplifier as a function of emitter resistance.

included in the equivalent circuit. It has been shown by test that the transistor input equivalent circuit can be represented by figure (3.3)[5]. Typical input capacitances have been measured in this circuit and found to be of the order of 1000 micro-micro-farads or more.[6]

Calculation will show that this network will reduce the frequency of the oscillator. This effect would be more pronounced in the amplifier stage than in the emitter follower. No laboratory work was done on determining the effect of these capacitances on oscillator frequency except to measure the input impedance at three different frequencies. It was found that for an R_{e2} of seven hundred ohms the input impedance of the grounded emitter amplifier was 57,000 ohms at 1,000 cycles per second, 53,000 ohms at 10,000 cycles per second and 10,000 ohms at 100,000 cycles per second.

Another assumption was made that the input impedance was not a function of h_{12}, the feedback parameter of the transistor. This assumption should not introduce any sizeable error since h_{12}, is a small quantity and only affects the input impedance by a small amount.

Further tests were run on this oscillator to determine the validity of the general equation for frequency of oscillation

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[5]T. L. Martin and Associates, Transistor Equivalent Circuit Criteria, University of Arizona, page 28 and 29.

[6]Ibid, page 30-31.

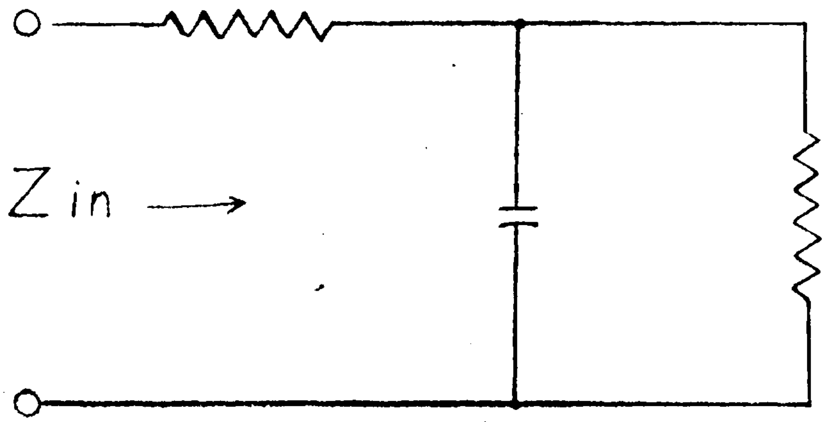

Fig. 3.3. An equivalent circuit of the input impedance of a transistor.

and the results are tabulated in table (3.1)

In this series of tests resistor and capacitor substitution boxes were used in conjunction with a terminal box for mounting the transistor. Precision components were not used. The significant part of this series of tests is that the calculated frequency was always between seven and nine percent higher than the measured frequency except in the high frequency case. The error then rises rapidly. Test #6 shows an error of nearly 30%.

(3.3) DISTORTION.

An important consideration for any sine wave oscillator is how much distortion is contained in the wave shape. With this particular oscillator the distortion can be controlled by providing adjustable current feedback. This controls the gain of the amplifier and by the same token controls the distortion. The oscillator wired to check the design equations had less than one per cent distortion as measured by a distortion analyzer.

(3.4) FREQUENCY DRIFT.

Although no heat chamber tests were run on the oscillator a cursory examination of its drift characteristics was made. The oscillator tended to drift only slightly with temperature change. The transistors were heated from room temperature to an estimated 130°F. The frequency change was too slight to

C_1	C_2	R_1	R_2	R_{d1}	R_{e2}	f(calc.)	f(meas.)
.01	.001	10k	100k	68k	470	1220 cps	1130 cps
.02	.002	10k	100k	68k	470	610 cps	560 cps
.03	.003	10k	100k	68k	470	407 cps	370 cps
.08	.008	10k	100k	68k	470	150 cps	140 cps
.14	.014	10k	100k	68k	470	87 cps	80 cps
.001	.0001	10k	100k	68k	100	15.5kcps	11 kcps
.5	.1	93k	100k	100k	470		9 cps

TABLE I. Test results of the shunt R oscillator.

measure with another oscillator used to form a Lissajous pattern on a cathode ray oscilloscope. Long term drift was not thoroughly investigated. However, the oscillator was left operating for six hours and the change in frequency was so slight that a nearly steady Lissajous figure was maintained without readjustment of the comparison oscillator. If drift were an extreme problem in a specific application of the oscillator a more detailed study would be warranted.

(3.5) SUMMARY OF THE SHUNT R OSCILLATOR TESTS.

The shunt R transistor phase shift oscillator provides a stable, distortion-free sinewave. It requires only nominal size components for low frequencies. Test results indicate that a range of frequencies from about one to thirty thousand cycles is available. However, the circuit is not well adapted for the higher frequencies for several reasons. For one thing the transistor input capacity begins to have a serious effect on the frequency. As a result it is extremely difficult to determine the circuit parameters for a specific frequency. The oscillator seems to be best suited for frequencies up to ten thousand cycles. At these frequencies the circuit parameters can be calculated with a reasonable degree of accuracy, probably within about 10%.

A complete design procedure for this oscillator and for the shunt C oscillator will be given in chapter four.

(3.6) THE SHUNT C OSCILLATOR.

It is possible to obtain higher frequencies with a shunt C phase shift oscillator than with the shunt R type. The input capacitance of the transistor has less effect than with the shunt R because the phase shift capacitors parallel the capacity and thereby lessen its effect.

In order to check the validity of equation (2.10) a shunt C oscillator was wired so that the conditions listed below existed. These are the conditions of equation (2.10).

$$m = h = 1, \quad n = k = 0.1,$$
$$C_1 = C_2 = aC_4 = aC_3.$$
$$R_3 = R_4 = aR_1 = aR_2$$

From equation (2.10)

$$f = \frac{0.267}{R_1 C_1}$$

In this instance

$$R_1 = 10{,}000\,\Omega, \quad R_3 = 100{,}000\,\Omega,$$
$$C_1 = 1000\,\mu\mu f, \quad C_3 = 100\,\mu\mu f.$$

Then

$$f_{calculated} = 26.7\ \text{Kcps}.$$

The measured frequency of the oscillator was 24.5 KCPS. There is an 8% difference in the two values. This discrepancy can probably again be attributed to the input capacitance of the transistor. This input capacitance will tend to reduce the frequency of the oscillator, although to a lesser degree than in the shunt R oscillator. In the shunt R oscillator to obtain 20,000 CPs the components used resulted in a calculated frequency of several times this value.

Further tests run on this oscillator are tabulated in table (II).

This data shows that it was possible to obtain by experimentation good waveforms up to 500,000 cycles. The equations were not tested in this area because of the large areas of uncertainty concerning transistor capacitances.

(3.7) DISTORTION.

The distortion of this oscillator was extremely low. On the test oscillator at 24,500 cycles no distortion was observeable. The distortion was not measured on a distortion analyzer. In general this oscillator as well as the shunt R oscillator can be made to be practically distortion free by proper gain adjustment.

(3.8) FREQUENCY DRIFT.

The drift characteristics of the shunt C oscillator seemed to be much the same as the shunt R oscillator. It was very stable with large temperature changes. No noticeable frequency change occurred in four hours of continuous operation.

(3.9) SUMMARY.

The two transistor oscillators gave superior results when

C_1	C_2	R_1	R_2	R_{d1}	R_{e2}	f(measured)
.002	.0001	190	10k	15k	220	170 kcps
.001	.0001	150	10k	68k	220	225 kcps
.001	.0001	140	1.5k	15k	330	365 kcps
.001	.0001	100	470	15k	220	500 kcps

TABLE II. Test results on the shunt C oscillator. These tests were to determine upper frequency limit of the oscillator.

compared to the single transistor oscillators described in Chapter two. When these single transistor oscillators were constructed in the laboratory it was found that they both experienced bad frequency drift with temperature changes. It was possible to change the frequency noticeably by merely holding one's fingers on the transistor for a short time. The frequencies obtainable out of them seemed to be limited.

In tests run by this writer while exploring various circuits the lowest obtainable frequency was about 500 cycles per second. The highest frequency attained was 3,500 cycles per second. These tests were not extensive enough to establish that these are the frequency limits. They did indicate that the range of these single transistor oscillators was limited.

In using two transistors as has been done in this investigation several real advantages seem to have been effected. They are:

(1) Wide frequency ranges are available using reasonably small components.

(2) High transistor voltage and current gains are not required, resulting in better frequency stability and gain stability.

(3) Transistor parameters, as such, need not be measured and are not critical.

(4) Reliability is increased because both transistors can be operated well below their maximum dissipation rating.

The most undesirable feature of this oscillator is the complexity of the design equations. By making the assumptions given in this chapter, simple equations evolve. Test results indicate errors of less than ten per cent in frequency calculations. This is of the same magnitude as is found from the single transistor oscillator equations.

Little has been said about the gain in this chapter. The reason for this is that no matter how careful gain calculations are made it is almost always necessary to make a slight gain adjustment for good wave shapes. A variable resistor should always be used for the emitter resistor of the amplifier stage for this reason.

CHAPTER 4

DESIGN PROCEDURE

(4.1) INTRODUCTION.

In the development of the oscillator circuits and the resulting equations written to describe them, it was found necessary to make assumptions in order to simplify the design problem. The transistor measurement problem was eased by using the techniques outlined in section (3.3). In order to accomplish one of the objectives of this thesis these and other ideas must be incorporated into a design procedure. The purpose of this chapter is to develop a useable design procedure.

(4.2) CHOOSE THE TRANSISTORS.

The choice of the transistor to use can be quite arbitrary. At low frequencies any transistor which can provide sufficient gain can be used. The higher the current gain the better, however, because the current gain determines to a large degree the amount of current feedback that can be used. The current feedback reduces transistor drift.

High current gain produces high input impedance which may be desirable, especially at low frequencies.

The transistor "alpha cutoff" frequency will be important if a higher frequency oscillator is desired. In general the higher the transistor cutoff frequency the higher the oscillator frequency can be made.

Other than these considerations the only others that seem important are the general ones concerning voltage, current and power ratings.

(4.3) DESIGN AND CONSTRUCT THE AMPLIFIERS.

The design of the amplifiers consists, at this point, of selecting the values of R_{e1}, R_{c2}, and R_{d3} and the voltage source. The remaining components affect frequency and are chosen from other considerations. These values are chosen using the common transistor considerations of quiescent point, liniarity and dissipation. Construct the amplifiers.

(4.4) MEASURE TRANSISTOR IMPEDANCES.

After constructing the amplifier the next step is to measure and plot the input impedance of the emitter follower as a function of the bias resistor, R_{d1}. A test setup for doing this is shown in figure (4.1). In this test various values of R_{d1} are placed in the transistor circuit as shown. For each value of R_{d1}, R_x is adjusted until $V_2 = 1/2\ V_1$. Then R_x is equal to the input impedance. After several values of R_x as a function of R_{d1} have been measured then a plot of input impedance versus R_{d1} can be made as was done in figure (3.1).

The same technique can be applied to the grounded emitter

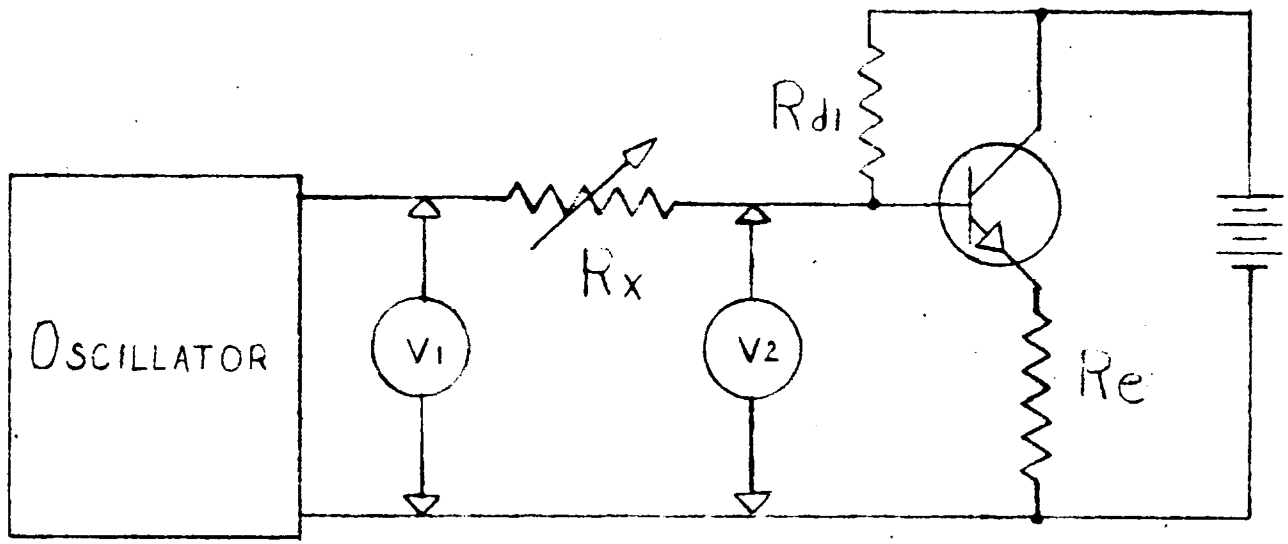

Fig. 4.1. Test setup for measuring input impedance to transistor amplifiers.

amplifier. The only difference is that R_{d2} is varied. When plotted a curve similar to figure (3.2) is obtained.

A single measurement is all that is necessary for each of the output impedances. In this case it is probably safe to assume that the output impedance of the amplifier stage is equal to the load resistor. The output impedance of the emitter follower can be measured or calculated. Its effect on the gain and frequency is small.

In making the latter measurements the test setup shown in figure (4.2) can easily be used. The open circuit voltage, V_{oc} at the output of the transistor is measured. Then the resistor load is added. The resistance, R_z, is decreased until the voltage is one half the open circuit voltage. Then R_z is equal to the output impedance. In general terms

$$R_{out} = \frac{V_{oc} - V_1}{V_1} R_z, \qquad (4.1)$$

where

V_{oc} is the open circuit voltage,

V_1 is the output voltage under load,

and R_x is the series resistance.

(4.5) DETERMINE RESISTANCES AND CAPACITANCES.

The resistance-capacitance product is determined by the desired frequency. In order to use the design equations derived in Chapter two the RC product of both sections must be

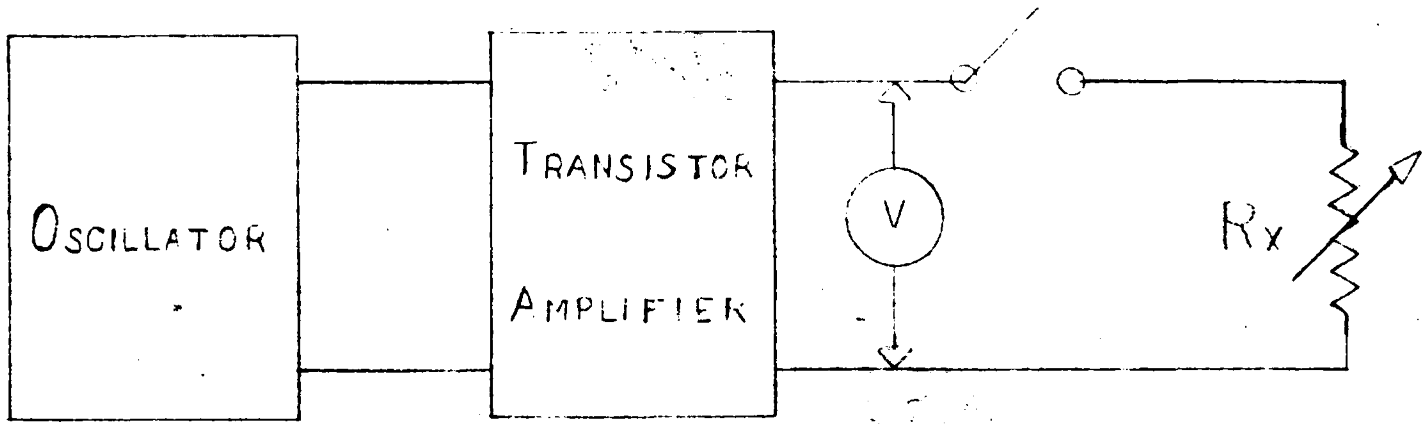

Fig. 4.2. Test Setup for measuring output impedance of the transistor amplifiers.

equal. This means:

AR_1 must equal R_2 and

C_1 must equal AC_2,

because R_1C_1 then equals R_2C_2.

Determine the value of R_1 and R_2 by letting R_1 be equal to ten times the output impedance of the emitter follower and by letting R_2 be equal to ten times the output impedance of the grounded emitter amplifier. In symbolic form

$$R_1 \cong 10R_{o1}, \text{ and}$$

$$R_2 \cong 10R_{o2}.$$

The values of C_1 and C_2 can be established by using the appropriate design equation with the value of R just calculated substituted into the frequency equation.

To find the value of R_{d1} reference to the plot of input impedance versus R_{d1} is all that is necessary. Let the input impedance be equal to R_2 and find the corresponding value of R_{d1} on the curve.

An exact determination of R_{e2} at this point is almost impossible. R_{e2} should be as large as possible while still maintaining sufficient gain to sustain oscillation. Therefore, it becomes expedient to choose R_a first. The approximate value of R_{e2} can be calculated from the gain requirement and from the approximate gain equation. It can be shown that

$$\text{Amplifier gain} \approx \frac{R_c}{R_e}. \tag{4.2}$$

Using these equations and finding a value for R_{e2} the input impedance of the amplifier corresponding to this value of R_{e2} can be found. Select R_a so that it in parallel with the input impedance of the grounded emitter amplifier results in a value equal to R_1.

R_{e2} should be made variable in order to provide gain control. This will allow adjusting the gain for the least distortion.

The selection of R_{d2} is made from the normal transistor operating point considerations. It will usually be very large as compared to the values of input impedance and can thus be neglected in frequency calculations.

(4.6) SUMMARY.

In this chapter a design procedure has been presented which gave good results in the laboratory. In outline form the procedure was:

(1) Choose the transistor.

(2) Design and construct the basic amplifier circuit.

(3) Measure output impedances.

(4) Measure and plot input impedances.

(5) Set $R_1 \cong 10\ R_{o1}$, and $R_2 \cong 10\ R_{o2}$.

(6) Calculate C_1 and C_2 from the design equations.

(7) Select R_{d1} so that $R_{i1} \cong R_2$.

(8) Set R_{e2} equal to R_c divided by the gain required.

(9) Calculate the value of R_a.

CHAPTER 5

SYNOPSIS

In chapter one the objects of this work were written. They were; to develop a useable design procedure and design equations which would allow the user to calculate the components of the two phase shift oscillators for a desired frequency. This has been done and the results are summarized in this chapter.

The transistor phase shift oscillators shown in figures (5.1) and (5.2) are good oscillators using the usual oscillator criteria as basis for judgment. They both exhibited good frequency stability, low distortion, and should be reliable.

Design equations show that for the shunt R oscillator, figure (5.1),

$$\omega^2 = \frac{D + F}{R_1^2 C_1^2 (DB + FA)}$$

when

$$R_1 C_1 = R_2 C_2,$$

D, F, B, and A are as defined in section (2.3). If the design procedure is followed

$$f = \frac{1}{2\pi \sqrt{1.2}\, R_1 C_1},$$

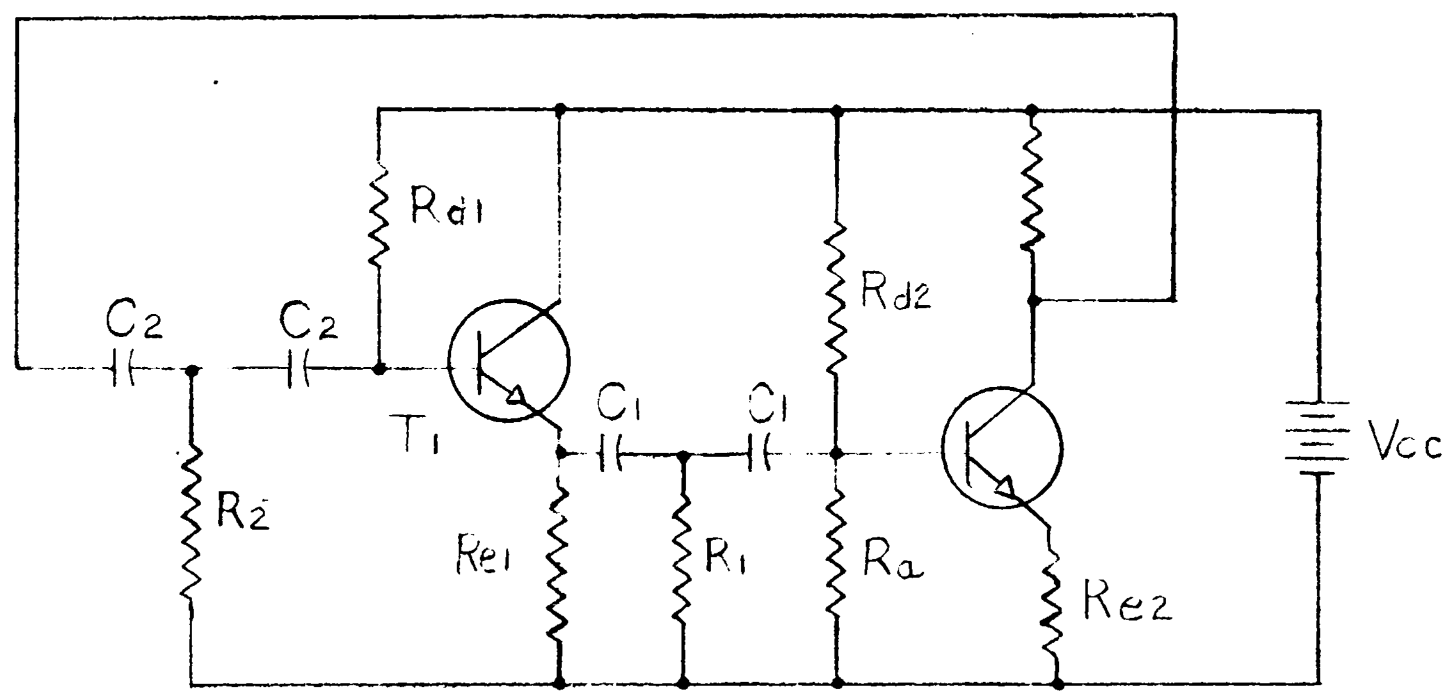

Fig. 5.1. The shunt R phase shift oscillator in the form described in the design procedure.

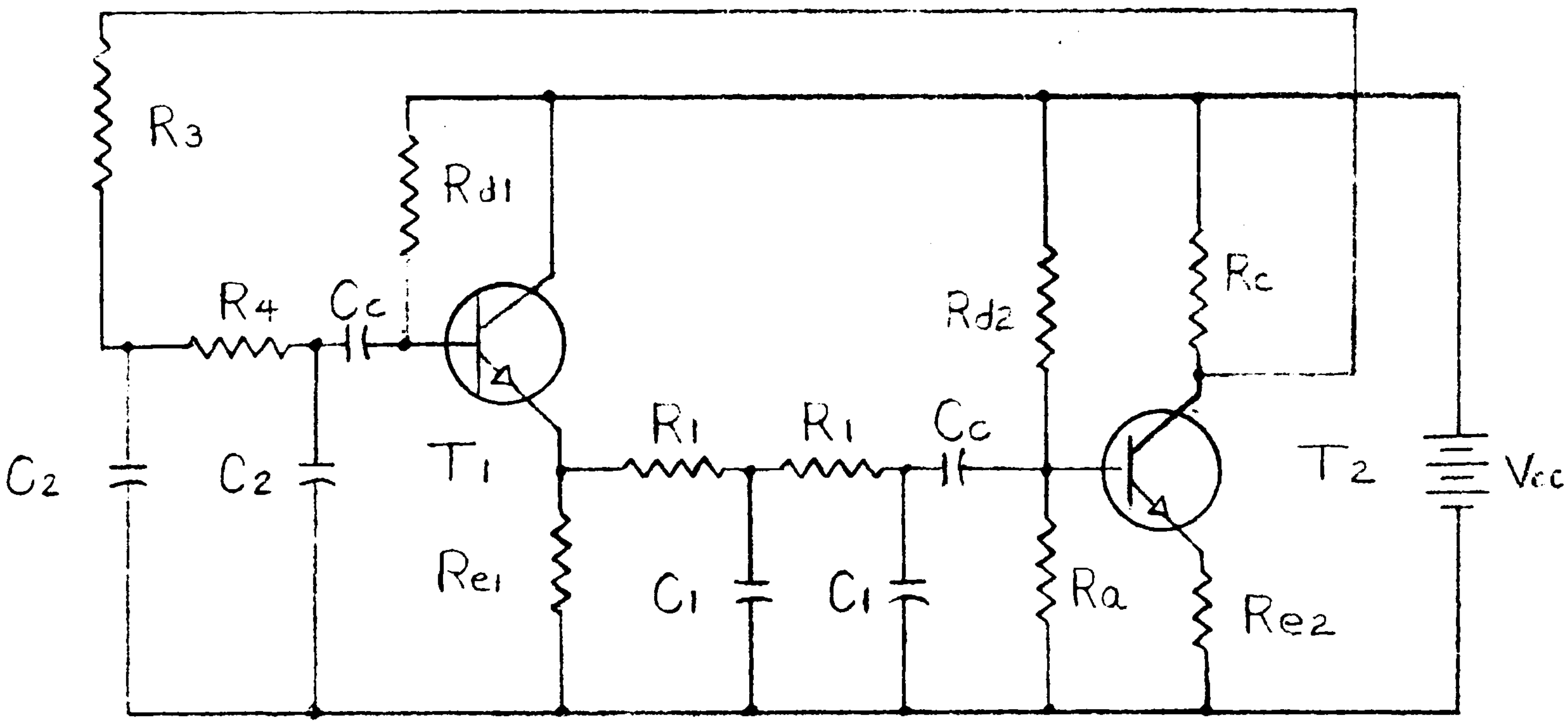

Fig. 5.2. The shuntCphase shift oscillator in the form described in the design procedure.

and

the gain required $\approx$ 21

For the shunt C oscillator

$$\omega^2 = \frac{BD_1 + B_1D}{R_1^2C_1^2(B_1A + BA_1)}$$

where A, A_1, B, B_1, C, C_1 are as defined in section (2.4) and $R_1C_1 = R_2C_2$.

If the design procedure is used:

$$f = \frac{1.68}{2\pi R_1C_1},$$

and the gain required = 26.5.

The design procedure is summarized below.

(1) Choose the transistor from gain and frequency requirement considerations.

(2) Design and construct the basic amplifier circuit.

(3) Measure output impedances.

(4) Measure and plot the input impedance of the grounded emitter amplifier as a function of the emitter resistor.

(5) Measure and plot the input impedance of the emitter follower as a function of R_{d1} the bias resistor.

(6) Set $R_1 = 10\ R_{o1}$, $R_2 = 10\ R_{o2}$.

(7) Calculate C_1 and C_2 from the design equations.

(8) Select R_{d1} so that $R_{i1} = R_2$.

(9) Set R_c/R_{e2} equal to the required gain and solve for R_{e2}. Find the corresponding value of input impedance of the grounded emitter amplifier.

(10) Calculate R_a by setting the parallel combination of the input impedance from step (9) and R_a equal to R_1.

(11) Make R_{e2} a variable resistor.

Using this design procedure the frequency of the oscillator can be expected to be not more than ten per cent different than the calculated value with some reservation. The shunt R oscillator should be used below 20,000 cycles per second. Above 10,000 cps the error in calculated frequency will rise above 10 per cent unless a transistor with low input capacitance is used.

The shunt C oscillator has its best application above 10,000 cps. Calculation error remains below 10 per cent up to 30,000 cps.

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