ABSTRACT

The general theory of the common connections of synchrocontrol elements is outlined under static and continuous rotation conditions. The harmonic response functions of such devices are analyzed using modulation theory and measurement techniques both direct and indirect presented. Discussion of the difficulties in experimental verification complete the dissertion.

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SYNCHRO CONTROL DEVICES

C O N T E N T S

C H A P T E R 4

FURTHER SYNCHRO PROPERTIES AND USES

C H A P T E R 5

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CHAPTER 10

THE HARMONIC RESPONSE FUNCTION OF THE BASIC SYNCHRO SYSTEM

P R E F A C E

Quite often in the study of automatic control systems or servomechanisms the student becomes so involved with the analytical aspects of system design that he tends to take the components necessary to transform the design ideas into actual working devices almost for granted and hence often losses sight of the limitations involved in the use of such components. It is also quite easy to simply accept "given" quantities in idealized text-book problems without realizing what methods and techniques are really necessary to measure such things as inertial loading constants, friction loading constants, and harmonic response functions in actual practice. Often it is only when some experience in working with real components has been gained does one begin to realize some of these points. It is the hope of the author that the choice of the topic for this dissertation will lead to some measure of this realization.

In many instances the synchro is looked upon as a device whose main purpose is for position indication, that is, where a synchro rotor at one location indicates the angular position of a second synchro rotor at some remote location. A simple example of this would be the employment of a synchro generator and a synchro receiver to indicate the position of an antenna. To look upon the synchro in this manner is to loose sight of its many and far more important uses in connection with automatic feedback control systems where the synchro is used to provide an electrical error signal proportional to some physical quantity such as the difference in angular position between two rotating shafts. The main emphasis of this thesis will be this application of synchros to automatic feedback control systems.

Part of the material contained herein represents a condensation and sorting out of the work done to date on synchros as found in most of the presently available references. The references used here are listed at the end of the appendix. In many cases expansion of the original article has been done to develop a specific item into a more general coverage. This for example has been done in connection with the analysis of the basic synchro pair operated with relative sinusoidal oscillation of the rotors. The original article dealt only with oscillation about one point, the correspondence point, and this has been expanded to cover the more general situation of oscillation about any point within the region of reasonably linear operation of the synchro pair. Some work not found in the references has also been included such as the analysis of the synchro pair with the relative displacement angle between the two rotors of the pair changing at a constant rate. This is listed in the table of contents under "Operation of the Synchro Pair With Continuous Rotation of the Generator Rotor".

Several possible uses of synchro devices beyond their normal applications are treated both in an experimental and analytical manner. The harmonic response function of the basic synchro system is dealt with at some length and in considerable detail. A new technique, the word new being employed strictly in a relative sense, for the determination of the harmonic response function is discussed and thoroughly outlined.

Even though it was possible to actually set up equipment and perform many actual tests using synchro devices, some aspects of synchro operation were not able to be checked experimentally due to limitations on equipment availability and construction facilities. This was the situation with respect to the errors inherent in synchro system operation. It would have been very informative to verify, using an actual synchro system, some of the analytically developed expressions for synchro errors but the cost of equipment necessary for such work was prohibitive.

In general the thesis covers almost all aspects of synchro operation and although some items are not [redacted] treated in as much detail as they could be this work should serve as a useful overall reference on synchro systems.

A C K N O W L E D G E M E N T S

The author is indebted to several persons for their invaluable assistance and advice during the preparation of this dissertation, in particular to Mr. R. Woods for his assistance and extreme patience with regard to the mechanical construction required throughout the process of this work, to Mr. T.J. White for obtaining necessary electronic components, to Miss D. Simpson of the Engineering library for her general assistance and more specifically for her efforts in obtaining several valuable articles from the General Electric Co. library in Schenectady, N.Y., to Miss L. Walker for her secretarial aid, and to his fellow graduate students for their interest and advice. The author also wishes to express his appreciation to his thesis examining committee, Mr. R.A. Johnson, Mr. J.P.C. McMath, Mr. W. Shepherd, of the Electrical Engineering Department of the University of Manitoba, and Mr. P. Shane of the Manitoba Power Commission, for their time and efforts in connection with the reading of this work, and to the National Research Council for the bursary which made this work possible. I am also very deeply indebted to Mr. R.A. Johnson not only for his assistance in the role of advisor for this thesis but more for the inspiration which he has given me during these last three years to seek new and higher goals. To my wife for her encouragement and many long hours of typing I am also sincerely grateful.

K. A. I.

CHAPTER 1

SYNCHRO SYSTEM COMPONENTS

Introduction

An important component of almost all control systems is the device which is used to convert input and output variables from their original physical form to a form that is more convenient to use. Devices of this type are usually referred to as transducers. In an electro-mechanical system such transducers are used to convert the position or velocity of a shaft output into electrical signals which are more easily handled in amplifiers and associated networks.

The primary requirement to be fulfilled by a transducer is accuracy. Since the transducer always functions outside of the control loop, its inaccuracy cannot be compensated for by means of feedback. A servo therefore cannot be more accurate than its transducer. It is normally desirable that transducers be linear, that is, that the output variable be directly proportional to the input variable. Any departure from linearity would give rise to repeatable error in a nominally linear transducer and thus reduce the accuracy. In many servomechanism applications, two transducers with closely matched characteristics are used, one at the input and one at the output. Any mismatch between these two units would provide another source of repeatable error and therefore adversely affect the accuracy of the system.

In addition to accuracy and precision, another desirable characteristic of transducers is high input impedance, that is, the transducer should require a minimum of power for its operation.

There are a number of devices which will perform this desired conversion between mechanical and electrical energies. Some of these are synchros, pickoff, linear transformers, differential transformers, and E transformers, but the synchro, or selsyn* as it is sometimes known, is the most widely used. The synchro usually resembles a small synchronous motor in construction with the actual design depending upon the operation the synchro is to perform in the system. The various synchro types are the generator or transmitter, the receiver, the differential generator or differential transmitter, the control transformer, and the differential receiver.

The electrical output from the synchro is an amplitude modulated wave with the carrier frequency being the same frequency as the a.c. supply to the generator rotor and the modulating frequency being directly related to the relative angular motion of the generator and control transformer rotors. It is the modulation that provides the useful output. In most applications the generator rotor only moves within a variance of a few degrees and the frequency of this modulating wave is quite low. Upper limits imposed on the modulating frequency are discussed in the section dealing with

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* General Electric Co. trade name.

discriminators. This modulated a.c. output signal is fed from the synchro control transformer into a unit known as a discriminator which extracts the useful portion of the wave. This signal is then fed to a servomechanism which makes use of the level and polarity of the signal to instigate a driving mechanism to return the original system to the normal operating condition.

The synchro also finds some use as a remote position indicator where the angular position of some device is transmitted and subsequently indicated on a display at a position which is removed from the actual location of the device.

In a study of synchros and their operation, it becomes fairly evident that the discriminator belongs more naturally with the synchro system than it does with the servomechanism or actual error-correcting driving devices so that the discriminator shall be included in this study of synchro systems. Thus the range to be covered in this dissertation will be from the mechanical input signal to the synchro generator to the output signal of the discriminator. It shall be assumed herein that servomechanisms exist which will accept the output from the discriminator and, depending on the polarity and magnitude of this output signal, instigate the desired effect to remove the error which the synchro system is detecting. These devices shall therefore not be dealt with in this study.

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Construction of a Synchro

The construction of a synchro is very similar to that of a miniature, three phase synchronous alternator. The stator is constructed with standard slotted steel punchings containing a three phase winding that is usually wye-connected. In order to minimize slot effects, the stator slots are often skewed. There are three different types of rotors in common use; the salient pole type, the umbrella type, and the cylindrical or drum type. All of these rotors are wound for two electrical poles, and the windings are brought out on slip rings so that continuous rotation is possible. Salient pole rotors are used for synchro generators and motors. Both of the other types of rotors are used in control transformers where it is desirable to have a uniform reluctance all the way around the air gap so as to keep any torque which may be produced by the rotor down to a minimum. Both the stator and rotor windings of control transformers usually have a higher impedance than those of the generator so that it is possible to excite several control transformers from one generator.

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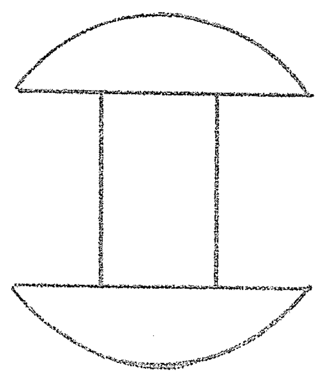
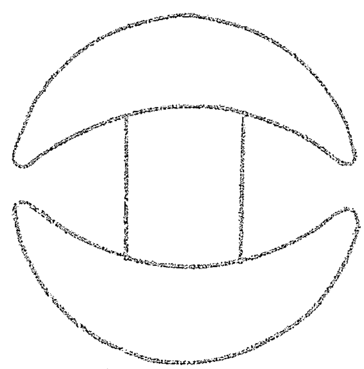
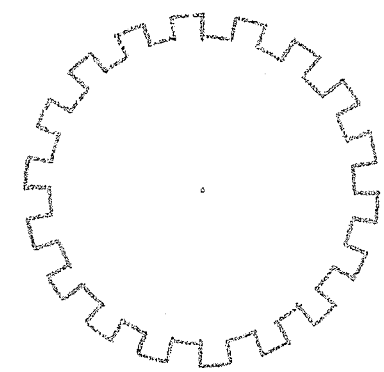


SALIENT POLE



UMBRELLA



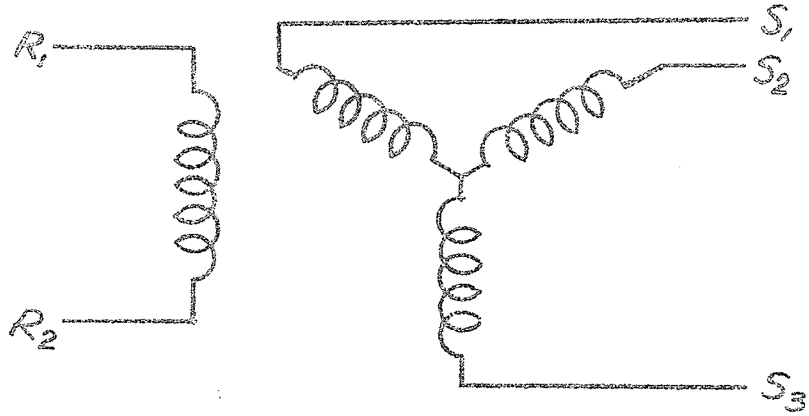
CYLINDRICAL

TYPICAL SYNCHRO ROTOR CROSS-SECTIONS

Fig. 1.

INDEX OF SYNCHRO TYPES

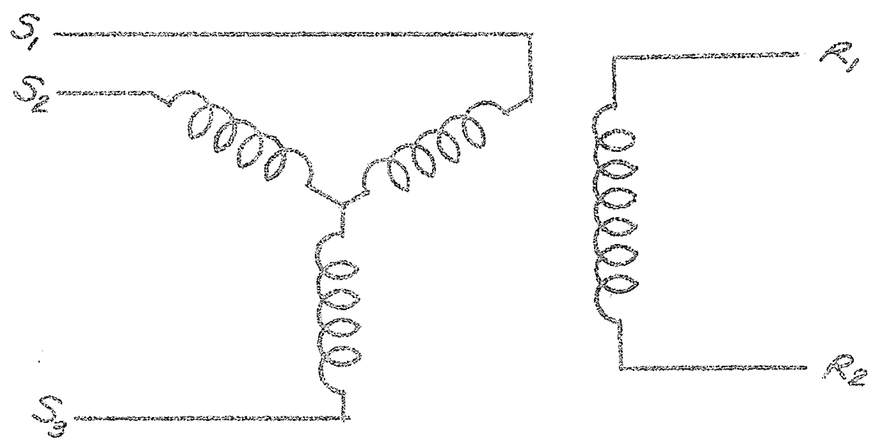
1. Generator or Transmitter.



Function - When its two input leads are excited by an alternating voltage, it applies to its three output leads , voltages whose relative magnitudes and polarities uniquely define the angular position of its shaft with respect to its stator.

Application - Servomechanisms, when used with a control transformer, or a remote indicating system , when used with a receiver.

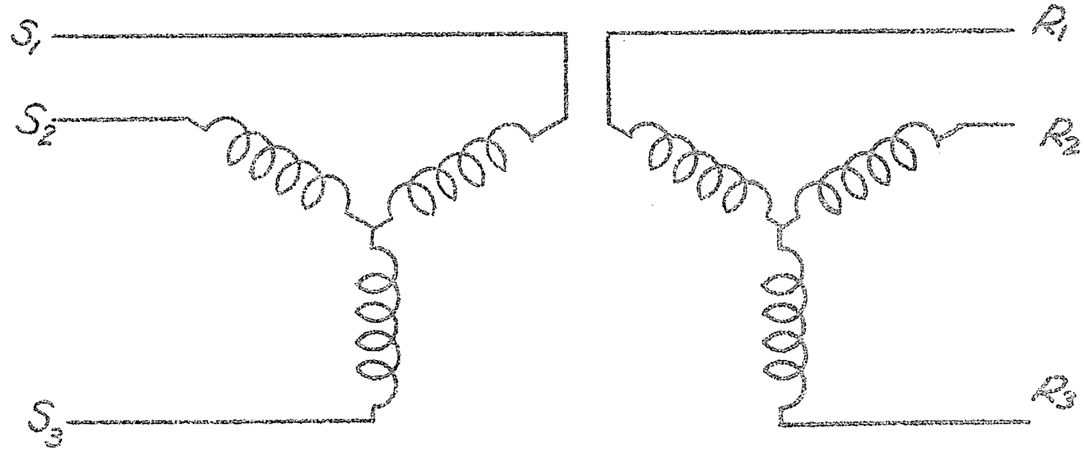
2. Control Transformer.



Function - When its three input leads are excited by voltages which define an angle, it applies to its two output leads a voltage which is proportional to the sine of the difference between this angle and the angular position of the shaft with respect to its stator.

Application - Servomechanisms, (used in conjunction with a generator).

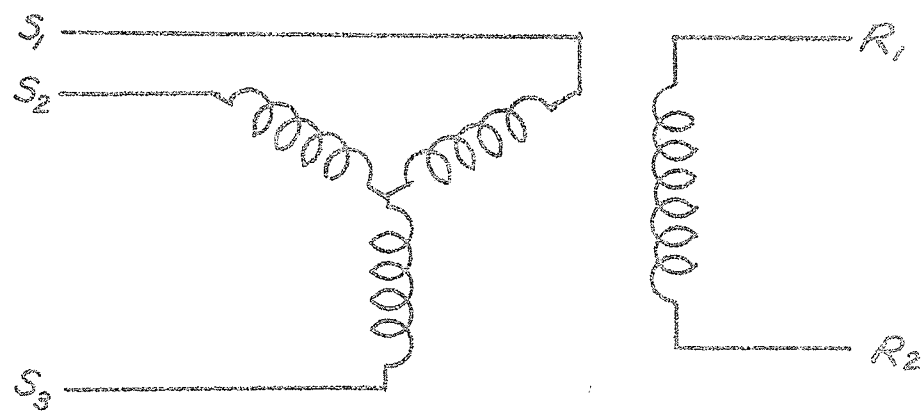
3. Differential Generator or Differential Transmitter:



Function - When its three leads are excited by voltages which define an angle, it applies to its three output leads, voltages whose relative magnitudes and polarities uniquely define a second angle which is the sum of the first angle and the angular position of the differential generator shaft with respect to its stator.

Application - Servomechanisms, when used with a control transformer, or remote indicating systems, when used with a receiver.

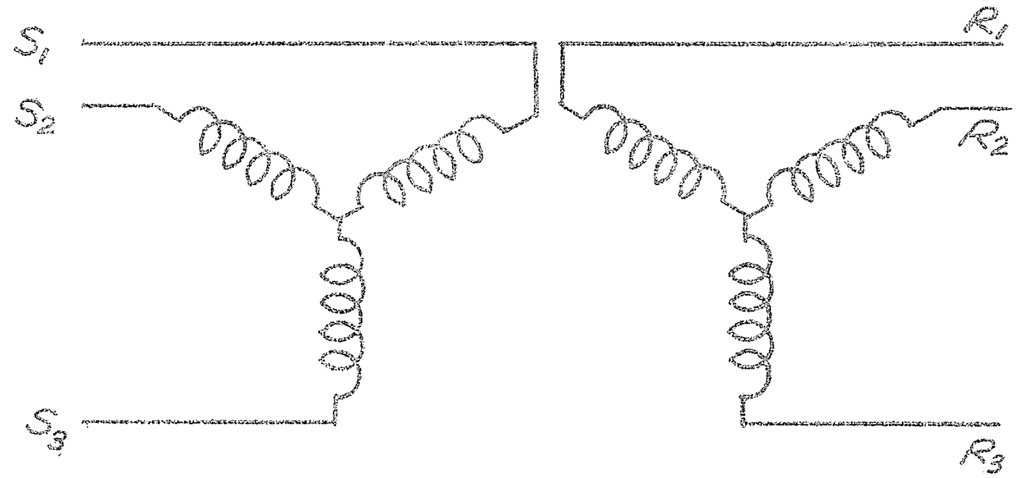
4. Receiver:



Function - When its two primary leads are excited by an alternating voltage and its other three input leads are excited by voltages which define an angle, it applies a torque to its shaft proportional to the sine of the difference between the angle and the angular position of its shaft with respect to its stator.

Application - Remote indicating systems.

5. **Differential Receiver:**



Function - When three of its input leads are excited by voltages which define an angle, and the other three of its input leads are excited by voltages which define a second angle, it applies a torque to its shaft proportional to the sine of the difference ~~angle~~ of the two angles.

Application - Remote indicating Systems.

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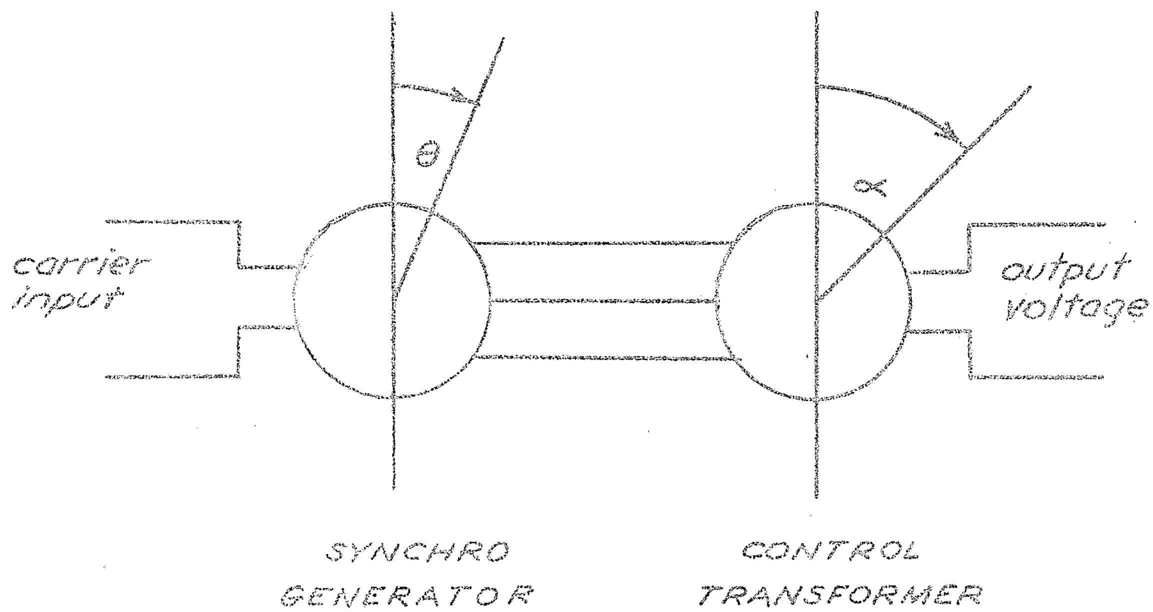
Synchro Generator, Control Transformer Pair

This is the most common form of electromagnetic transducer used to convert the angular position of a rotating shaft into an electrical signal. This type of system requires a synchro generator or synchro transmitter and a synchro control-transformer. These are set up as shown in fig. 2 with the three stator windings of each connected together and with an a.c. supply connected to the rotor of the generator. The useful output voltage is obtained from the rotor of the control transformer. This output voltage from the control transformer is used as an indication of the relative shaft positions of the two machines. Thus the system acts as a comparator of the two shaft positions and as a transducer, converting this difference of shaft positions into an electrical signal.

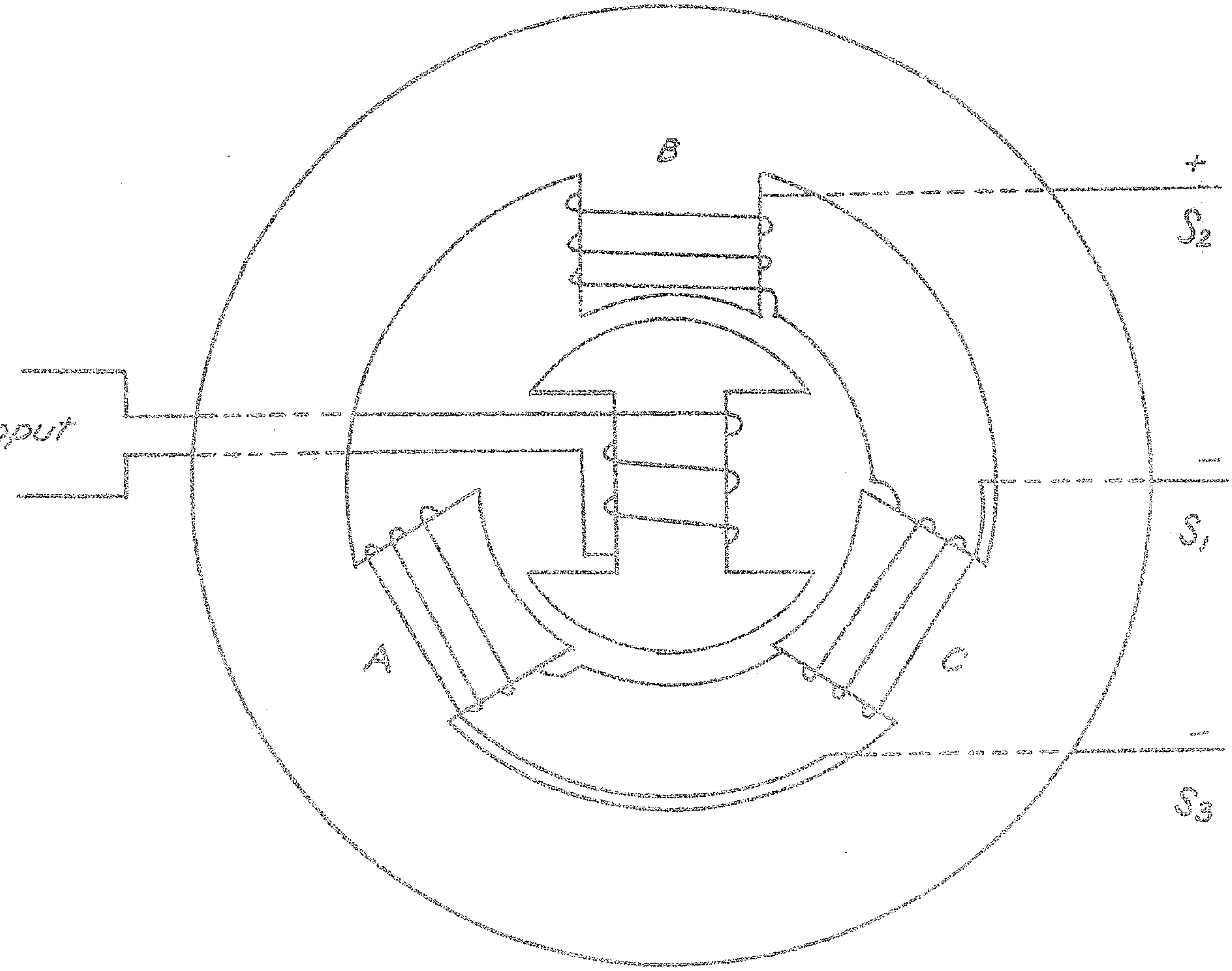
Synchro Generator

The principle of the synchro generator is the same as that of any transformer. An alternating current in the primary winding results in a magnetic field within the vicinity of the primary which changes with intensity and direction with the frequency of the alternating current. If a second winding is placed in the same magnetic field, a voltage is induced in the secondary and a current will flow through the secondary load. This is the result of electromagnetic induction by an alternate building-up and decaying of a magnetic field. If primary and secondary can be moved with respect to one another, then the



BASIC SYNCHRO PAIR

fig. 2



SYNCHRO GENERATOR

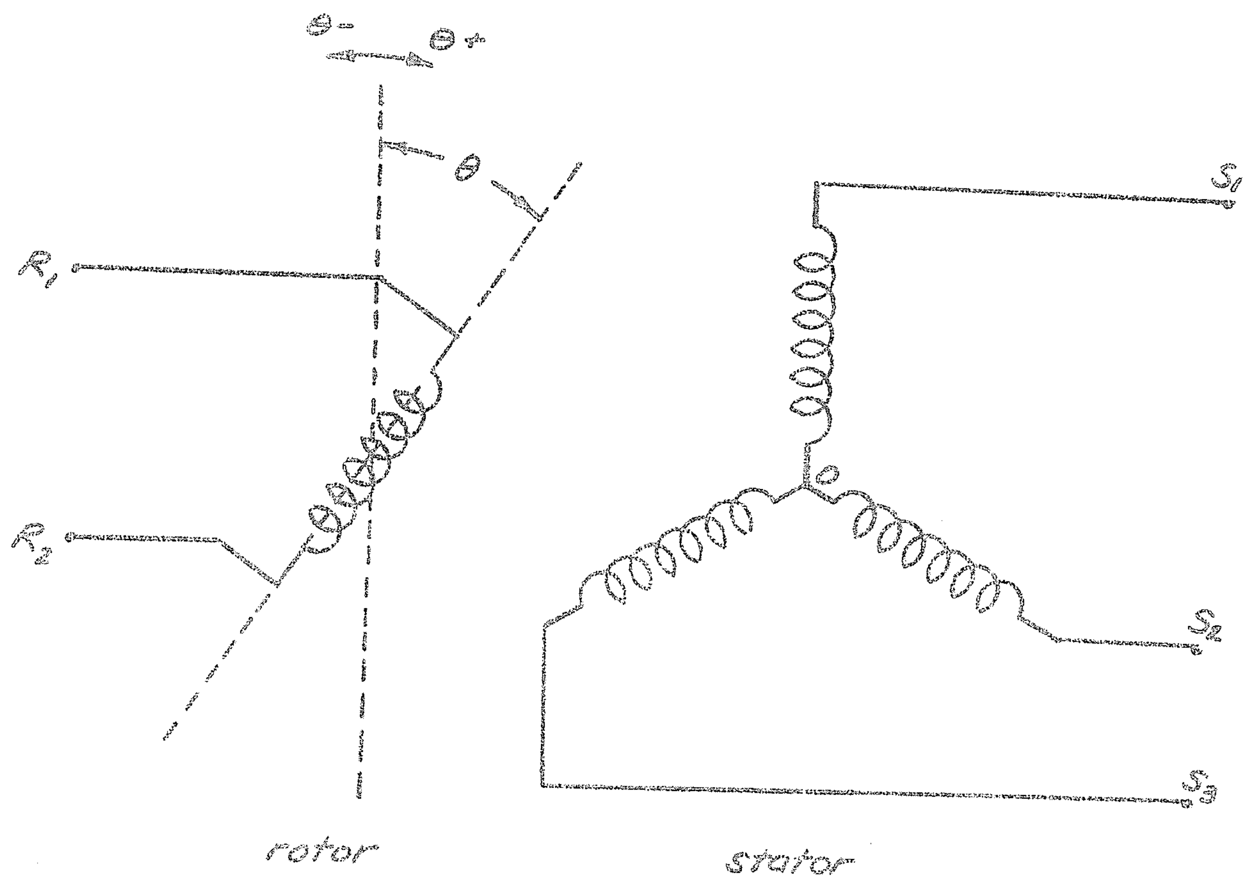
Fig. 3

maximum voltage is induced when the secondary is located so that a maximum flux from the primary passes through its windings. Depending on the direction of the flux lines with respect to the secondary, the polarity of the voltage in the secondary will be either in one direction or the other.

The primary of the synchro generator is wound on a rotor while the secondary is distributed in three separate windings on a stator as illustrated in fig. 3. Depending on the position of the rotor, more or less of the magnetic flux lines will cut through any one of the three windings. In the position shown, which is known as the zero position, voltages induced in windings A & C are equal and in the same direction. Hence the voltage between S_1 and S_3 is zero.

The voltage induced in B is of opposite polarity to either winding A or C. Thus at a given moment, the terminal of S_2 is of positive polarity as indicated, while at the same time polarities at S_1 and S_3 are negative. With the expanding and decaying of the magnetic field, polarities reverse, but a voltage still appears between terminals S_1 and S_2 and between terminals S_2 and S_3 but not between S_3 and S_1. Rotating the primary winding will gradually change this condition since the number and direction of flux lines which cut through the secondary windings is altered.

To analyze this more closely, consider the schematic



SCHEMATIC DIAGRAM of SYNCHRO GENERATOR

Fig. 1.

diagram in fig. 4. If the axis of a coil carrying an alternating current makes an angle θ with the axis of a second concentric coil, the induced emf in the second coil will be $K \cdot \cos \theta$, where K is a constant dependent upon the frequency and the magnitude of the current in the primary coil, the structure of the coils, and the characteristics of the magnetic circuit. If two additional secondary coils are added with their axes 120° and 240° from the axis of the first secondary coil, the emf will be, -

$$E_{os_1} = K \cdot \cos \theta$$

$$E_{os_2} = K \cdot \cos (\theta - 120^\circ)$$

$$E_{os_3} = K \cdot \cos (\theta - 240^\circ)$$

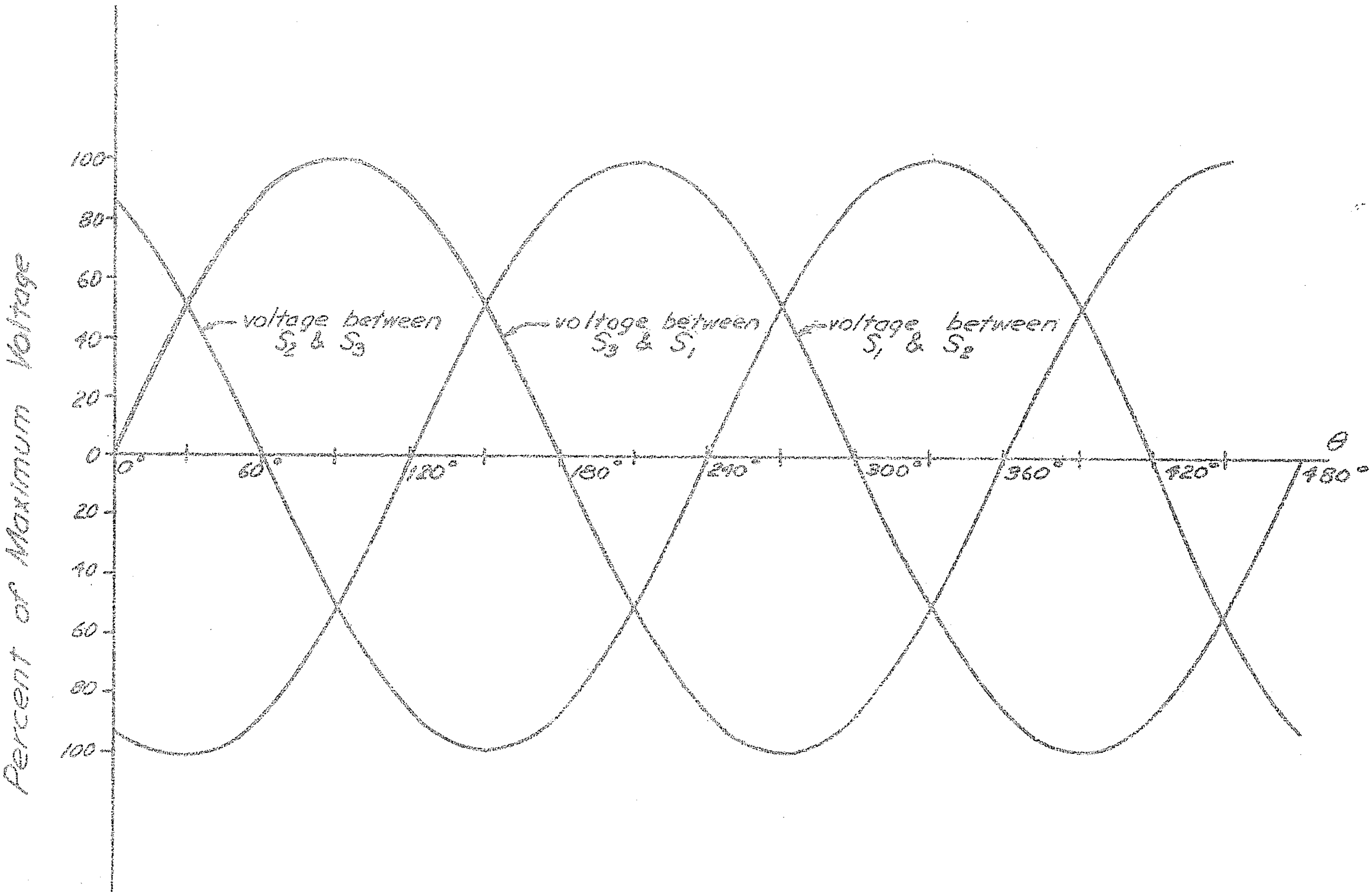
where the subscripts indicate the points between which the voltage is measured, and their order gives the sense of the measurement, (for example, E_{os_2} means the point S_2 is positive with respect to the point 0). The terminal voltages will then be:-

$$E_{s_2 s_1} = K \cdot \sqrt{3} \cos (\theta + 30^\circ)$$

$$E_{s_1 s_3} = K \cdot \sqrt{3} \cos (\theta + 150^\circ)$$

$$E_{s_3 s_2} = K \cdot \sqrt{3} \cos (\theta + 270^\circ)$$

Thus, it is seen that by applying an a.c. supply voltage to the generator, voltages are induced in the stator coils and consequently across the three stator output leads, which are dependent on the relative position of the rotor and the stator coils.



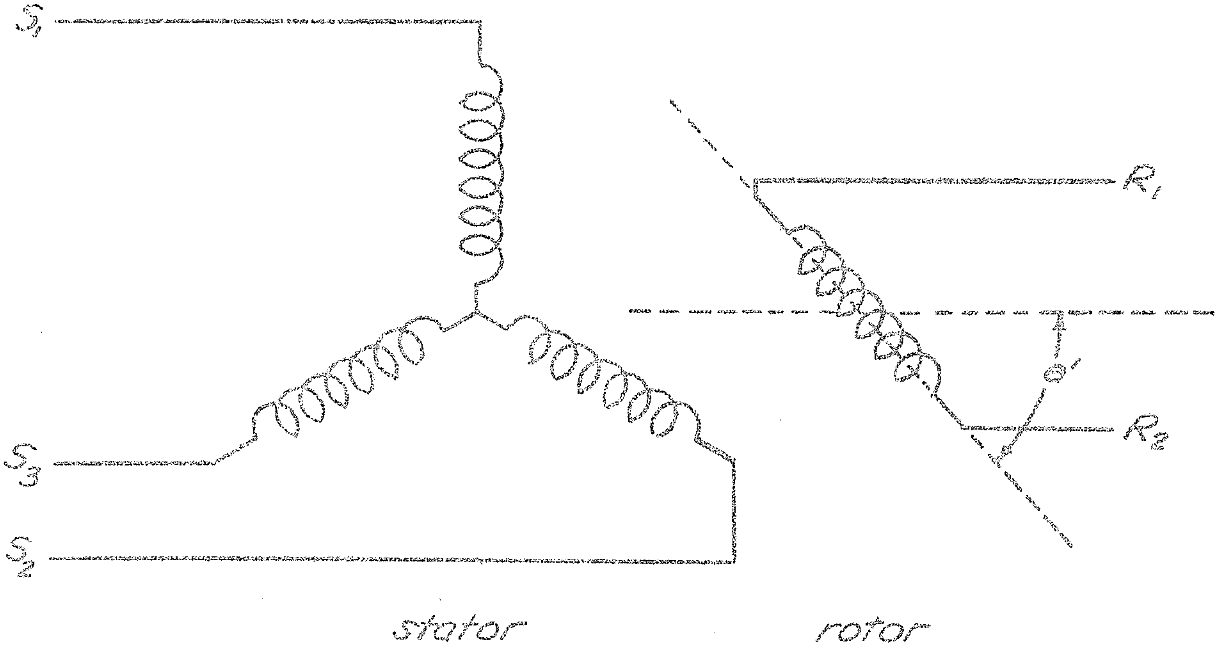
SYNCHRO GENERATOR OUTPUT VOLTAGES vs. ROTOR POSITION

fig. 5

Synchro Control Transformer

The control transformer is of essentially the same construction as the generator. The major difference is that the shape of the rotor in the control transformer is cylindrical, or of the umbrella type, while the rotor in the generator is of salient pole construction. This is done to prevent the magnetic field from producing a torque on the rotor of the control transformer. The only electric power connected to the control transformer is the signal power from the synchro generator. This is applied to the stator windings, which represent the primary of the transformer. The rotor windings therefore represent the secondary of the transformer. The voltage which will be induced in the secondary will depend on the position of the rotor with respect to the magnetic field which is set up by the currents flowing in the stator windings. For a field that corresponds to the zero position in the generator, the voltage induced in the secondary of the control transformer will be zero when the control transformer rotor is displaced by 90° from the generator rotor. This 90° displacement applies also for any other position of the rotor in the synchro generator. These last statements demand that the stators of both the generator and the control transformer positions be measured with respect to a common reference direction.

In the diagram of fig. 6 of the control transformer



SCHEMATIC DIAGRAM OF SYNCHRO CONTROL TRANSFORMER

fig. 6

windings, it is seen that if the rotor is set at 90° to the resultant flux which is produced in the control transformer by the currents in the three stator coils, no voltage will be produced across the rotor terminals R_1 R_2. If the angle of rotation θ' of the control transformer is less than θ (referring to fig. 4), an a.c. voltage will appear across R_1 R_2 with a slight phase lag, due to the reactances of the elements of the units, with respect to the line voltage. If θ' is greater than θ, an a.c. voltage is again observed across R_1 R_2 but with an additional 180° phase shift. The complex rms voltage across R_1 R_2 may be expressed as:

$$E_{R_1 R_2} = E_{max_{rms.}} \sin(\theta - \theta')$$

To explain this matter of the additional 180° phase shift, it may be more advantageous to look at the schematic of the control transformer which uses a different reference direction in defining the angle of turn of the control transformer rotor. Referring to fig. 7 it is seen that the angle θ here is set up in the same frame of reference as in the case of the generator, that is, as the angle between the axis parallel to the axis of S_1 and the axis of the rotor.

Considering the case where $\theta_t = \theta$, where θ_t is the angle of turn of control transformer rotor and θ is the angle of turn of the generator rotor, and neglecting the phase shift occuring between the generator and the control transformer due

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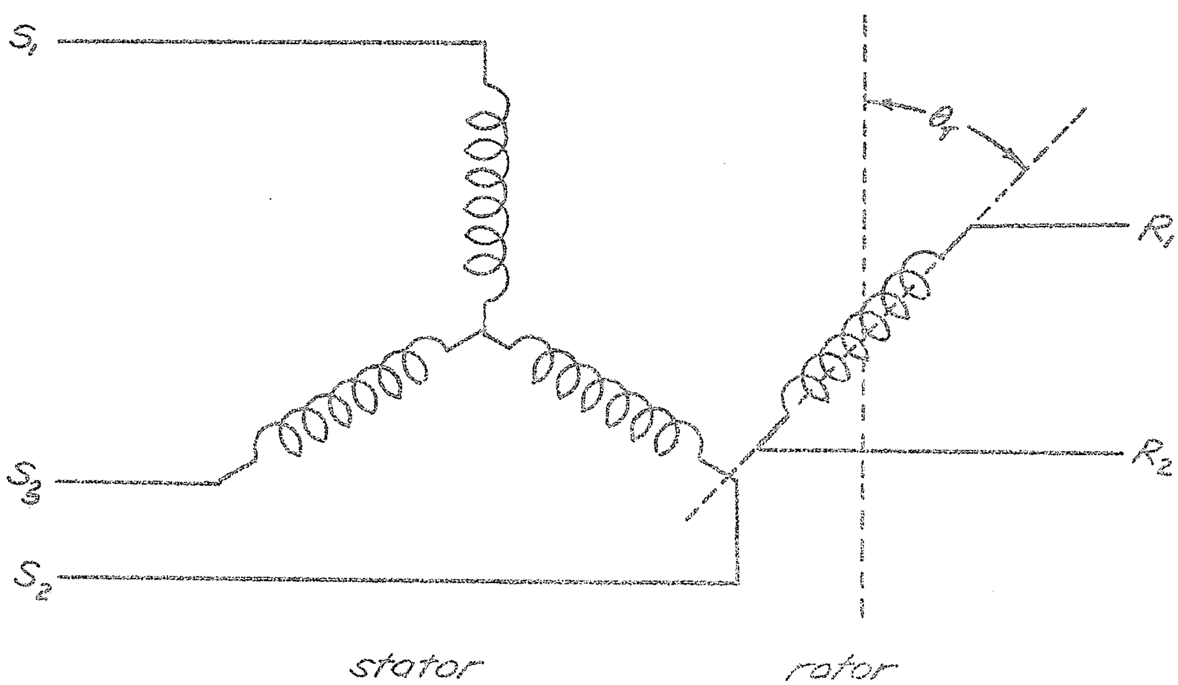

SCHEMATIC DIAGRAM of SYNCHRO CONTROL TRANSFORMER

fig. 7

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to reactances of the windings, the voltage appearing across R_1 R_2 of the control transformer will be $M \cdot E_R$, and where E_R is the voltage applied to the generator rotor and M is a constant to account for the losses between the two units. Let the flux producing the voltage $M \cdot E_R$ be represented by ϕ, therefore $M \cdot E_R$ will be proportional to ϕ, or $E_R = P\phi$, where P is a constant of proportionality. If it is assumed that the generator rotor remains fixed and the transformer rotor changes in position, the flux will remain constant but the flux parallel to the axis of the transformer rotor will decrease. Since the steady state voltage produced across a coil is directly proportional to that flux which is parallel to its axis, E_R thus becomes:-

$$E_R = P\phi \cos (\theta_t - \theta)$$

If ϕ remains constant, this equation can be written:-

$$E_R = E_{max} \cdot \cos (\theta_t - \theta).$$

Thus it is obvious that when $90^\circ < |\theta_t - \theta| < 270^\circ$, E_R will become negative, or if preferred, will assume an additional 180° phase shift. This therefore explains why, in the previous analysis, if θ' becomes greater than θ, E_R takes on an immediate 180° phase shift.

CHAPTER 2

SYNCHRO SYSTEM OPERATION

......../21 a.

Operation of the Synchro Pair

With the previous explanation of the operation of each of the components of the simple two unit synchro system it becomes relatively simple to explain the basic synchro pair. The alternating current flowing in the generator rotor sets up, by transformer action, voltages across the coils of the stator. These voltages cause currents of different magnitudes, but all of the same phase, to circulate in the stator windings. The result is that an alternating flux pattern is set up in the stator of the control transformer. This flux pattern has ideally the same space orientation as that in the generator. If the reluctance of the air gap of the control transformer is independent of rotor position, (this is usually a fairly valid assumption for cylindrical rotor synchros) the output of the control transformer is an a.c. voltage proportional to the cosine of the angular difference between the axes of the rotors of the generator and control transformer. Thus if the voltage applied to the generator rotor is $E \sin \omega_c \cdot t$, then the instantaneous control transformer out voltage is:

$$e_o = E_{o_{max}} \cdot \cos(\theta_t - \theta) \cdot \sin(\omega_c \cdot t - \beta) \text{ or}$$

$$e_o = E_{o_{max}} \cdot \sin(\theta' - \theta) \cdot \sin(\omega_c \cdot t - \beta)$$

where θ_t, θ', and θ are the angles denoted previously, and $E_{o_{max}}$ is the maximum value of the output voltage. The phase shift β between input and output voltages corresponds to the

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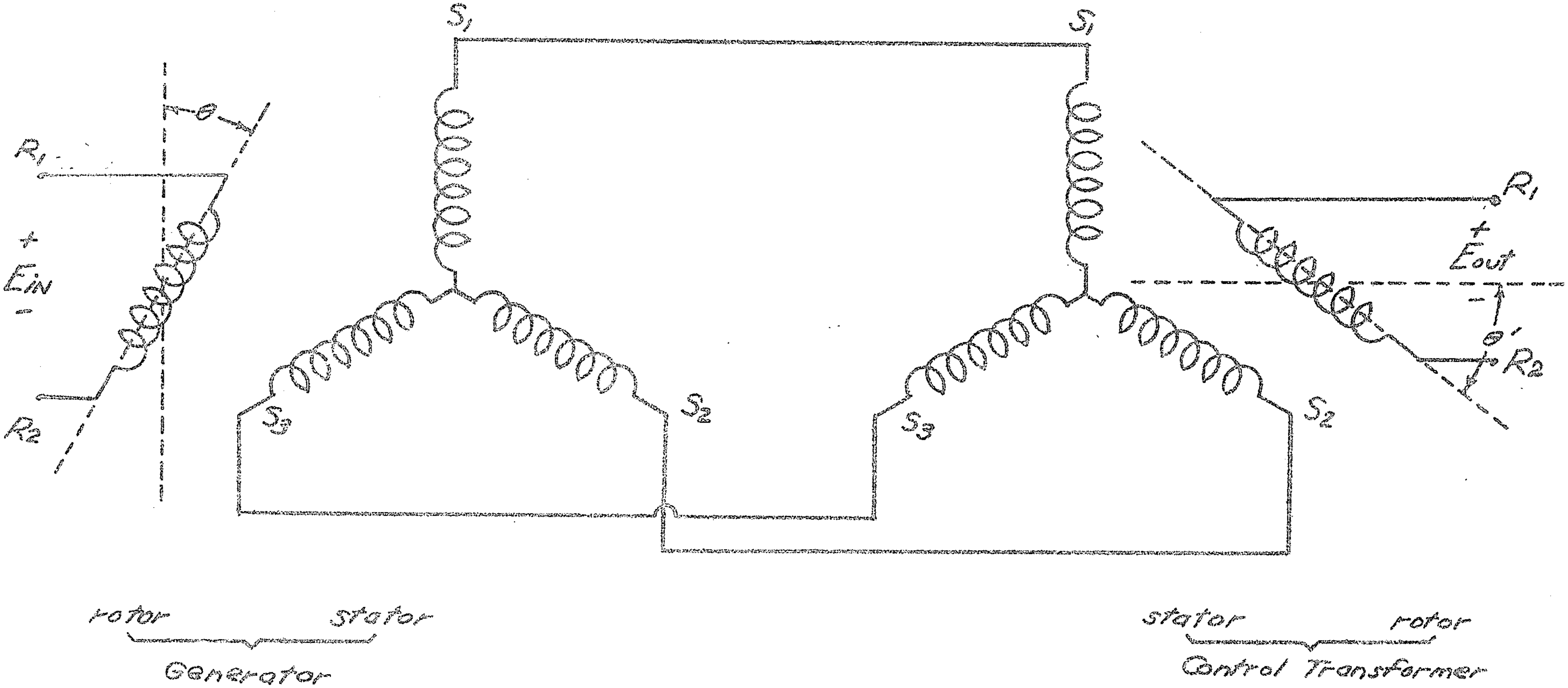

BASIC SYNCHRO PAIR

fig. 8

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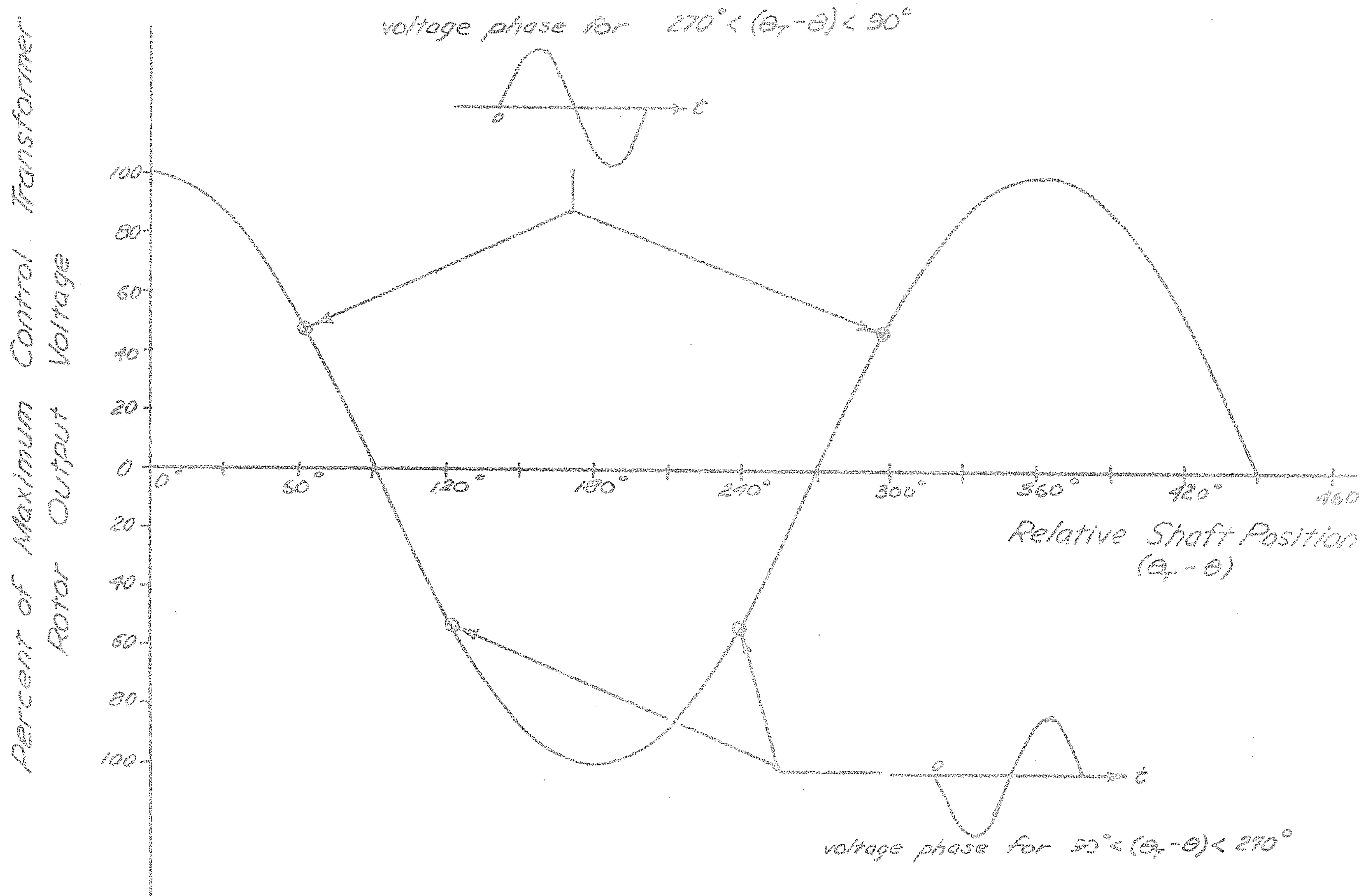
sort of phase shift found between the primary and secondary voltages in an ordinary transformer and is due to the impedances of the windings. In actuality, β is usually of the order of only a few degrees and is therefore usually neglected.

A plot of the output voltage from the control transformer vs. the angular difference between θ_t and θ, for a fixed angular position of the generator rotor, is shown in fig. 9. The polarity of this voltage is indicated by whether the voltage is plotted as a positive quantity or a negative quantity, as shown in the diagram by inserted voltage vs. time plots for the same starting instant. The negative voltage is, as mentioned previously, 180° out of phase with the positive voltage.

If the synchro generator is placed at any angle, there will be two positions of the control transformer shaft which will yield zero output voltage from its rotor windings. A servomechanism must be able to distinguish these two positions, which are 180° of shaft rotation apart.



CONTROL TRANSFORMER ROTOR VOLTAGE vs. SHAFT POSITION INCLUDING VOLTAGE PHASE RELATION

Fig. 9.

Indicating Synchros

There is a class of synchros which are used for remote positioning and indication that do not require servomechanisms to position the shafts. A typical example is the combination of a synchro generator and a synchro receiver. These two units are electrically identical, however, the receiver has a mechanical damper attached to its shaft while the generator does not. The synchro receiver has a constant voltage applied to its rotor, and the stator leads are connected to the corresponding leads of the synchro generator. If the shaft of the generator is held fixed, the shaft of the synchro receiver rapidly rotates until it is at the same angle as the generator shaft. If the shaft of the generator is moved, the receiver follows that movement.

Such a pair of synchros is used in cases where it is desired to indicate remotely the position of a shaft. The receivers, however, have only a small amount of torque to exert against any load and are consequently used only for indicating a shaft position on a dial. Their inaccuracy makes them undesirable for even this purpose and a servomechanism, which of course is much more complex, must sometimes be substituted. Another disadvantage is their poor dynamic response. If the shaft of a generator is oscillated sinusoidally at a certain critical frequency, the shaft motion of the receiver can amplify this movement as much as ten times. This indicates that this type of synchro system is very oscillatory and not nearly as highly damped as the response of a servo can be made. Furthermore, dynamic inaccuracies on the part of the receiver or static errors caused by loads or sticky bearings change the magnitudes of the voltages on the stator leads supplying it. Thus if any other receiver is connected to the same leads, its stator voltages and hence its rotor position will be in error because of the error of the first receiver.

This type of synchro is very seldom, if ever, used as a component of servomechanisms and therefore discussion of its characteristics shall not be further expanded upon here.

Operation of the Synchro Pair with Continuous Oscillation of the Generator Rotor

Thus far it has been assumed that the angles θ and θ', referring to fig. 9, have remained constant with time. Suppose however that either θ or θ' (or both) is oscillated sinusoidally with a relatively low frequency of ω_s radians per second. (It will be shown later in the discussion of the operating characteristics of discriminators that practical limitations force ω_s to be about 0·1 to 0·25 times the carrier frequency of the system). Thus the relative angular position of the two shafts can be expressed as:

$$\theta - \theta' = \sigma_m \cdot \sin\left(\omega_s \cdot t + \theta_0\right) + \Delta\theta$$

where θ_0 accounts for the displacement between the time when

$(\theta - \theta' - \Delta\theta)$ is zero and the $t = 0$ point referred to the carrier wave. $\Delta\theta$ represents any constant angular displacement about which the relative angular difference $(\theta - \theta')$ oscillates. σ_m is the amplitude of the sinusoidal displacement between θ and θ'. From previously developed theory the output voltage from the control transformer can be expressed as:

$$e_o = E_m \sin(\theta - \theta') \sin(\omega_c \cdot t - \beta)$$

where ω_c is the carrier frequency and β is the phase shift of the carrier signal between the input to the synchro generator rotor and the output from the control transformer. Substituting for $\theta - \theta'$ yields:

$$e_o = E_m \sin \left[\sigma_m \sin(\omega_s \cdot t + \theta_o) + \Delta\theta\right] \cdot \sin(\omega_c \cdot t - \beta).$$

Now if σ_m and $\Delta\theta$ are small this expression can be reduced by introducing the approximation:

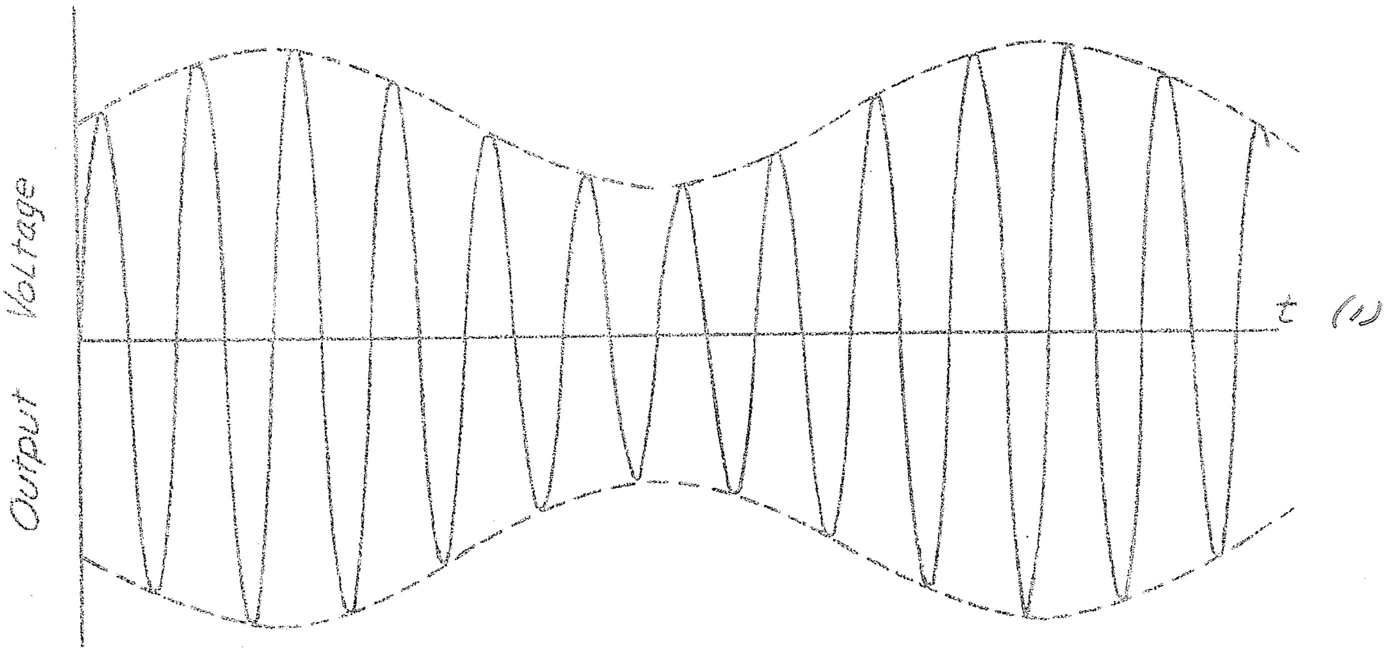
$$\sin \left[\sigma_m \sin(\omega_s \cdot t + \theta_o) + \Delta\theta\right] \doteq \sigma_m \sin(\omega_s \cdot t + \theta_o) + \Delta\theta$$

thus yielding:

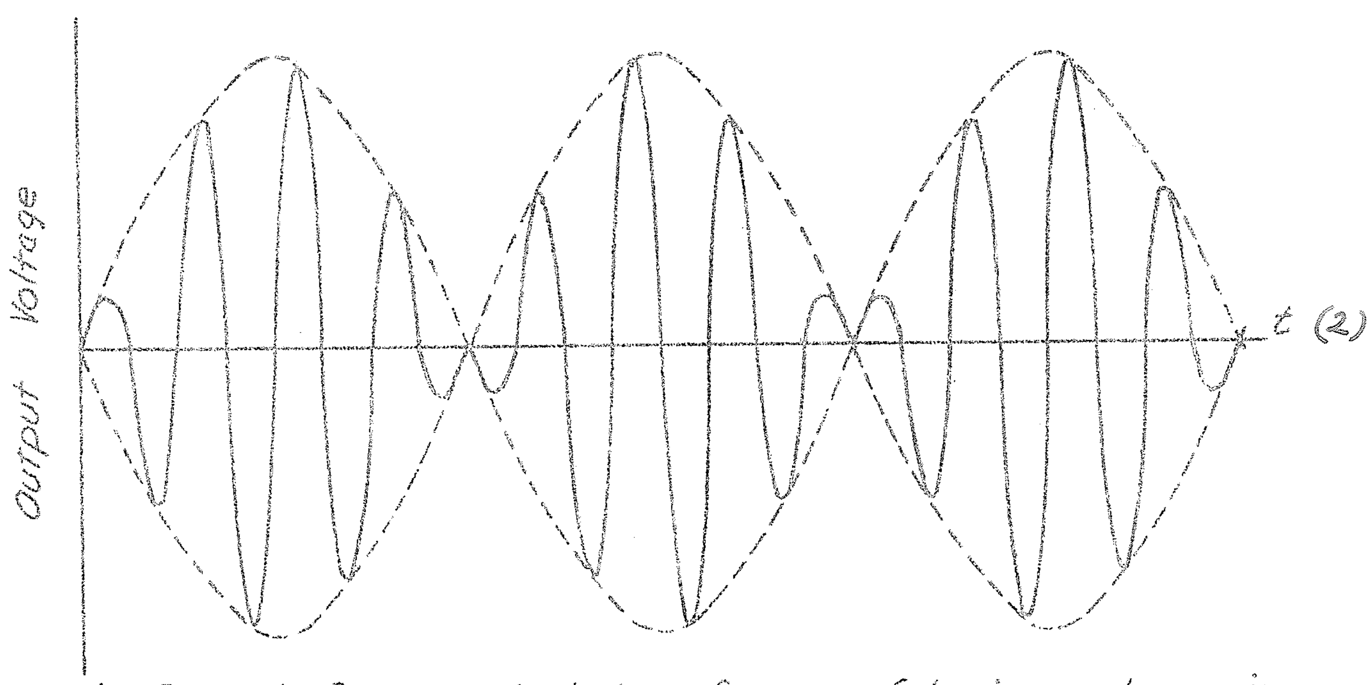
$$e_o = E_m \left\{ \sigma_m \sin(\omega_s \cdot t + \theta_o) + \Delta\theta \right] \cdot \sin(\omega_c \cdot t - \beta).$$

This voltage will appear as a wave of the same frequency as the original carrier sine-wave amplitude modulated at a frequency corresponding to the angular frequency of oscillation of the rotor (ω_s). This is shown in fig. 10a.

In normal control systems the synchro pair operates close to the point where the pair is in correspondence. Thus for oscillation at this operating point $\Delta\theta$ would be zero and



Output from control transformer of basic synchro pair for sinusoidal oscillation of generator rotor about a point away from the correspondence point. ($\Delta\Theta > \sigma_m$)



Output from control transformer of basic synchro pair for sinusoidal oscillation of generator rotor about the correspondence point.

fig. 10 a.

the output voltage equation becomes:

$$e_o = E_m \cdot \sigma_m \cdot \sin(\omega_s \cdot t + \theta_o) \sin(\omega_c \cdot t - \beta)$$

Considering the situation where β is small enough to neglect and also where the time axis of the sinusoidal oscillation is so chosen that $\theta_o = 0$ the output voltage equation can be expressed as:

$$e_o = E_m \cdot \sigma_m \cdot \sin(\omega_s \cdot t) \sin(\omega_c \cdot t)$$

Placing this in an equivalent form it becomes clear that this voltage waveform is a type of suppressed carrier modulation with e_o given by:

$$e_o = \frac{E_m \cdot \sigma_m}{2}\left[\cos(\omega_c - \omega_s)\cdot t - \cos(\omega_c + \omega_s)\cdot t\right]$$

Thus the waveform is composed of two sidebands having frequencies $(\omega_c - \omega_s)$ and $(\omega_c + \omega_s)$ but no carrier frequency component. The form of this voltage is typical of the output of a large number of transducers used in servomechanisms in which the input variable is used to modulate the amplitude of an a.c. carrier.

In this analysis for continuous rotor oscillation there is an inherent characteristic which must be carefully regarded. The type of waveform that will be produced across the output of the control transformer rotor will depend on the relative magnitudes of σ_m and $\Delta\theta$. For $\Delta\theta$ larger than σ_m the output will appear as a normal amplitude modulated carrier wave such as is shown in fig. 10a. As $\Delta\theta$ becomes

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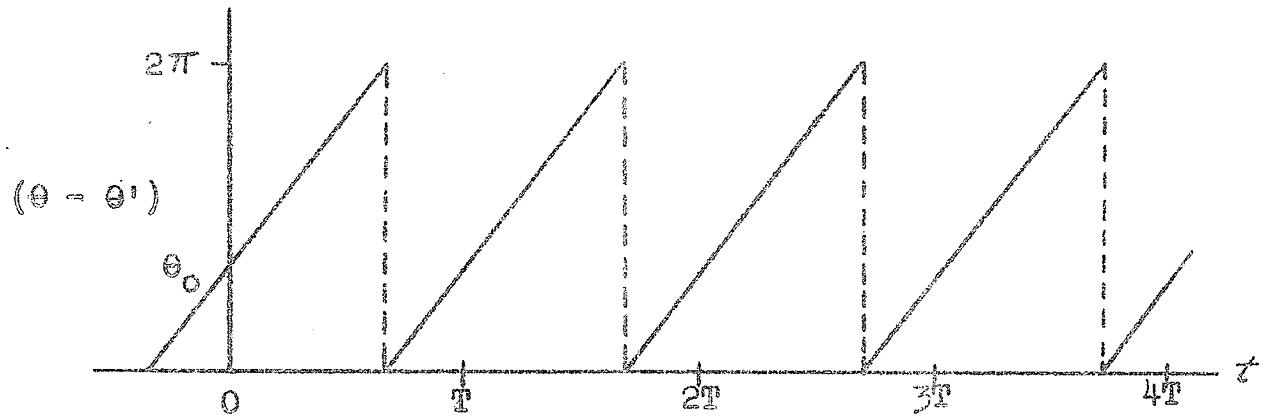
less than σ_m however the value of $\sigma_m \cdot \sin(\omega_s t + \theta_o) + \Delta\theta$ will no longer always be greater than zero and for certain portions of the oscillation cycle it will become negative. This negative value corresponds to a 180° phase shift in the signal and hence the waveform departs from the normal amplitude modulated form of fig. 10a(1). For oscillation about the point where the two units produce a null output, that is where $\Delta\theta = 0$, the waveform appears as in fig. 10a(2) (page 27) and is the same type of waveform as would be obtained for continuous rotor rotation. It must be noted however that whenever $\Delta\theta$ does not equal zero the form of the output wave will change and besides the two sideband components will then contain a component at the carrier frequency and hence the system no longer produces suppressed carrier modulation.

Operation of the Synchro Pair with Continuous Rotation of the Generator Rotor

This analysis is very similar to the analysis for continuous oscillation except here the displacement $(\theta - \theta')$ follows a pattern such as shown below. This pattern occurs whenever the generator rotor has an angular velocity of ω_m radians per second relative to the rotor of the control transformer regardless of which one of the rotors is moving or whether both are moving.



$$T = \frac{2\pi}{\omega_m}$$

For one complete generator rotation:

$$\theta - \theta' = \frac{2\pi t}{T} + \theta_o \text{ but } T = \frac{2\pi}{\omega_m}$$

therefore: $\theta - \theta' = 2\pi t \cdot \frac{\omega_m}{2\pi} + \theta_o = \omega_m t + \theta_o$

Substituting in the instantaneous output voltage equation:

$e_{out} = E_m \sin(\theta - \theta') \sin(\omega_c t - \beta)$ yields:

$$e_{out} = E_m \sin(\omega_m t + \theta_o) \sin(\omega_c t - \beta)$$

It is therefore seen that the output from the control transformer rotor, with continuous rotation of the generator rotor with respect to the control transformer rotor, appears as an amplitude modulated sine wave with the modulating frequency equalling the relative frequency of rotation of the two shafts. It might be worthy of further emphasis to note that when the two shafts are rotating at the same angular velocity the output

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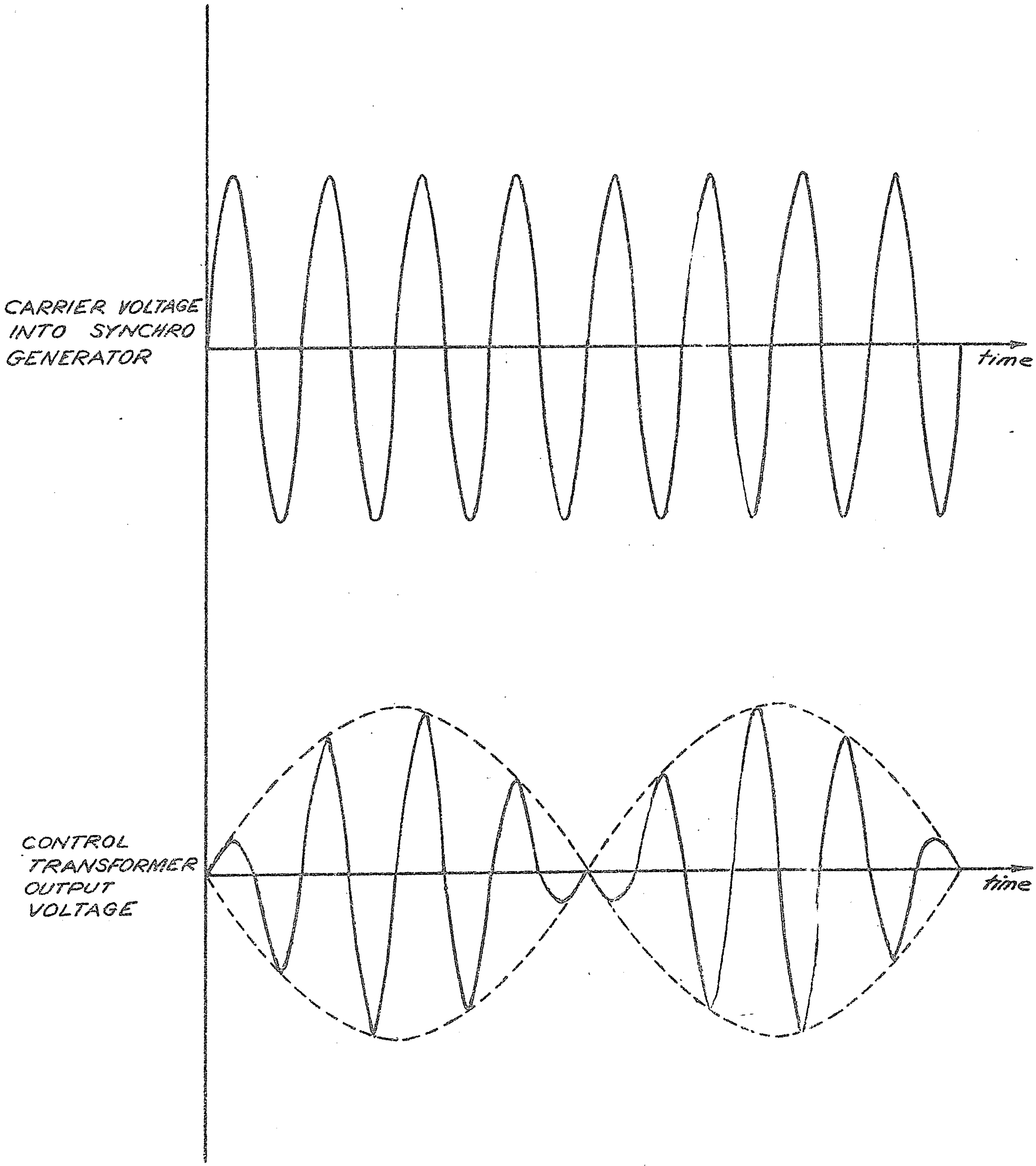
will be constant at an amplitude dependent on the relative shaft positions which is denoted by θ_o. If rotation is started in such a manner that $(\theta - \theta') = 0$ at $t = 0$, and if β is small enough to be negligible, the output equation can be simplified and placed in the equivalent form:

$$e_{out} = \frac{E_m}{2} \left[\cos (\omega_m - \omega_c) t - \cos (\omega_m + \omega_c) t \right]$$

This is the same form of suppressed carrier output signal as was obtained with continuous oscillation of the rotor when $\Delta\theta = 0$. The carrier is completely eliminated from the output and the sidebands differ in frequency from the carrier by an amount equal to the frequency of generator rotation.

The waveform of the output signal is shown in fig. 10b. and displays the effect of the 180° phase shift in the carrier wave whenever the relative angular displacement $(\theta - \theta')$ is in the 180° to 360° region.

This analysis is, as stated previously, based on a relative angular velocity of the synchro generator and control transformer shafts of ω_m radians per second. Caution must be exercised in applying this development to the typical control system such as the speed control of a d.c. motor. In the speed control system the d.c. motor whose speed is being regulated drives a synchro generator. This of course constitutes continuous rotation of the generator shaft at, for a particular situation, a constant shaft angular velocity. The synchro



VOLTAGE WAVEFORMS for CONTINUOUS ROTATION of SYNCHRO ROTOR

fig. 10b.

control transformer is driven in continuous rotation by a second "reference" prime-mover, at the speed at which it is desired to have the d.c. motor operating. Considering the situation where the d.c. motor and the reference prime mover are running at exactly the same speed, and the synchro generator and control transformer rotors are in correspondence, thus yielding zero output voltage from the control transformer, this situation will remain as long as the two units continue to rotate at the same speed. If the d.c. motor shaft begins to speed up or slow down this will throw the two synchros out of correspondence and hence produce a non-zero output from the control transformer. This output signal will now obey the general output voltage equation:

$$e_{out} = E_m \sin \omega_m t \ \sin (\omega_c t - \beta)$$

Now in this type of system ω_m will in general not be constant over any great time interval since the system immediately begins to correct itself and hence the modulating frequency will be continually varying and continues to do so until the two shafts are rotating at the same speed and are once again in correspondence. The value of the modulating frequency will be dependent on the speed of response of the control loop. Analytical representation of the correcting signal, that is the output signal from the control transformer, therefore becomes a bit cumbersome whereas in a qualitative sense the situation is relatively easily described by simply picturing a correcting signal being produced every time the two shafts move out of correspondence in any way and this correcting signal being reduced in magnitude as the system drives itself back into the correct mode of operation. To produce an analytical expression for e_{out} for this type of situation would necessitate knowledge of the speed of response of the system in analytical form and then, with the proper modification, the previous development for the output voltage from the control transformer for continuous constant speed shaft rotation can be used to describe the type of control system which is probably the most common application of synchros in continuous rotation. In an actual frequency response domain analysis of speed control system this degree of complexity in the analytical treatment is not necessary since only the small signal transfer function across the synchro pair is required. In obtaining this transfer function it is usually assumed that the amount of deviation from the correspondence position of the two shafts is quite small at all times and hence the output voltage is taken as being directly proportional to the angle of relative displacement of the shafts.

C H A P T E R 3

T H E D I F F E R E N T I A L G E N E R A T O R

The Differential Generator or Differential Transmitter

If it is desired to position the controlled shaft of a servomechanism to a shaft position which is the sum or difference of two shaft angles, a synchro differential generator can be used. A very important example of the use of such device is in a gunfire control system when it is desired to add angles, such as lead angles or parallax angles. The differential generator (or transmitter) is very similar in construction to the synchro generator and to the control transformer except that the differential generator is wound with three windings on both the primary and secondary. It is usually wound with a 1:1 turns ratio. The rotor is of the slotted drum type and the windings of the rotor and the stator are of the standard distributed type with each of the windings acting as if they were physically separated 120° from the other two. Despite its construction similarity to a three phase machine, it operates on single phase like other synchro devices discussed thus far.

When the stator leads are connected to the output leads of a synchro generator the currents flowing produce a flux in the differential generator which is in the same orientation as the flux in the synchro generator. This takes place in the same manner as the reproduction of generator flux in the control transformer of the basic synchro pair.

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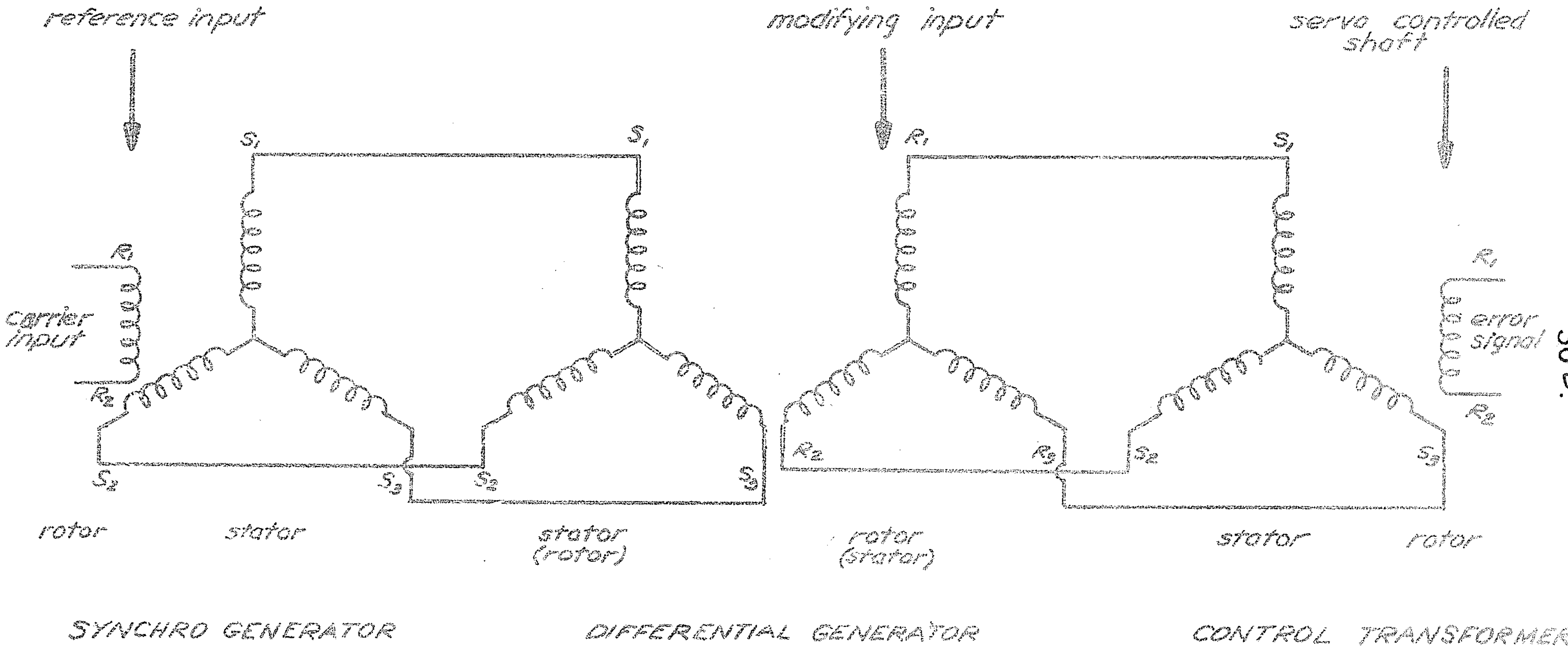


SYNCHRO SYSTEM USING DIFFERENTIAL GENERATOR

fig. 11

The rotor of the differential generator however has voltages induced across its terminals of a magnitude and polarity which are a measure of the angular position of the rotor relative to the resultant flux produced by the stator currents. Since the rotor of the differential generator has the same 120° orientation of its windings as does the stator, a given flux in the differential generator produces the same three output voltages across its R_1 R_2 R_3 leads as would be produced across the S_1 S_2 S_3 leads of the ordinary synchro generator which is feeding the differential unit except that these voltages are now a function of the differential generator rotor angle as well as the angular position of the synchro generator rotor. It is worthwhile to note in passing that in practice it is also possible to use the rotor of the differential generator as the primary and the stator as the secondary without changing the effect of the unit.

To explain the details of the operation of the differential generator consider the analysis of a system with a differential generator inserted between a generator and a control transformer. For simplicity, it is assumed that the synchro generator is perfect, free from errors and unbalances. From the discussion of the operation of the synchro generator it is known that the output voltages from the synchro generator are:

$$E_{os_1} = K \cos \theta$$

$$E_{os_2} = K \cos (\theta - 120^\circ)$$

$$E_{os_3} = K \cos (\theta - 240^\circ)$$

Considering the loads reflected into the primary of the differential generator to be purely resistive and also balanced, the following mesh analysis can be made (referring to fig. 12a.)

$$\begin{bmatrix} E_{os_1} - E_{os_3} \\ E_{os_1} - E_{os_2} \end{bmatrix} = \begin{bmatrix} 2R & R \\ R & 2R \end{bmatrix} \cdot \begin{bmatrix} I_b \\ I_a \end{bmatrix}$$

$$E_{os_1} - E_{os_3} = E_{s_3 s_1} = \sqrt{3}K \cos (\theta - 30^\circ)$$

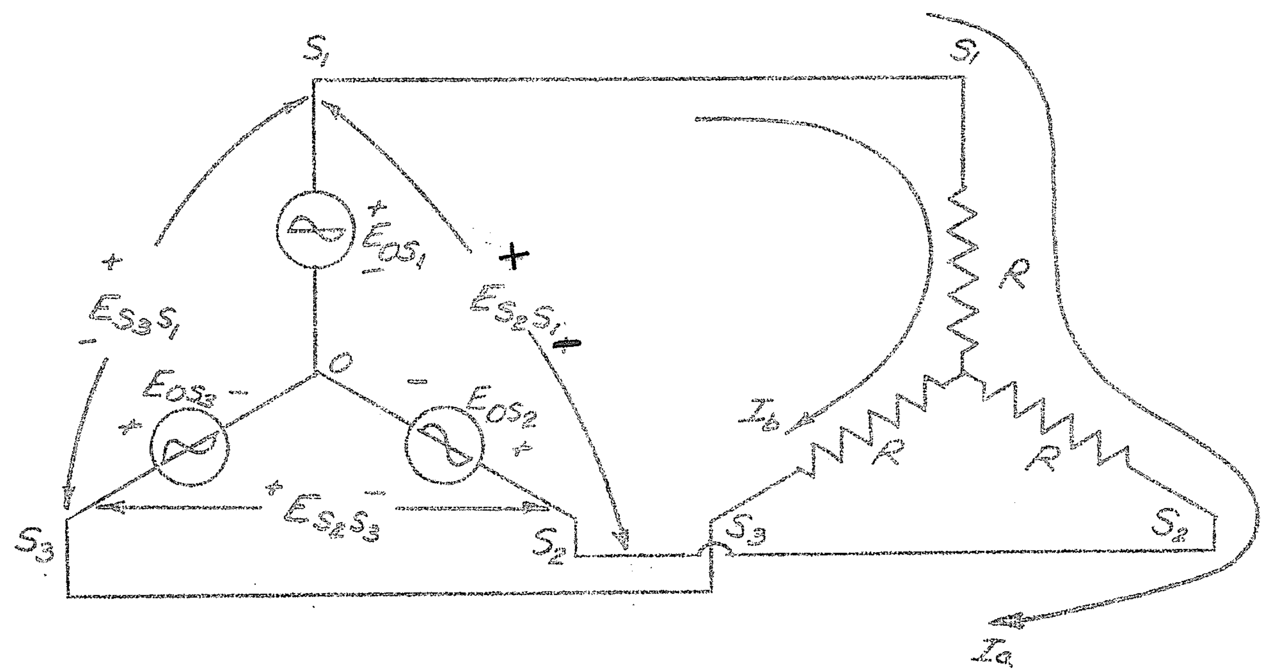
$$E_{os_1} - E_{os_2} = E_{s_2 s_1} = \sqrt{3}K \cos (\theta + 30^\circ)$$

Substituting these into the matrix equation and solving for the currents by Cramer's Rule yields:

$$I_a = \frac{K}{R} \cos (\theta + 60^\circ) = I_m \cos (\theta + 60^\circ)$$

$$I_b = \frac{K}{R} \cos (\theta - 60^\circ) = I_m \cos (\theta - 60^\circ)$$

The analysis is also simplified by assuming that the differential generator has unity turns ratio and that the coefficient of coupling between stator and rotor circuits is unity. From the standard transformer theory it is known that the demagnetizing effect of the secondary current must be balanced by an automatic increase of primary current to such an extent that



APPROXIMATE EQUIVALENT CIRCUIT OF a SYNCHRO GENERATOR SHOWING LOADING WHEN CONNECTED TO a DIFFERENTIAL GENERATOR.

fig. 12a.

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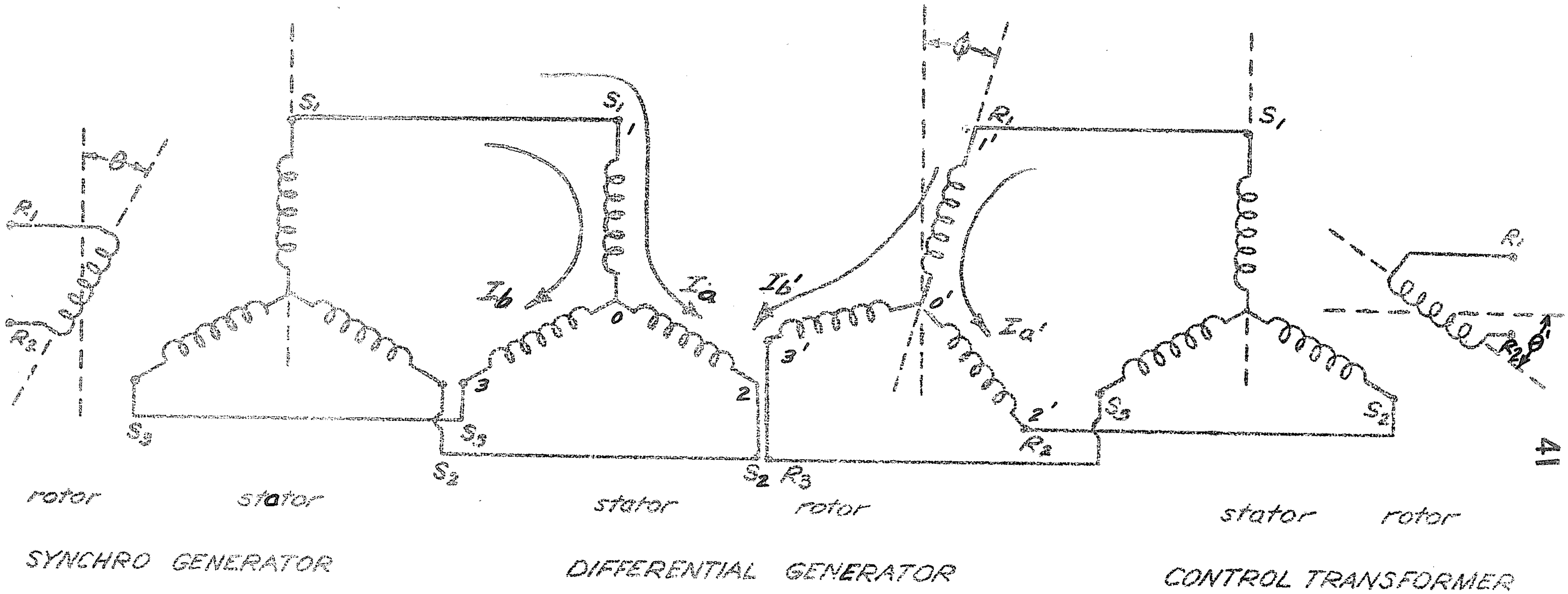
the resultant of the primary and secondary mmf's (ampere-turns) will be just sufficient to maintain the flux at a value corresponding to the counter emf demanded by the new conditions. These conditions must also be satisfied in a differential generator and therefore the amp-turns acting along any axis in the differential generator stator must be equal and opposite to the amp-turns acting along the same axis in the rotor. For an example of this consider the axis passing through the leg 0'2' of the rotor (see fig. 12b.). The amp-turns contributed by the rotor along this axis are:

$$mmf_{0'2'} = N \cdot I_{a'} + N\,(I_{a'} + I_{b'}) \cos 60^\circ + N \cdot I_{b'} \cos 120^\circ$$

where N is the effective number of turns on each winding. This equation assumes the windings to be so distributed that the amp-turns measured in any axis are proportional to the cosine of the angle between the axis under consideration and the axis of the winding contributing the mmf. Thus the angle between the reference axis 0'2' and the leg 1'0', carrying current $(I_{a'} + I_{b'})$, is 60°, and the contribution to the mmf along the reference axis is given by the second term in the expression. The angles between the three stator windings and the axis 0'2' are $(60^\circ - \phi)$, $(-\phi)$, and $(120 - \phi)$ for the legs 10, 02, and 03 respectively. The rotor angle ϕ is measured as shown in fig. 12b. It is noted that the direction in which the current is assumed to flow in the winding is important. The stator amp-turns contributed to axis 0'2' are:



SYNCHRO SYSTEM WITH DIFFERENTIAL GENERATOR

Fig 12 b.

$$\text{mmf} = N(I_a + I_b)\cos(60^\circ - \phi) + N \cdot I_a \cos\phi + N \cdot I_b \cos(120^\circ - \phi).$$

Equating these two expressions for mmf along O'2' yields:

$$N\left[I_{a'} + (I_{a'} \pm I_{b'})\cos 60^\circ + I_{b'}\cos 120^\circ\right] =$$
$$N\left[(I_a + I_b)\cos(60^\circ - \phi) + I_a\cos\phi + I_b\cos(120^\circ - \phi)\right]$$

$$I_{a'}(1 + \cos 60^\circ) + I_{b'}(\cos 60^\circ + \cos 120^\circ) =$$
$$I_a\left[\cos(60^\circ - \phi) + \cos\phi\right] + I_b\left[\cos(60^\circ - \phi) + \cos(120^\circ - \phi)\right]$$

$$\tfrac{3}{2}\, I_{a'} + 0\, I_{b'} = I_a\left[(\tfrac{1}{2} + 1)\cos\phi + \frac{\sqrt{3}}{2}\sin\phi\right] +$$
$$I_b\left[(\tfrac{1}{2} - \tfrac{1}{2})\cos\phi + \left(\frac{\sqrt{3}}{2} + \frac{\sqrt{3}}{2}\right)\sin\phi\right]$$

$$\tfrac{3}{2}\, I_{a'} = \sqrt{3}\left[I_a\cos(30^\circ - \phi) + I_b\sin\phi\right]$$

now substituting: $I_a = I_m\cos(\theta + 60^\circ)$

$$I_b = I_m\cos(\theta - 60^\circ)$$

yields: $I_{a'} = I_m\cos(\theta - \phi + 60^\circ)$.

In a similar fashion, and by using leg O'3' as the reference axis, it is seen that:

$$I_{b'} = I_m\cos(\theta - \phi - 60^\circ)$$

The form of $I_{a'}$ and $I_{b'}$ is the same as that of I_a and I_b except that the angle θ has been replaced by the angle $(\theta - \phi)$. Hence the effect of the differential generator is to change the control transformer angle for zero output voltage to $\theta' = (\theta - \phi)$ rather than $\theta' = \theta$, the value necessary when the differential generator is not present. Thus this shaft angle of the differential generator is added (or subtracted) from the

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angle of the input.

Therefore, the output voltage for the control transformer can be written:

$$E_{R_1R_2} = E_m \sin (\theta - \theta' \pm \phi)$$

where $E_{R_1R_2}$ is the output voltage and E_m is the maximum value of this voltage. The sign attached to ϕ depends on the direction of rotation of the differential generator rotor and must be determined experimentally.

It is of course possible to use more than one differential generator when more than one additional input must be inserted into the system.

In practice, differential generators are subject to the same sort of errors as other synchros. Also the coefficient of coupling between stator and rotor circuits is less than unity and therefore the differential generator requires a fairly large magnetizing current, which must be supplied by the synchro generator. This current produces heating in the generator, and it places a limit on the number of differential generators that can be excited by a synchro generator of a given size. The magnetizing current may be reduced quite considerably, at least as far as the synchro generator is concerned, by the use of a synchro capacitor. This is a set of three capacitors usually connected in delta and connected across the lines connecting the synchro generator and the

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differential generator. This permits more control transformers to be supplied by a given generator. A capacitor across each of the stator leads of a control transformer also tends to reduce the carrier wave phase shift between the generator excitation and the control transformer output.

Synchro capacitors are also sometimes used in simple generator control transformer systems, particularly when one generator supplies several control transformers. Since the impedance of the control transformer is usually quite high, a smaller capacitor is required in this application. The fundamental limitation on the number of control transformers that can be excited from a single generator, however, is usually the inaccuracy resulting, particularly if all the control transformers are not in synchronism.

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CHAPTER 4

FURTHER SYNCHRO PROPERTIES AND USES

Synchro Phase Shift

As mentioned previously, one problem in connection with the application of synchros is the presence of a phase shift between the excitation voltage and the output voltage of the control transformer, even if the residual voltages are zero. This phase is caused by the reactance of the synchro windings and is in such a direction that the control transformer voltage lags the excitation voltage to the generator. This must be taken into account in adjusting the phase of the voltage applied to the fixed field of a two phase motor and also the reference voltage of a phase-sensitive detector. This phase shift is usually around 3° to 5° for small units but may run higher than this for very large synchros.

Frequency Specifications

It is worth noting that 400 cycle control units, as with any transformer, can be made smaller than corresponding 60 cycle types. In addition the dynamic errors mentioned previously are reduced by a factor of about seven.

Power Synchros

It is possible to produce synchronous action between two mechanical systems by use of power synchros or the synchro-tie. This system makes use of two identical wound-rotor induction motors operating with both stators paralled and both rotors paralled. When this system is once synchronized,

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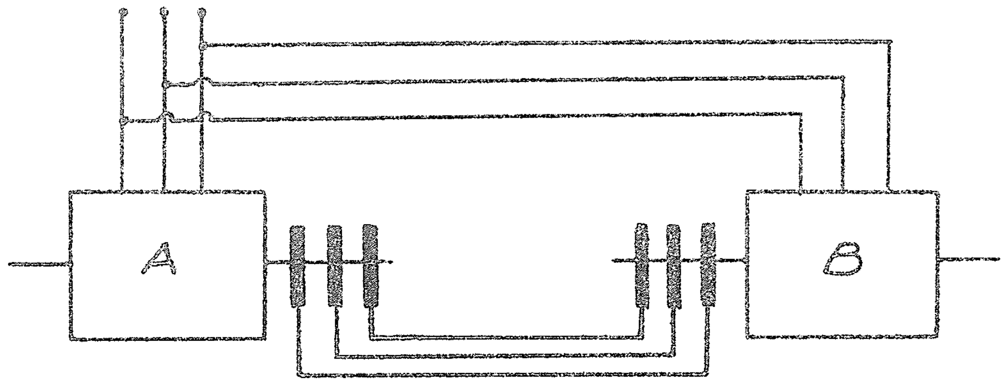
the shafts of the two machines are in synchronism even though there is no mechanical connection between them and this requires that they turn at the same speed. Even under these conditions, however, the space angle between the stator flux and rotor flux of one machine may differ from that of the other machine. The fluxes produced by both machines can be considered equal and the rotor voltages for a given slip must be equal in magnitude. The currents in the two rotors are identical.

The electrical circuit of one phase of the rotor circuit is shown in fig. 14. When the machines are identical and operating in synchronism, the rotor voltages induced in corresponding phases of machines A and B are E_A and E_B. The impedances of the rotors are $(R_a + jX_a)$ and $(R_b + jX_b)$. When the loads on the two motors are equal, they must develop equal torques. The torque developed in either machine is equal to a constant multiplied by the product of its rotor voltage, its rotor current, and the cosine of the angle θ_a or θ_b between I_2 and E_a or E_b. That is:

$$\text{Torque in A} = K \cdot E_a \cdot I_2 \cdot \cos \theta_a$$

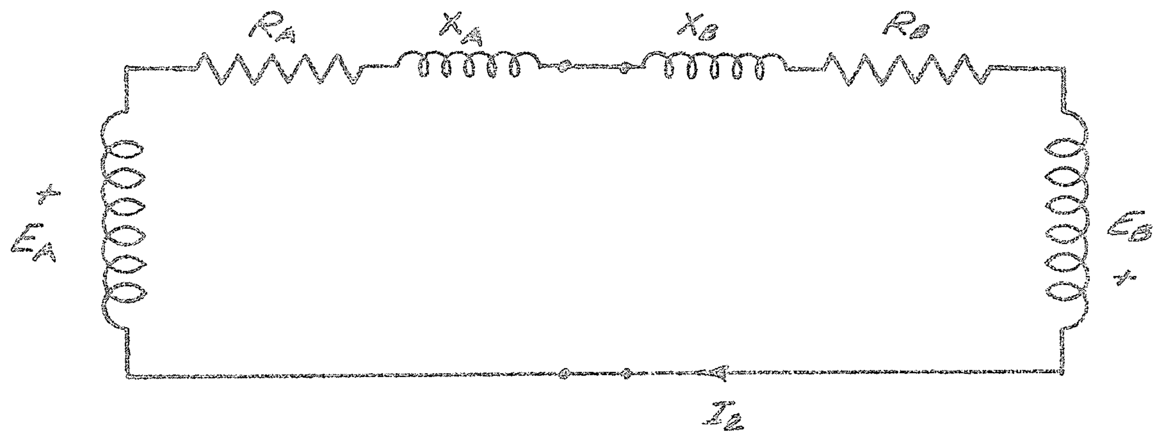
$$\text{Torque in B} = K \cdot E_b \cdot I_2 \cdot \cos \theta_b$$

Unbalanced loads will therefore require a difference in the value expressed by these equations. Since $E_a = E_b$ and I_2 is common to both, this difference must result from different rotor power factor angles θ_a and θ_b. Consider for instance,



POWER SYNCHRO CONNECTION

fig 13



EQUIVALENT CIRCUIT FOR POWER SYNCHRO CONNECTION

fig 14.

that the load on shaft B were to reduce to zero. The torque developed in machine B would then cause acceleration of its rotor. When it has gained position such that the rotor conductors have moved sufficiently forward, the resultant forces on the rotor of B will be zero, regardless of the magnitude of the rotor current. If the rotor of machine B was to advance from this position, it would experience forces that would tend to restore it to the original position. Thus, for different loads on the two motors, their rotors can occupy different space-phase positions but they must run in synchronism.

The system has one undesirable feature. If the two rotors come to rest, when the power is shut off, such that E_a and E_b are 180° out of time phase, no resultant current I_a will flow in the secondary circuits and consequently there will be no starting torque when reconnected to the voltage supply. This condition can be overcome by switching circuits inserted in the primary side of one machine.

This type of system, although not used extensively, has been applied successfully on several occasions. One example is the operation of two motors on opposite ends of an elevating bridge. It is not adaptable for installations requiring large values of torque.

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Other Uses of Synchros

The types of synchro systems discussed up to this point do not cover all modifications and applications of synchros. For example a receiver similar to the differential generator may be built with triple stator windings. This permits applying to the stator, as well as to the rotor, signals which correspond with the angular displacements of two different shafts. The resulting rotor position of this differential receiver is then proportional to the sine of the difference between the angles of the other two positions. Another possibility is to obtain a primary feedback signal from small angular deviations, using for this purpose the synchro generator without any other synchros. Fig. 5 (page 14) showed that for limited deviations from zero, the voltage change is practically linear with change in angle.

Since these remaining applications of synchros are mostly just modifications and glorifications of the basic systems already discussed, an outline of their operation shall not be included.

C H A P T E R 5

S Y N C H R O E R R O R S

Synchro Errors

Synchros, when used in connection with servomechanisms, are not as perfect as is intimated in the foregoing description of their operation. Actually, synchros suffer from two major faults:- inaccuracy and residual voltages. The servo can be made no more accurate than its error detector and if the synchco does not accurately indicate the difference between two shaft angles, then the servo will be in error by at least this inaccuracy. If this voltage null does not occur when the command shaft has precisely the same angle as the controlled shaft, then the servo will be in error and there is nothing to indicate to the amplifier and motor that such an error exists.

Although synchros can be built having a static error of but a few minutes, the typical error of a synchro generator control transformer combination is on the order of about 18 minutes maximum and is a function of rotor position. One way of determining the error of a synchro-pair is first to affix dials to the shaft of the generator and the control transformer. Suppose then that the shafts are brought into correspondence so that the voltage output is a null at electrical zero. If the shaft of the generator is rotated 10°, in general the shaft of the control transformer must be rotated an angle of 10°, to achieve a null once more. The difference between the angle to which the control transformer must be rotated, and 10°, is the error of the synchro-pair. If this experiment is repeated throughout 360°, a typical graph of the error as a function of shaft angle is obtained as illustrated in fig. 15.

If it is desired to find the error merely of a synchro generator, this can be done by defining zero error as say electrical zero, affixing a very accurate dial to the shaft, and measuring the stator voltages induced for any angular position of the shaft. Since these stator voltages define an angle, the angle corresponding to the measured voltages can be calculated. If this calculated angle differs from the measured angle, the difference is the error. Similarly, voltages having the exact proper relative magnitudes for a given angle can be fed to a synchro control transformer and the shaft angle necessary to cause a null measured. If this shaft angle differs from the angle corresponding to the applied voltages, an error exists and is equal to this difference.

Provided that the synchro has been properly designed, the cause of synchro errors is generally the result of manufacturing difficulties. If the stator coils of a generator for example are not identical, errors can result. These discrepencies in the coils can be caused by different numbers of turns on the coils, or by different wire resistances caused by using wire from different manufacturers or even from the same manufacturer but from different spools. For salient-pole machines, the length and shape of the poles are important. In

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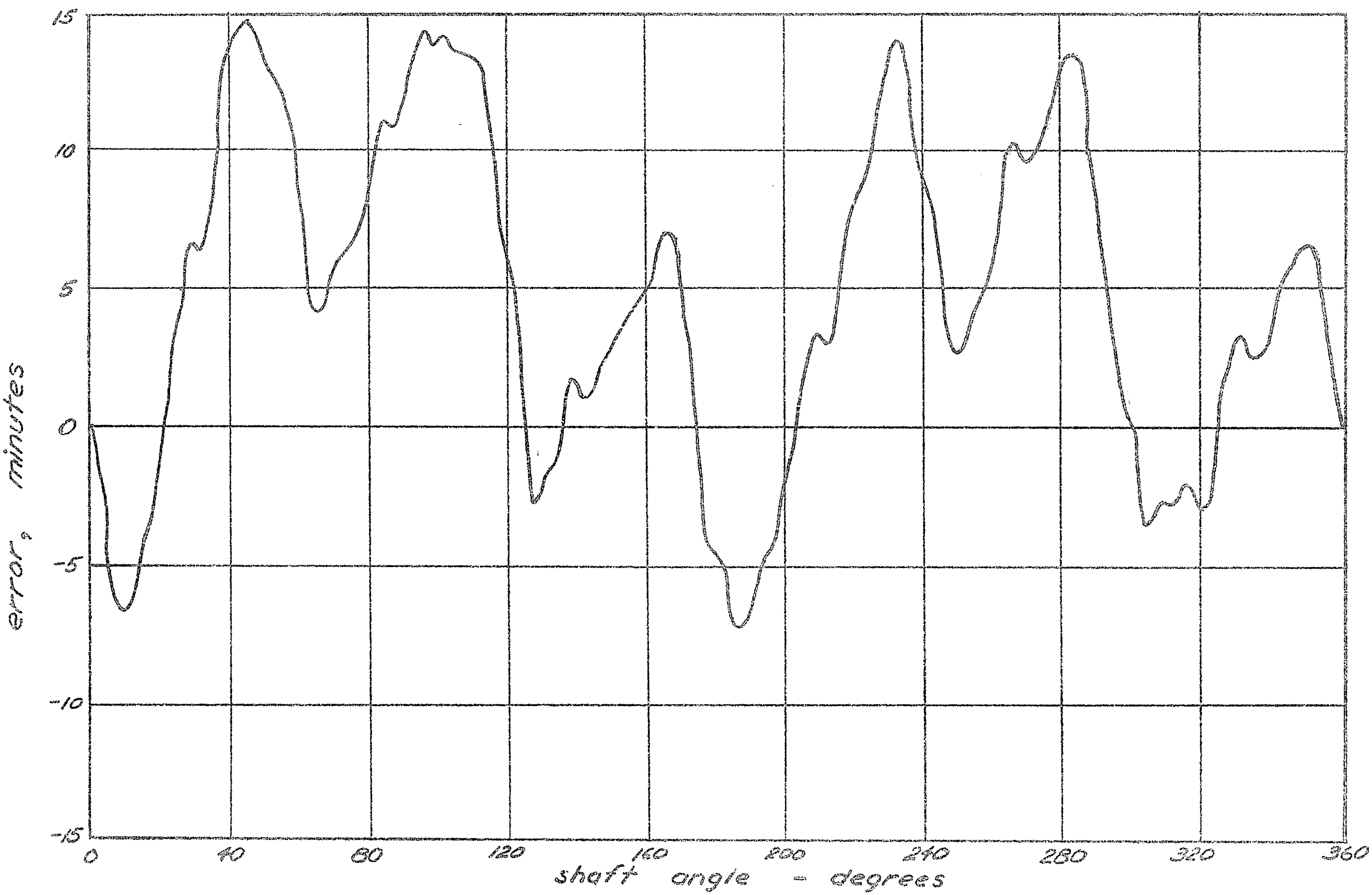

ERROR OF A SYNCHRO PAIR

fig. 15

machines with round rotors, the rotor must be exactly round. This is not too difficult to do, but it is difficult to grind the stator of the synchro so that it is exactly round. Ellipsing of either the stator or the rotor causes errors. For example, ellipsing of either the rotor of the stator an amount such that the difference between the major and minor axes of the ellipse is as small as one ten-thousandth of an inch, results in a spread of four minutes error in a typical small synchro.

The rotor must be placed exactly in the center of the bore of the stator and the laminations must be of uniform magnetic property so that even the direction of rolling of the steel sheet has to be taken into account. For this reason, such laminations are skewed when stacked. The slots are also skewed, so that the reluctance of the slots is evenly distributed lengthwise over the magnetic path. The angle of skew is carefully designed. The length of rotor laminations compared with the stator laminations is also important. To control this factor the laminations are usually cemented together to eliminate the flare that might exist if they were merely stacked and bound mechanically.

The static synchro error can be resolved into various space harmonics. The largest error is usually the second harmonic, that is, the error goes through two complete cycles for every 360° of shaft rotation. This is caused mainly by ellipsing of the stator. If the rotor was ellipsed, the error could also be fundamental. Another common space harmonic is the sixth harmonic, caused by the "three-phase" winding distribution. Another harmonic present is equal to the number of slots in the rotor or stator or is double the number of slots and is caused by the non-sinusoidal flux distribution as a result of the reluctance of the slots or of the varying resistance or reluctance of the coils themselves.

The accuracy of synchros is therefore fundamentally a problem of how accurately parts can be manufactured.

Dynamic Errors

Although the static error is the error of principal concern, there is also a dynamic error present in synchros. This is caused by the fact that the magnetic fluxes existing in synchros are being cut by the wire of the rotor coils. If a synchro generator and control transformer were coupled together at null position, and the synchro generator excited from a source of alternating voltage in the usual way, the output of the control transformer would be zero if there were no static error or residual voltage present in the units. This would be true for any angular position of the coupled shafts. If however, they were rotated at a constant velocity and their stators held fixed, a voltage would appear across the rotor of the control transformer. At about 300 r.p.m. rotation of the shaft for typical 60 cycle units, the voltage would be about 1 volt. If the shafts had been displaced by 1 volt so that when they were standing still there would be about 1 volt output from the control transformer and again the shafts were rotated at 300 r.p.m., the output voltage would be ~~null~~ zero. Thus servos using synchros have an additional error when the shafts are rotating at high speed, which is roughly proportional to the speed of the shaft rotation. The phase of this voltage also shifts slightly from the phase of the synchro output voltage when the shafts are standing still. This direction of the dynamic synchro error for constant velocity of its shaft is in the direction to increase the steady-state error of a servomechanism when following a constant velicity command.

Residual Voltages

The other principal defect of synchros is the presence of residual voltages. It has previously been stated that when the shafts of the synchro pair were in correspondence, the voltage output of the control transformer was ~~at null~~ zero. This is not entirely true however due to an effect known as Residual Voltage. It is found that as the rotors of the two synchros approach correspondence, the output voltage does not

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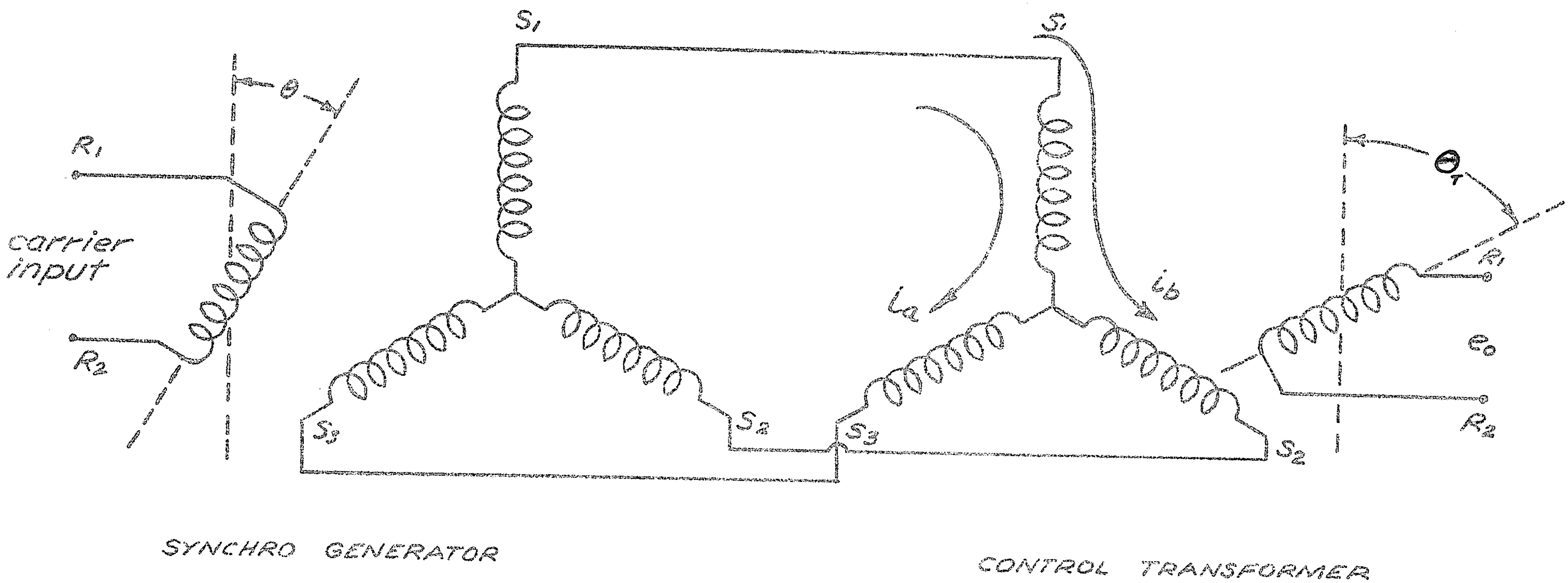
go to zero, but instead only goes to a minimum. This residual voltage remaining at the minimum consists primarily of a component at carrier frequency ω_c but 90° out of phase with the predominant phase of the output signal for large differences in rotor displacement. It is therefore referred to as the quadrature voltage or quadrature error.

It can be shown that one reason for the existence of the quadrature error is that the impedances making up the stator circuit do not all have the same phase angle. Consider the system shown in fig. 16. The two currents i_a and i_b are alternating at the carrier frequency ω_c. Their magnitude depends on θ, the rotor angle of the generator, and although ideally they are in time phase, in practice a small difference in the phase angles of the impedances of the stator results in a small phase angle β between the currents. Thus let:

$$i_a = I_a \sin \omega_c \cdot t$$

$$i_b = I_b \sin (\omega_c \cdot t - \beta)$$

where I_a and I_b are the maximum values of i_a and i_b respectively. The control transformer output voltage is produced by the mutual inductance between the control transformer stator and rotor windings. This mutual inductance is a periodic function of θ_T the rotor angle. For the usual cylindrical rotor construction this inductance may be assumed to be a sinusoidal function of θ_T. If we let M_c be the



BASIC SYNCHRO PAIR

Fig. 16

maximum value of this mutual inductance then the output voltage may be written:

$$e_o = M_c \cdot \cos \theta_T \cdot \frac{di_a}{dt} + M_c \cdot \cos \theta_T \cdot \frac{di_b}{dt} +$$

$$M_c \cdot \cos (240^\circ - \theta_T) \cdot \frac{di_a}{dt} + M_c \cdot \cos (120^\circ - \theta_T) \cdot \frac{di_a}{dt}$$

$$= M_c \cdot \cos (\theta_T + 60^\circ) \cdot \frac{di_a}{dt} + M_c \cos (\theta_T - 60^\circ) \cdot \frac{di_b}{dt}$$

substituting for i_a and i_b:

$$e_o = \omega_c \cdot M_c \left[I_a \cos (\theta_T + 60^\circ) \cos \omega_c \cdot t + I_b \cdot \cos (\theta_T - 60^\circ) \cdot \cos (\omega_c \cdot t - \beta) \right]$$

By expanding the $\cos (\omega_c \cdot t - \beta)$ term this becomes:

$$e_o = \omega_c \cdot M_c \left[\left\{ I_a \cos (\theta_T + 60^\circ) + I_b \cos \beta \cdot \cos (\theta_T - 60^\circ) \right\} \cos \omega_c \cdot t + \left\{ I_b \sin \beta \cdot \cos (\theta_T - 60^\circ) \right\} \sin \omega_c \cdot t \right]$$

This equation indicates that the rms control transformer output voltage cannot go to zero for any value of θ_T unless I_b is zero, a condition occuring for only two positions of the generator rotor. Thus it is seen that there will always be some voltage output, even when the rotors are in correspondence, unless the generator rotor happens to be in the two positions which make $I_b = 0$. The second term in the equation is the quadrature error referred to above, and for small β, it is approximately proportional to the phase difference between the stator currents. It is clear that this voltage can be kept to a minimum by making certain that the impedances of the stator circuit all have the same phase angle. Usually synchros are wound quite carefully and the windings are almost identical, but unless the connecting cable is perfectly constructed, it may introduce unbalancing impedances that will give rise to considerable error.

Errors Due to Harmonic Voltages

In addition to the effects described above, harmonic voltage components are also found near the synchro nulls. Some of these are due to harmonics present in the supply voltage, but most of them are due to non-linearities in the iron. Quadrature voltages of fundamental frequency are also sometimes caused by unbalanced capacitive coupling between stator and rotor, although this effect is usually not too important in low frequency transducers like synchros.

The phase discriminator or demodulator, which in one form or another is always used to recover the actual error information from the synchro signal, can, by proper adjustment, be made insensitive to all quadrature and harmonic voltages and hence these residual effects do not result in any appreciable servo error. If it is desired to amplify the ~~desired~~ synchro error signal in an a.c. amplifier, this amplifier must be designed to handle input amplitudes that are at least as large as the maximum quadrature signal, and preferably somewhat larger. Hence a large amount of quadrature voltage limits the amount of a.c. gain that can be used ahead of the discriminator. If more gain is needed, a d.c. amplifier can be used after the discriminator. Since d.c. amplifiers tend to drift, it is usually desirable to have as much of the required gain as possible in the a.c. amplifier. For this reason it is important to reduce the quadrature error to a minimum. Another disadvantage of a large quadrature signal is that it results in a considerably larger ripple output from the discriminator and therefore requires more filtering.

The presence of harmonic voltages in the excitation voltage applied to the synchros adds somewhat to the harmonic voltage output of the control transformer, but even a pure voltage source will not eliminate the harmonic voltages produced in the synchro themselves.

CHAPTER 6

DOUBLE SPEED SYSTEMS

Double Speed Systems

It was noted in the last chapter that synchros suffered from two basic limitations, inaccuracy and the presence of noise voltages at null. The first of these problems can be greatly helped by the use of double-speed transmission. The terminology "double-speed transmission" is a confusing but conventional term which means that the position of a shaft is transmitted electrically by two synchros that rotate at different speeds as the command shaft is rotated. Such a scheme also reduces the effect of noise voltages, however it introduces another problem requiring the use of the synchronizing network. Before considering double-speed systems though, a preliminary consideration, error reduction using gearing, must be regarded.

Error Reduction Using Gearing

The basic problem in improving synchros is improving their accuracy. With the previously described application of synchros coupled directly to the command shaft and the controlled shaft, any inaccuracy of the synchros becomes directly an inaccuracy of the servomechanism. Suppose that, instead of direct coupling, the synchro generator and the synchro control transformer are each geared to their respective shafts by a ratio of 5:1. Assume that angles v and c, which define the angular positions of the command and controlled shafts

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respectively, are made equal to zero and the synchros are adjusted by positioning the stator of the control transformer so that the final output signal is zero. Suppose now that the command shaft is moved through 10°. This means that the generator shaft rotates 50° if the gearing is accurate. The command shaft is then rotated by hand until a null voltage exists at the output of the control transformer. If the synchros were perfect, this angle would be precisely 10° and the control transformer would have to be rotated through 50°. Suppose, however, that the synchro pair has an error of 10 minutes at this point. This would mean that the shaft of the control transformer would still have to be rotated through 10 minutes to effect a null. This would be accomplished by rotating the angle c through only 2 minutes. The absolute error therefore has been cut by a factor of 5, which is equal to the gear ratio between the command shafts and the synchros.

If the voltage from the control transformer were fed to a servo-amplifier, which in turn fed a servomotor geared to the controlled shaft, it is seen that the error introduced by the synchro transmission system would have been cut by a factor of 5 since the servo operates to make the actuating signal equal to zero. In a similar manner, a gear ratio of say, 25:1, or almost any other gear ratio, could have been used. Since the synchro rotates at a higher speed for a given controlled shaft speed, the synchro dynamic error at

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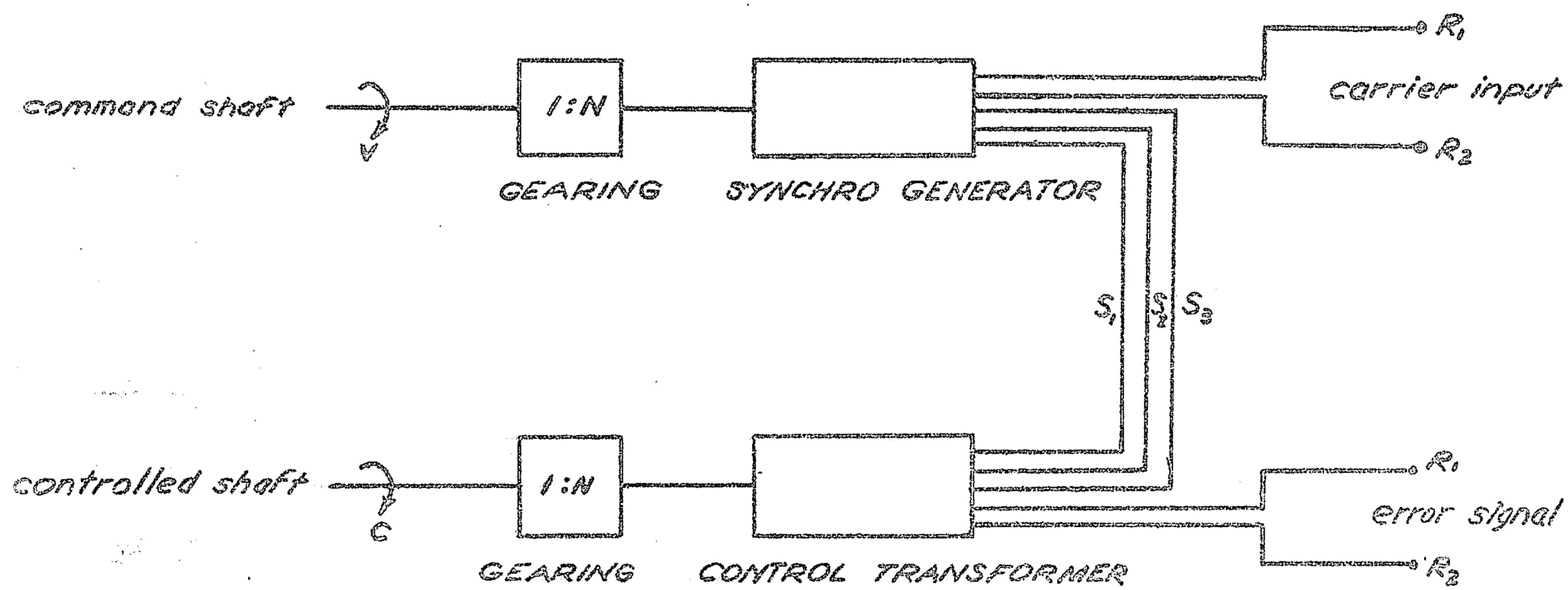
the synchro shaft would be increased by a factor of the gear ratio, the synchro dynamic error referred to the controlled shaft would remain unchanged. The inertia of the synchro reflected to the servomotor however, would increase under the use of geared-up synchros.

It is also noted that the voltage sensitivity of the synchro system has been increased by the factor of the gear ratio. A motion of 1° of the command shaft for instance causes a voltage of 5 volts to appear at the control transformer output leads if the synchros have a sensitivity of 1 volt/degree. This means that the servo amplifier used can have one-fifth the gain that would be necessary if one speed transmission were used. The problem of pickup in the input leads of the amplifier has therefore been reduced.

A difficulty however has been introduced by the use of the scheme shown in fig. 17. Suppose the command shaft is clamped so that it cannot move and the controlled shaft is moved through an angle of 72°, or 1/5 of a revolution. The synchro control transformer has then rotated through a complete revolution, and a null voltage again appears in the output of the control transformer. The shafts v and c, however, are now 72° apart. Thus, if the system were used in a servomechanism, the servo would drive to the nearest of the five null positions, only one of which would bring shafts



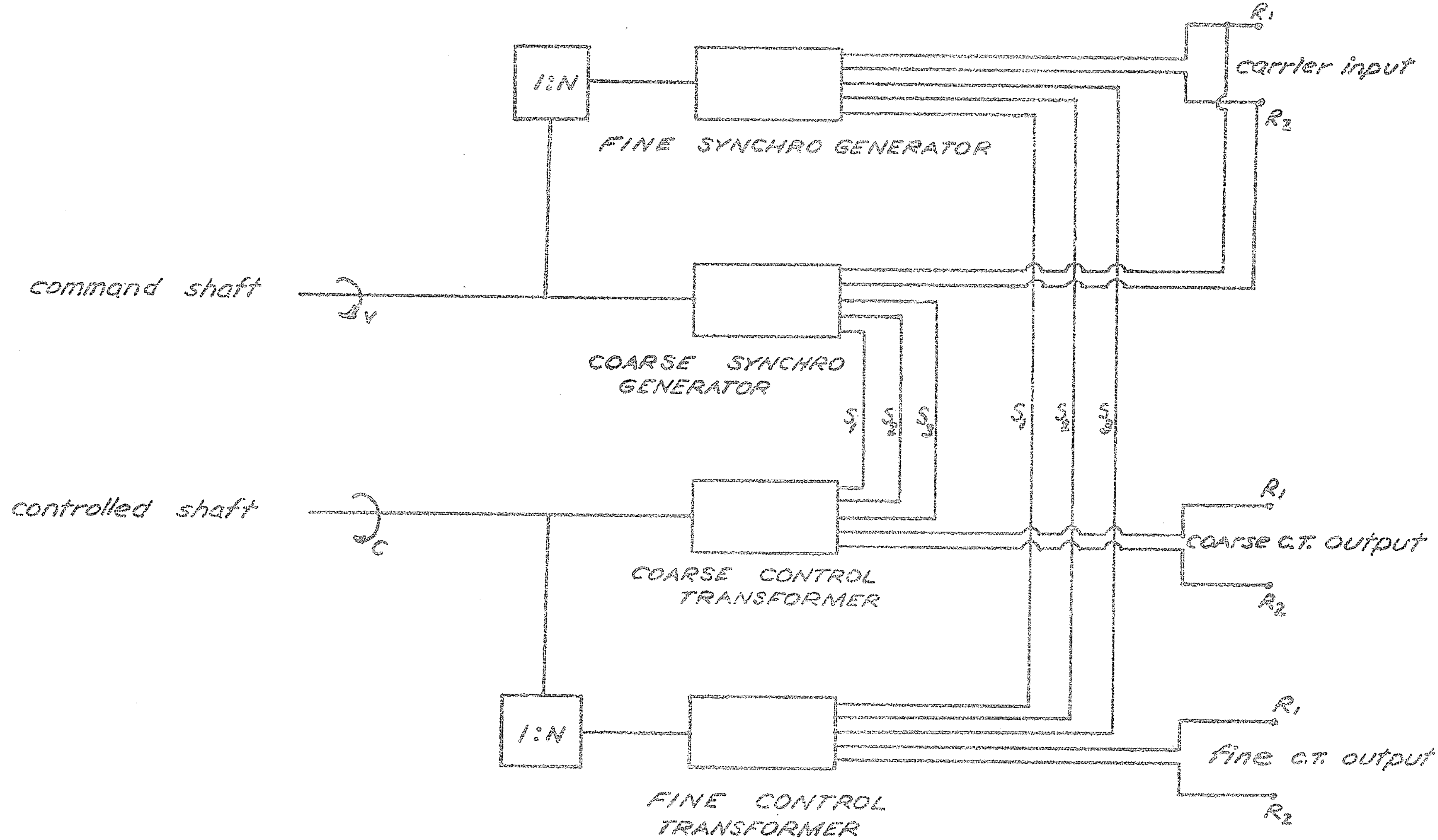
SYNCHROS GEARED 1:N to COMMAND SHAFT and CONTROLLED SHAFT

fig. 17

v and c into correspondence. It is necessary therefore, that additional information be fed into the servomechanism to tell it which of the five null positions is the correct one.

Basic Double-Speed System

This system can be set up according to the scheme shown in fig. 18. Double-speed synchro transmission required two synchro generators and two control transformers. The generator and control transformer are geared n:1 to the command shaft and the controlled shaft respectively, but in addition, a generator and a control transformer are coupled directly to the two shafts. The synchro system can be put into proper adjustment by placing the command shaft angle v at zero and the controlled shaft angle c at zero. The stators of each of the control transformers are then rotated to the positions at which the output of both the one-speed and the five-speed control transformers are equal to zero. If the command shaft is moved to any arbitrary angle, the controlled shaft can be rotated an equal amount by observing the output voltage of both the control transformers. This is accomplished by rotating the ~~command~~ controlled shaft and observing first only the voltage from the coarse, or single speed control transformer. When this voltage is zero the shafts are close to correspondence, though perhaps not exactly so because of errors in the

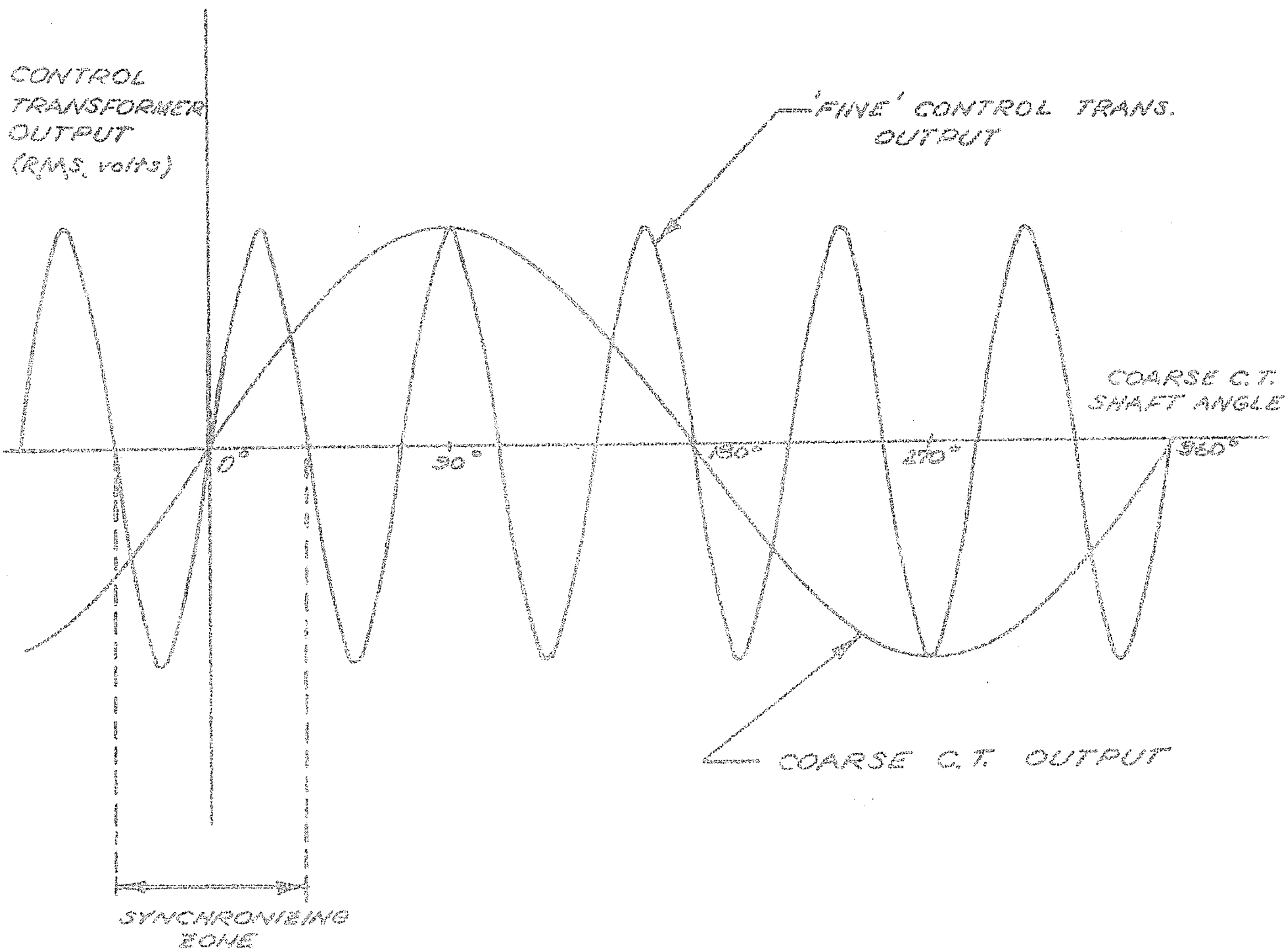


SYNCHRO DOUBLE-SPEED TRANSMISSION SYSTEM

fig. 18

synchros. If, however, the ~~command~~ controlled shaft is rotated through the small angle necessary to cause the fine control transformer output voltage to be equal to zero, it will be more exactly in correspondence with the command shaft. This is true not only because the sensitivity of this control transformer per degree of controlled shaft angle is higher and thus gives a greater signal voltage for a given lack of correspondence but also because the error in synchro transmission has been reduced by a factor of n. (When the fine control transformer voltage is at zero, the coarse control transformer voltage will be slightly different from zero because of error in the synchros). This system is similar to the hour and minute hands of a clock which constitute a form of double-speed system. The hour hand tells the approximate time, and the minute hand gives a more accurate indication. The minute hand alone though does not give sufficient information unless one knows the hour.

It is instructive to look at the voltages that are developed in the various synchros. Suppose that the command shaft is in any arbitrary position and that the controlled shaft is rotated until they are in correspondence. If the command shaft is held fixed, and the controlled shaft is rotated, the magnitude of the output voltages of each of the control transformers will be as shown in fig. 19 for one revolution of shaft c. It is seen that, in one revolution



DOUBLE-SPEED SYSTEM OUTPUT VOLTAGES

Fig 19.

of the shaft c, the fine control transformer voltage has a null of the proper polarity five times but the null of the coarse control transformer voltage has only one null of the proper polarity. If a servomechanism were used to operate with the synchro transmission system of fig. 18, it would be necessary to have some sort of circuit which would cause the servo amplifier to be fed from the coarse control transformer voltage when the actuating signal was large, and from the fine control transformer voltage when the actuating signal was small.

The simplest device that will do this is a relay whose coil is fed from the coarse control transformer voltage. If the coarse control transformer voltage is large, the relay picks up and its contacts feed the coarse control transformer voltage into the servo amplifier. When the coarse control transformer voltage becomes smaller, indicating that the servo has rotated the controlled shaft close to its correspondence position, the relay drops out and the contacts feed the fine control transformer voltage into the servo amplifier. The servo-controlled shaft is then positioned only by the output voltage of the fine control transformer and the accuracy of the transmission system is improved by the factor of gear ratio. When used in this application, the relay is known as a synchronizing network and is one of many types of devices that can be employed for this purpose.

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Fundamentals of Operation of the Synchronizing Network

It would appear that a problem would arise from the fact that the sensitivity of the synchros, and hence of the servo, changes by a factor of the gear ratio when the relay switches from fine to coarse. No difficulty exists however, since the servo amplifier is usually saturated by a very small actuating signal. Once the amplifier has saturated, its output voltage is approximately constant, irrespective of the magnitude of its input signal. At the point of transfer from the fine control transformer voltage to the coarse control transformer voltage, the amplifier is usually saturated by either signal, and hence no change of torque or of loop gain occurs. The servo is therefore designed for the fine synchro gain, since it will normally be operating with the fine control transformer supplying the actuating signal.

It will be seen from fig. 19 that there are two positions at which the coarse control transformer voltage is close to zero, indicating that the relay would drop out and feed the fine control transformer voltage to the servo amplifier. One position is that of correspondence where $c = v$, and the other is a position of c which is 180° away from its proper position. While it is true that the relay would drop out and feed fine control transformer voltage to the servo at the 180° position, the polarity of the fine control transformer voltage is the same as the polarity of the coarse control transformer voltage and the servo tends to drive the controlled shaft away from this point toward correspondence, so that no false null is introduced.

During normal application of a servomechanism, employing double speed synchro transmission, the servo is always operating on the fine actuating signal whenever the controlled shaft is expected to follow the command shaft closely. The only time it is not expected to follow closely is when the servo is de-energized or when the command shaft is moved so swiftly that the controlled shaft cannot follow, that is, when the command shaft is rotated at a velocity faster than the controlled shaft can rotate even with the motor travelling at full speed. It is only under these conditions that the synchronizing network performs its function of switching.

Gear ratios between fine and coarse synchros in common use are 25:1, 16:1, 36:1, and 20:1. As long as the gear ratio is odd, no additional problem is introduced. Problems caused by the use of even gear ratios are discussed in the reference given below*.

The span of the zone in which transfer from coarse to fine must take place is always equal to one revolution of the fine control transformer. If a gear ratio of 75:1 were used, the total width of this zone would be 1/75 revolution, or 4.8°. With this 75:1 gear ratio and using the synchronizing

* Ahrendt - Servomechanisms Practice - page 64
"Problems involved in using even gear-ratios"

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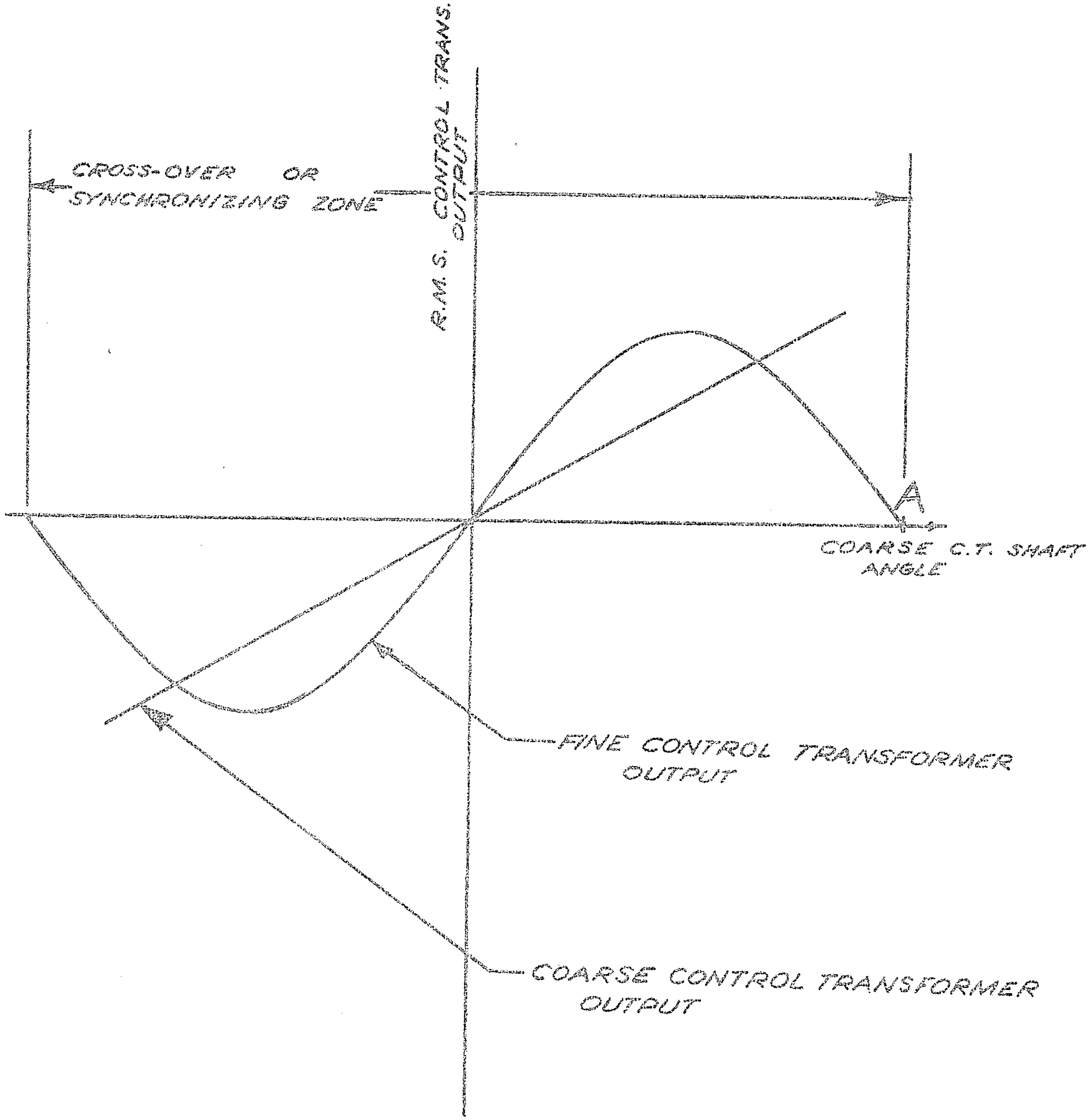
circuit described, the relay must transfer from coarse to fine at a coarse synchro voltage less than 2.4 volts, assuming a synchro gain of 1 volt/degree, as the servo error decreases. If the servo error is increasing, the relay must also transfer from fine to coarse at a coarse synchro voltage less than 2.4 volts. Suppose for example, that the servo has been deenergized and the controlled shaft is 5° away from correspondence. Suppose also that the relay is not properly adjusted and drops out at 6 volts output of the coarse control transformer. If the servo is re-energized, the relay will not pick up. The servo will drive toward a null position 4.8° away from the correct one and the servo will be in error by 4.8°.

In every circuit there is the practical problem of selecting the voltage at which the transfer shall take place. Actually, the transfer can occur at any place within the synchronizing zone, however, practical circuit problems place some restrictions on the transfer point normally chosen. For example, the transfer point could be at a position very slightly to the left of the position labeled A in fig. 20, where the position of A in degrees rotation of the coarse control transformer is given by:

$$\frac{1}{2}\left(\frac{360^{\circ}}{N}\right)$$

where N is the gear ratio between the coarse and fine control transformers. This would mean that when the servo is coming into synchronism the voltage being fed to the servo amplifier



SYNCHRONIZING ZONE for DOUBLE-SPEED SYSTEM

Fig. 20

would switch from the coarse to the fine voltage slightly to the left of point A. The servo would then operate on fine voltage and would drive to the zero position. At the instant of transfer, the actuating signal voltage would drop to a very low value and, while it would still drive the servo to the proper null position, might show some hesitation in doing so.

Another difficulty is that a slight shift in parameters, a shift in circuit value, or a shift in the relay characteristics might cause the point of transfer to shift to the right of A. This would result in the servo and the relay oscillating. Furthermore, if an error was present in the fine synchro it would mean that point A would not occur at exactly 180° rotation of the fine synchro away from a zero position. If the fine synchro were in error in such a direction that the fine control transformer voltage reached zero for an angle slightly less than A, the servo would again oscillate. For these reasons, a safe value of the transfer point would be about 90° of fine control transformer rotation away from zero at the point where the voltage output of the fine control transformer is a maximum. This position minimizes the possibility of the servo system oscillating due to a shift in parameters while still holding the transfer point far enough away from the null position that the coarse voltage would still saturate the amplifier.

From this analysis it is seen that, to have some margin of safety for relay differential and relay adjustment, the synchronizing zone cannot be too small and hence the gear ratio cannot be raised indefinitely. There is also a limitation in gear ratio caused by the inaccuracy of synchros, for example, if the coarse synchro error in a 75:1 gear ratio system were 1.5° at some angular position of v, and the relay was adjusted to transfer from fine to coarse at angle of 1.5° or less, then the relay would improperly transfer the amplifier input from fine to coarse even though the servo was in exact correspondence. The relay would again cause the servo to oscillate.

It must be remembered in these explanations that the actuating signal voltage from the control transformers will be zero for any angle from zero to 360° of the command and controlled shafts as long as they are at the same angle, that is in correspondence. It is therefore necessary to be concerned with synchro errors for any position of the synchro shaft. In choosing an angle for the transfer point, it is therefore necessary to choose one which is greater in magnitude than the largest angular error of the coarse control transformer.

Practical Synchronizer Circuitry

Since the experimental work connected with this dissertation will not delve into the realm of double-speed systems, and since many excellent reference texts* on this subject are

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* Ahrendt - Servomechanisms Practice
Gibson & Tuteur - Control System Components

*

available, further material on double-speed systems and their associated circuitry ~~shall~~ will not be included here.

CHAPTER 7

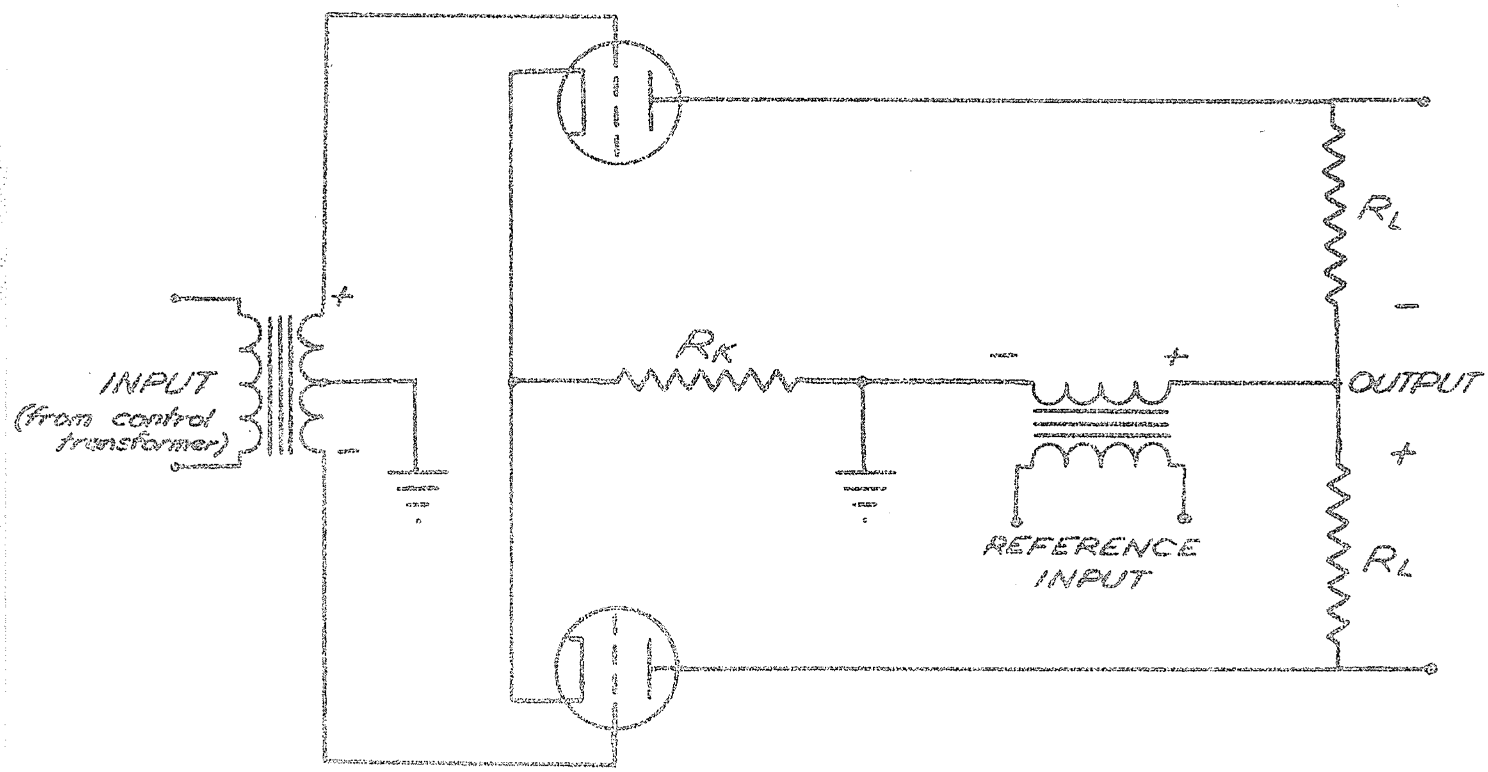
DISCRIMINATORS

Discriminators

To make the output voltage from the control-transformer suitable for use with many servomechanism systems it must be converted into a direct-current signal or a low frequency a.c. signal. The magnitude of this voltage must be directly proportional to the magnitude of the a.c. voltage from the control-transformer, and its polarity must be dependent on the phase of the control transformer signal. Such devices are commonly known as discriminators or demodulators but are actually phase-sensitive detectors or phase-sensitive rectifiers.

Basic Action of the Phase-Sensitive Discriminator

A simple circuit to accomplish the type of demodulation required is shown in fig. 21. This circuit is essentially a push-pull amplifier of standard design with plate load R_L and cathode-biasing resistor R_K. The feature that makes the circuit act as a phase sensitive discriminator is the a.c. plate supply voltage. This supply is referred to as the reference voltage, and it must be of the same frequency as the input signal. As a result of the a.c. supply, the amplifier is cut off every other half cycle, operating only during the half cycle in which the plate voltage is positive. If the phase of the input signal is such that during the ON period of the amplifier the upper grid (fig. 21) is positive and the lower one negative, more current flows in the upper triode than in



HALF-WAVE TRIODE DISCRIMINATOR

fig. 21

the lower, and hence the plate of the upper triode is more negative than that of the lower. If the phase of the input is reversed, the polarity of the output reverses. Typical output waveshapes obtained from the circuit for various phase relations between input and reference are shown in fig. 22. In drawing these figures it is assumed that the gain of the tubes is independent of the magnitude of the plate voltage. This is usually a good assumption during most of the time that the voltage is positive. Fig. 22(c) shows the effect of an input signal 90° out of phase with the reference. Here the average d.c. output is zero. The discriminator is therefore seen to reject quadrature signals. In general, if the input is less than 90° out of phase with the reference, the d.c. component of the output is reduced from the value it would have if the signal and reference were exactly in phase.

In all cases shown, the output contains a large a.c. ripple component. This ripple is normally highly objectionable because it results in saturation, heating, and general malfunctioning of amplifiers and power devices following the discriminator. Low-pass filters are therefore usually used in the discriminator output to attenuate this ripple. For most applications these are simple R-C networks.

The output waveforms shown in fig. 22(a) and 22(b) are similar to those obtained from a standard half wave rectifier,

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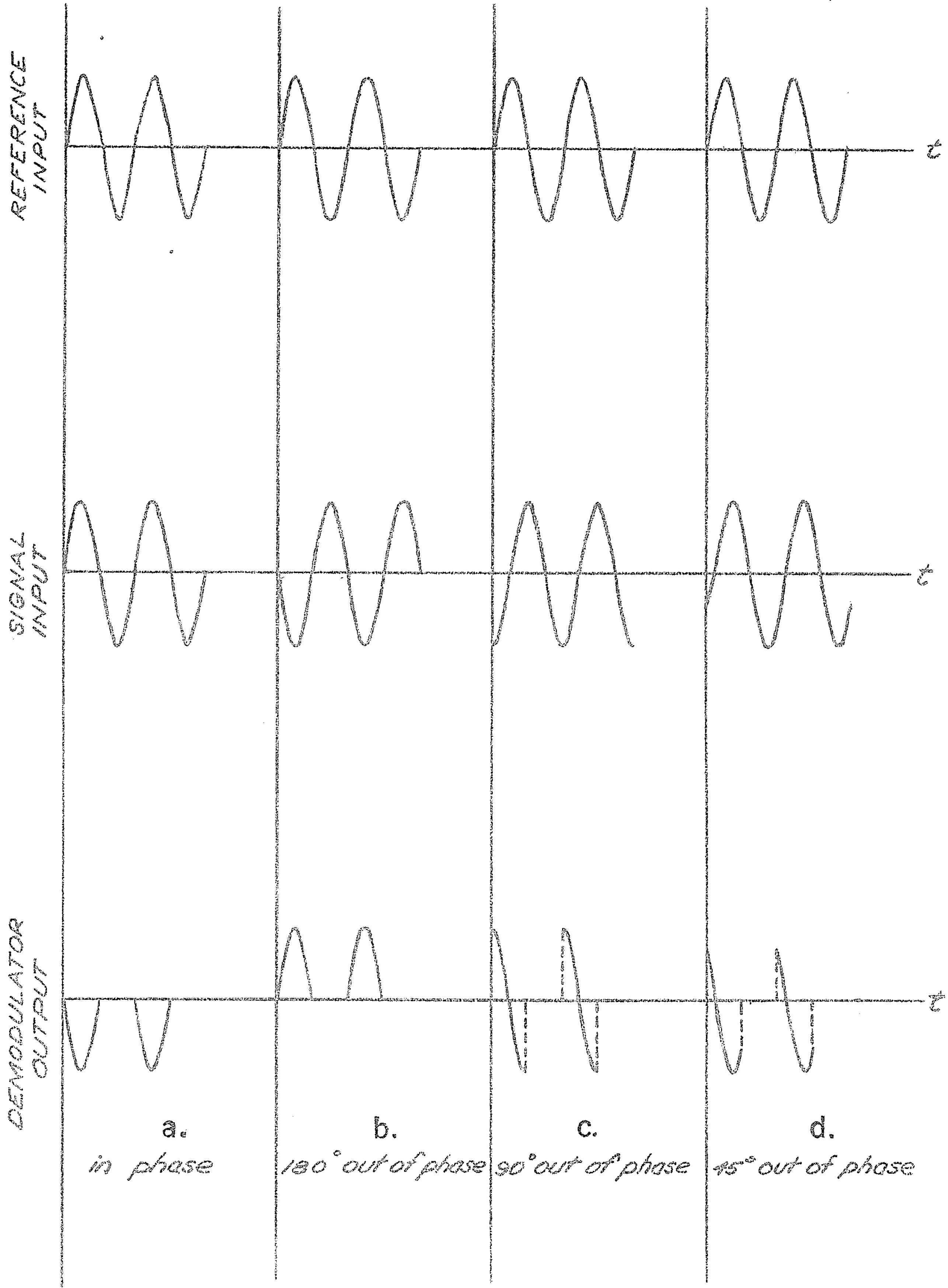


DISCRIMINATOR WAVESHAPES

Fig. 22

hence this discriminator circuit is referred to as a half-wave discriminator. Some of its other features are that it provides gain between input and output and that it delivers a push-pull output signal. If the two halves of the circuit are exactly identical, zero input results in both zero d.c. and zero ripple output. In practice, however, perfect balance of the circuit is difficult to achieve, and there is usually some output for zero input. Since the circuit unbalance may vary as a result of temperature and supply voltage changes, the output produced by the unbalance may also vary, that is, the circuit may drift. This is not unreasonable since the circuit is in part a d.c. amplifier, but it should be noted that since it is a push-pull circuit the drift should be quite small.

Analysis of the Half-Wave Discriminator

In analyzing the operation of the discriminator circuit considered in the preceding paragraphs, it is convenient to assume that the output voltage delivered during the ON period of the amplifier is equal only to the amplified input signal and is independent of the plate voltage. While this is not quite true, it is approximately correct during the major part of the ON period if the plate characteristics of the tube in question are fairly linear, that is, if r_p and μ are approximately independent of the operating point. Also it will be found that the exact form of the output signal will affect the analysis only in detail and not in principle.

If the gain during the ON period = G, then under the above assumption the operation of the circuit is simply explained by considering it to multiply the input signal by the square wave shown in fig. 22·5. This is a square wave having an amplitude of G/2 plus a d.c. component of G/2. It may be expanded into a Fourier series:

$$f_d\ (t) = \frac{G}{2} + \frac{2G}{\pi}\ (\sin \omega_c \cdot t + \frac{1}{3} \sin 3\ \omega_c \cdot t + \frac{1}{5} \sin 5\ \omega_c \cdot t\ \ldots\ldots)$$

where ω_c is the reference frequency in radians per second and $f_d(t)$ is the demodulating function.

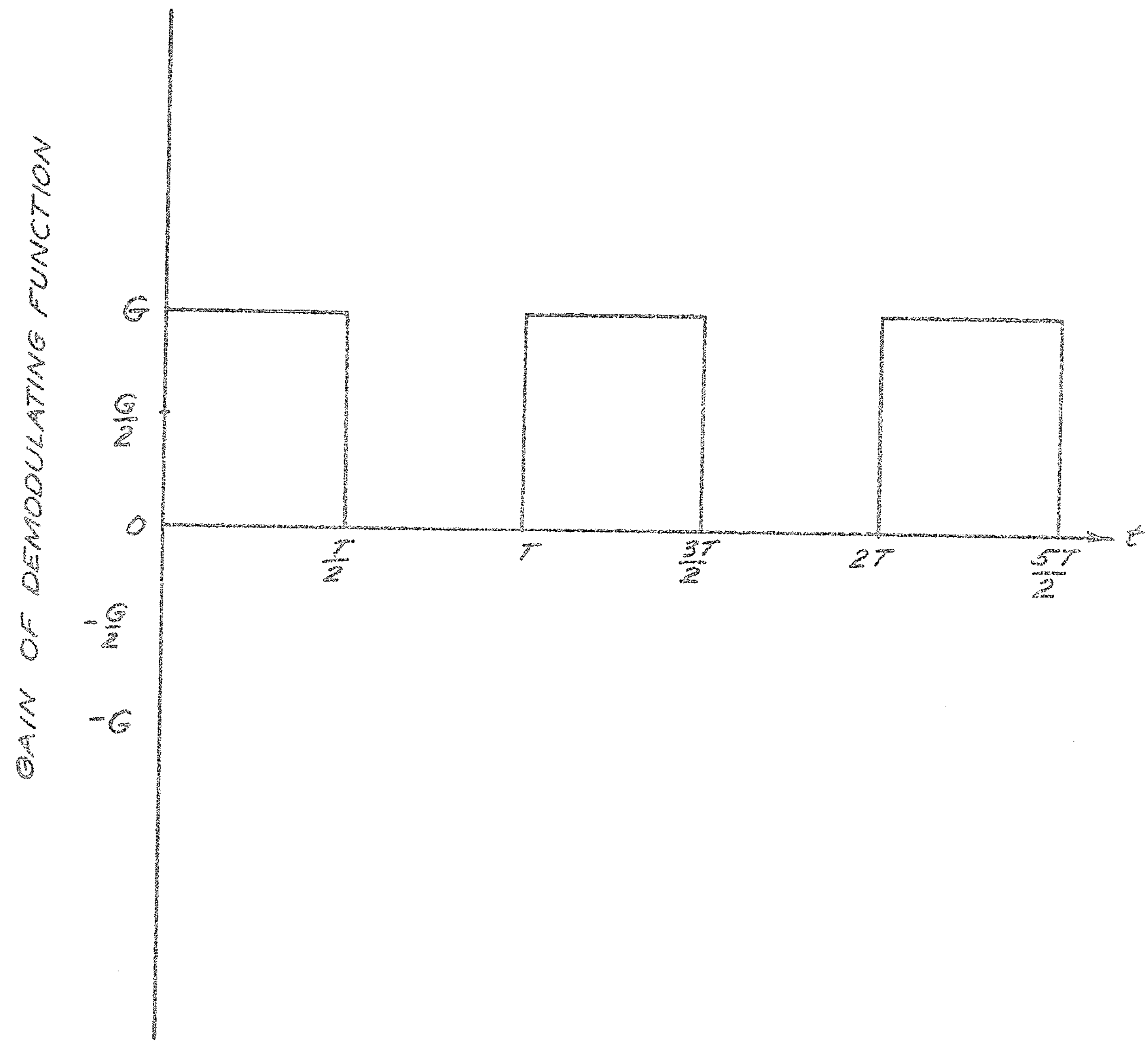
The input signal applied to the discriminator is assumed to be suppressed-carrier amplitude-modulated, and for the moment it is assumed that the carrier is a sinusoidal function of time. If the modulating function is F(t), then the input signal to the discriminator $v_i(t)$ is given by:

$$v_i(t) = F(t)\ \sin\ (\ \omega_c \cdot t - \beta\ )$$

where ω_c is the carrier frequency and β is the phase shift that exists between the signal input and the reference. The output of the discriminator $v_o(t)$ is the product of the input and the demodulating square-wave gain function:

$$v_o(t) = G \cdot F(t)\ \sin\ (\ \omega_c \cdot t - \beta)\ \left[\frac{1}{2} + \frac{2}{\pi}\ (\sin\ \omega_c \cdot t + \frac{1}{3} \sin 3\ \omega_c \cdot t + \frac{1}{5} \sin 5\ \omega_c \cdot t + \ldots\ldots)\right]$$

In expanding this expression, it is convenient to expand first



APPROXIMATE DEMODULATING FUNCTION

Fig. 22.5

the factor $\sin(\omega_c \cdot t - \beta)$ into $(\sin \omega_c \cdot t \cos\beta - \cos\omega_c \cdot t \sin\beta)$.
Multiplying this into the series and simplifying results in:

$$v_o(t) = G \cdot F(t) \Big\{ \cos\beta \Big[\frac{1}{\pi} + \frac{1}{2}\sin\omega_c \cdot t - \frac{2}{\pi}\Big(\frac{1}{3}\cos 2\omega_c \cdot t + \frac{1}{15}\cos 4\omega_c \cdot t + \frac{1}{35}\cos 6\omega_c \cdot t \ldots\ldots\Big)\Big] - \sin\beta \Big[\frac{1}{2}\cos\omega_c \cdot t + \frac{4}{\pi}\Big(\frac{1}{3}\sin 2\omega_c \cdot t + \frac{2}{15}\sin 4\omega_c \cdot t + \frac{3}{35}\sin 6\omega_c \cdot t \ldots\ldots\Big)\Big]\Big\}$$

The useful part of this output is provided by the first term in the first bracket, that is:

$$\text{useful } v_o(t) = \frac{G}{\pi}(\cos\beta) \cdot F(t)$$

All other terms represent ripple. It is noted that the original modulation function F(t) has emerged from the discriminator multiplied only by the constant factor $(\frac{G}{\pi}) \cos\beta$, without any phase lags or amplitude changes depending on frequency. It is concluded from this that the process of demodulation does not inherently introduce any phase lag (or lead) into the signal channel and that the phase lag normally found is entirely due to the ripple filters. As mentioned previously, these are almost always required to eliminate the objectionable effects of the ripple on the circuit following the discriminator.

The equation for useful $v_o(t)$ shows that the gain between input and useful output is proportional to the cosine of the phase shift between the signal input to the discriminator and the reference signal. The desirability of keeping this phase angle small if maximum gain is to be secured is therefore evident. However, it is also clear that phase shifts of the order of 10° or less are of no great consequence. Phase shift also has some effect on the magnitude of the ripple. This can be appreciated qualitatively by inspection of the second bracket of the general $v_o(t)$ equation. The terms in this bracket, which contribute a large ripple component, are proportional to the sine of the phase angle β and therefore vanish if the input signal and reference are in phase. A somewhat more quantitative idea of the effect of phase angle on output ripple is obtained by computing the magnitude of the harmonics from the general $v_o(t)$ equation as a function of β. The results of such a computation are given below for $G \cdot F(t) = 1$.

Harmonic	Magnitude
0	$0{\cdot}318 \cos\beta$
1	$0{\cdot}500$
2	$0{\cdot}210 \cdot \sqrt{1 + 3\sin^2\beta}$
4	$0{\cdot}043 \cdot \sqrt{1 + 15\sin^2\beta}$
6	$0{\cdot}018 \cdot \sqrt{1 + 35\sin^2\beta}$
8	$0{\cdot}010 \cdot \sqrt{1 + 63\sin^2\beta}$

This table shows clearly that the ripple voltage increases with β, but since the largest ripple-frequency component, the fundamental, is independent of β, the rms value of the ripple would not be expected to increase very much as β changes from

0° to 90°. It is illustrative, in this connection, to compute the rms value of the ripple for these two extreme values of β. For $\beta = 0^\circ$, the rms value of the ripple is 0·544 volts while for $\beta = 90^\circ$ the ripple increases to 0·70 volts. Thus it is clear that, although the phase shift does increase the ripple somewhat, the effect is not of any great importance in a half-wave discriminator.

At this point there is an important item which is worthy of some attention. This involves the determination of the maximum frequency of the input modulation F(t) that can be handled without ambiguity by a half-wave discriminator.

let $F(t) = V \sin \omega_s \cdot t$

substituting this into the general $v_o(t)$ equation yields:

$$v_o(t) = G \cdot V \left\{ \cos\beta \left[\frac{1}{\pi} \sin \omega_s \cdot t + \tfrac{1}{4} \cos (\omega_c - \omega_s)\, t - \tfrac{1}{4} \cos (\omega_c + \omega_s) t \ldots \right] - \sin\beta \left[\tfrac{1}{4} \sin (\omega_c - \omega_s) t + \tfrac{1}{4} \sin (\omega_c + \omega_s) t + \ldots \right] \right\}$$

(equation M).

The term yielding the useful part of the output is the first term of the first bracket, $G \cdot V \cdot \cos\beta \cdot \left(\frac{1}{\pi}\right) \cdot \sin \omega_s \cdot t$, with all else constituting ripple. Thus the lowest harmonic of the ripple is shifted downward by the modulating signal frequency and becomes $(\omega_c - \omega_s)$. It is clear that when $(\omega_c - \omega_s) = \omega_s$ or $\omega_s = \omega_c/2$, the signal frequency and the frequency of the lowest harmonic of the ripple are the same, and it is no longer possible to distinguish between signal and ripple. Hence, this contributes the important result that the absolute upper-frequency limit of a half-wave discriminator is equal to one-half the carrier frequency. Practical reasons such as the difficulty of constructing a ripple filter with a very sharp cut-off characteristic usually restrict the maximum permissible signal frequencies to values considerably below this limit. Thus a commonly observed rule of thumb applied to the design of feedback systems is that the loop-gain crossover frequency should be no more than one-tenth of the carrier frequency if a half-wave discriminator is used.

Equation M (page 88) contains additional information that is useful in the design of ripple filters. Usually these are simple low-pass RC networks. It is sometimes found that a simple filter of this sort designed to attenuate the ripple to a satisfactory degree also results in excessive attenuation and phase shift in the signal spectrum. In such cases advantage may be taken of the fact that the ripple is concentrated in frequency bands near the fundamental, second, fourth, and higher harmonics of the carrier frequency. This fact permits the use of a series of sharply-tuned, band-rejection filters, such as bridged-T or twin-T circuits which produce much less phase lag and attenuation at lower frequencies than the simple low-pass filter does. It can be shown that the phase shift produced by the filters can be made to approach zero, as the

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rejection band is reduced to zero width.* It would appear therefore that this might be the ideal adjustment, however, if more terms of the series indicated are written, it will be found that the frequency components found in the ripple are not simply the fundamental, second, fourth etc., harmonics of the carrier, but are actually $(\omega_c \pm \omega_s)$, $(2\omega_c \pm \omega_s)$, $(4\omega_c \pm \omega_s)$, etc. Hence if the input F(t) is confined to a spectrum ranging between d.c. and some maximum frequency $\omega_{s_{max}}$, the ripple will be found to be concentrated in bands of widths $2\omega_{s_{max}}$ centered at the fundamental, second harmonic, etc. The ripple filter should attenuate all frequencies in this band, and hence cannot be of zero width. It is concluded therefore, that in practice there is a certain minimum amount of phase lag contributed by a discriminator even if the most sophisticated type of ripple filter is employed.

According to the table listing the harmonics and their magnitudes, the fundamental and second-harmonic frequency components contribute the major share of the ripple. Hence if a band-rejection type filter is used to minimize the low-frequency phase lag, it is usually sufficient to attenuate only the fundamental and second harmonic by this type of network, with a simple low-pass structure to attenuate all the higher harmonics. Since this low-pass structure must

* Valley & Wallman -"Vacuum Tube Amplifiers" Radiation Lab. Series, vol. 18, McGraw-Hill Co. 1948.

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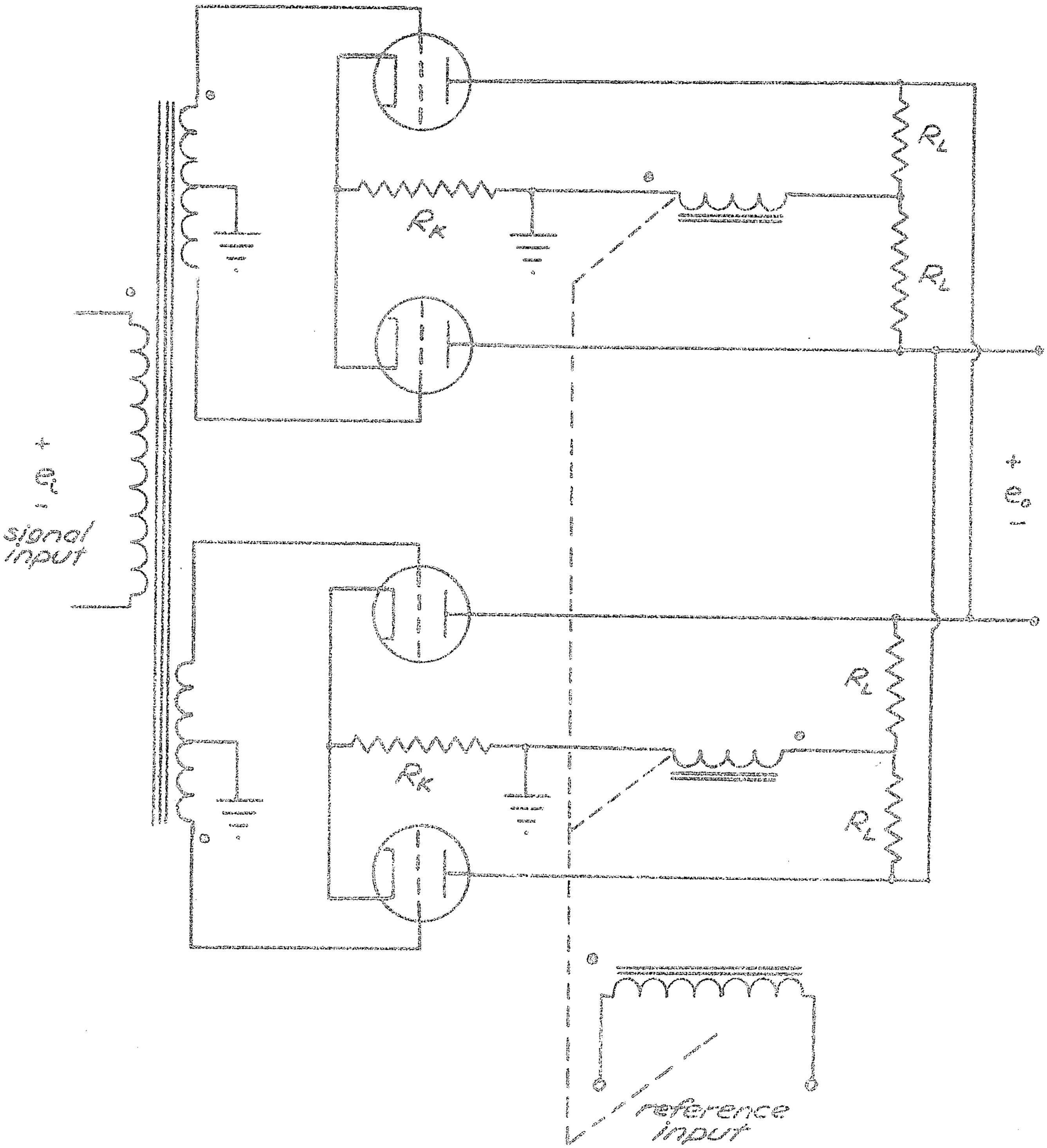
attenuate only the fourth and higher harmonic components, its bandpass may be relatively wide. Thus the phase shift contributed by this part of the filter at signal frequencies can also be reduced to a minimum. In this way it is possible to obtain good attenuation of all the ripple with relatively small effect on the signal band.

Full-Wave Discriminator Circuits

The analysis of the half-wave discriminator has shown that its upper frequency limit is equal to one-half the carrier frequency and that the lowest harmonic of the ripple is the fundamental or carrier frequency. Intuitively, it appears that the full wave discriminator, for which there is an output every half-cycle of the carrier, should be able to improve on this performance.

The circuit of a triode full-wave discriminator is shown in fig. 23. The dots shown on the transformers are the conventional polarity ~~ies~~ markings. When the upper two triodes conduct the lower two are disabled by the negative plate supply, and similarly when the lower two conduct, the upper two are disabled by the negative plate supply. If the phase of the input is such that the grid of the uppermost tube is positive when the upper two triodes conduct, then output signal e_o is positive in the direction shown. During the next half-cycle the lower two triodes conduct, and the



FULL-WAVE DISCRIMINATOR

fig. 23

connection of the input transformer is such that the lowermost grid is now negative. Hence the output signal is again positive. Thus for each cycle of the reference there are two pulses in the output, or a full-wave signal. Had the phase of the input relative to the reference been inverted, so that the uppermost grid was negative during the conduction period of the upper two tubes, the output would have consisted of negative pulses. If the phase of the input were such that during the conducting period of the upper triodes the uppermost grids were first positive and then negative, the output would consist of a pulse first positive and then negative. By this sort of reasoning the output waveshapes shown in fig. 24 can be obtained. It is noted that a signal 90° out of phase with the reference will result in zero output, and if the phase shift between signal and reference is somewhere between 0° and 90°, a reduced d.c. output results.

In analyzing this circuit, the assumption is made, as in the case of the half-wave discriminator, that the gain of the circuit is constant during the conduction period of either set of triodes. Also, as has been already noted, the operation of the circuit is such that, if during one half-cycle of the reference the output is positive for positive input, then for the next half-cycle the output is positive for negative input. Hence the circuit can be assumed to multiply the input by the square wave shown in fig. 25. This

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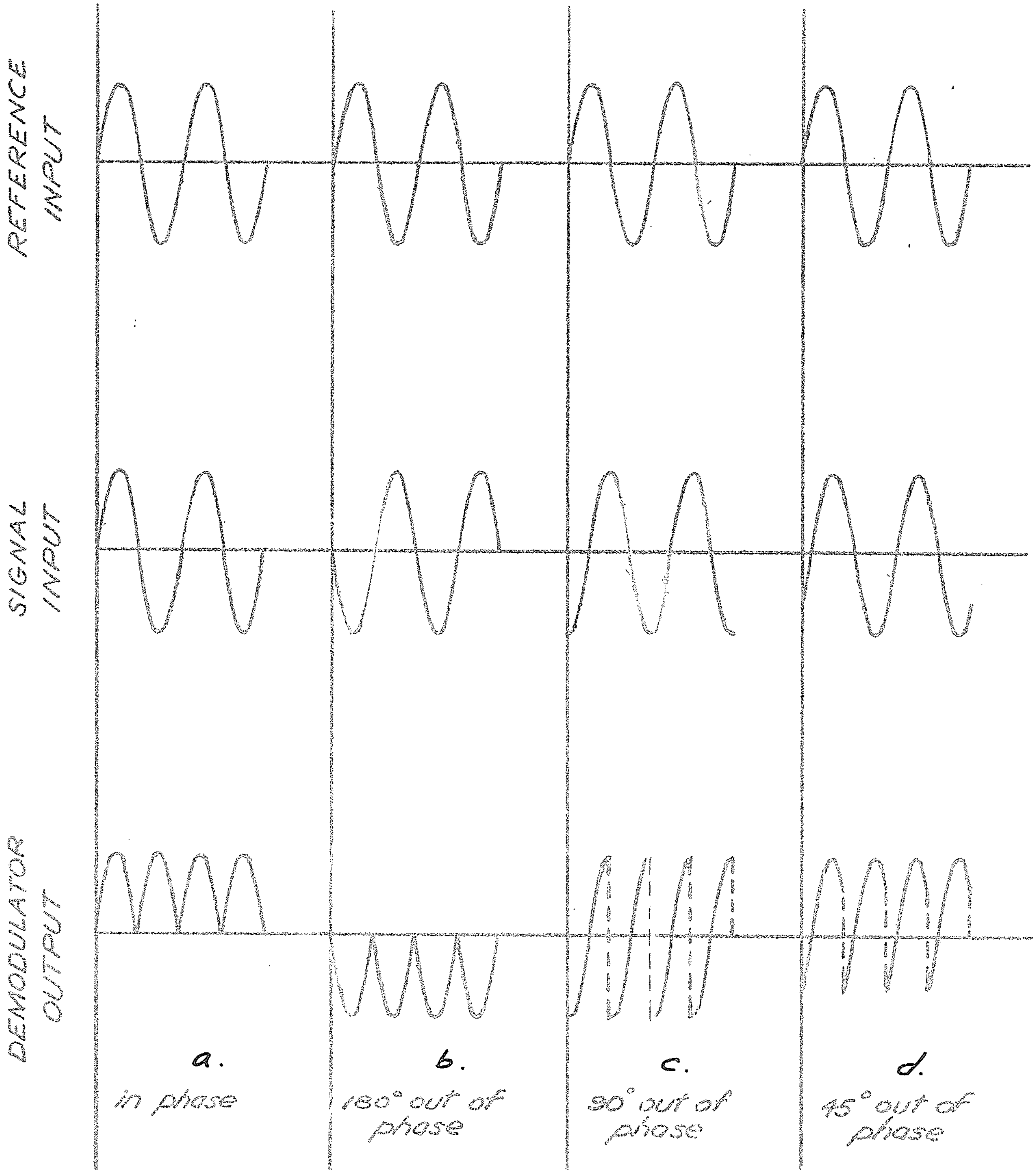
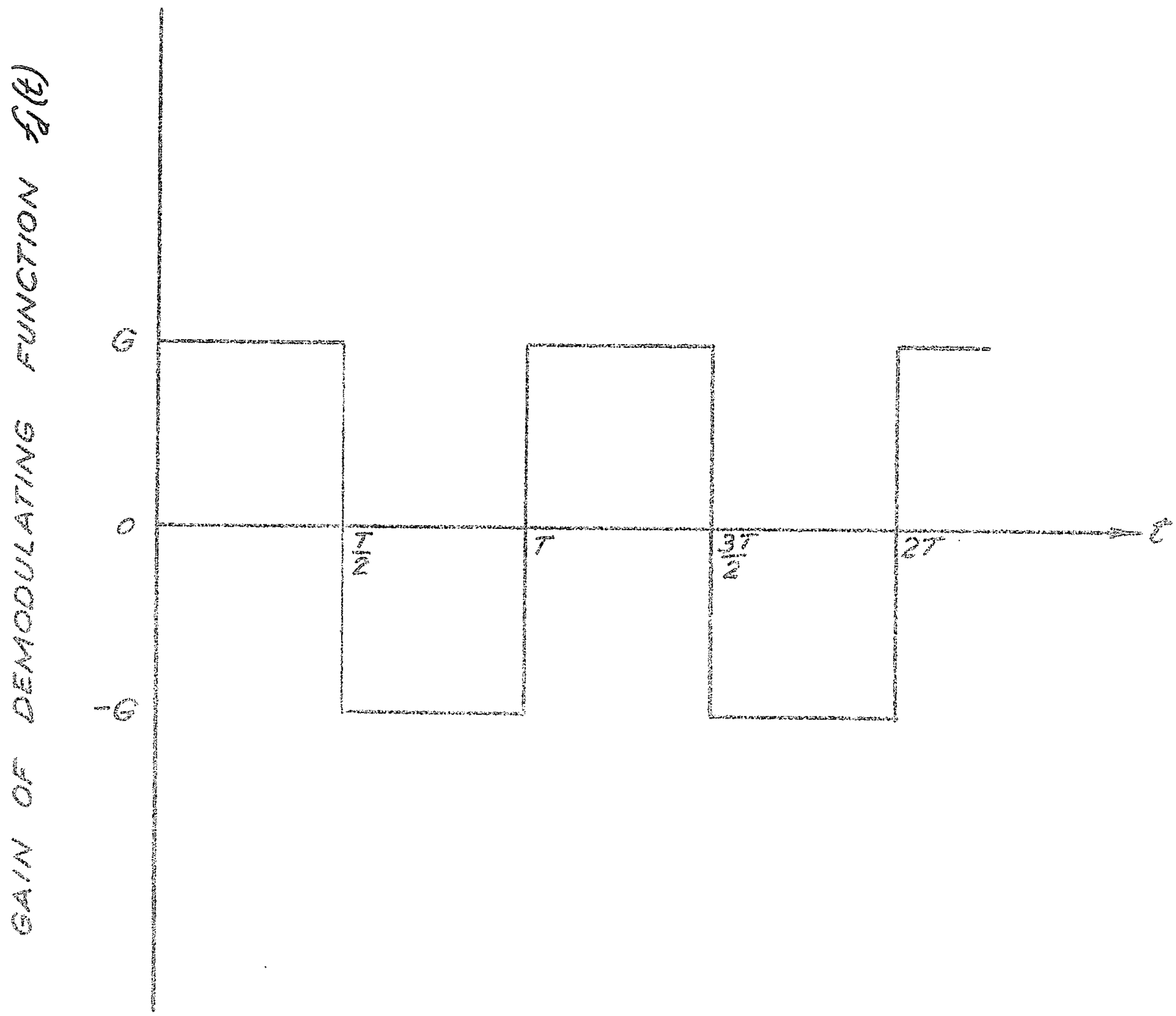


DISCRIMINATOR WAVESHAPES

fig. 24



APPROXIMATE DEMODULATING FUNCTION for FULL - WAVE DISCRIMINATOR

fig. 25

differs from the square wave considered in connection with the half-wave discriminator primarily in that its d.c. level is zero. The magnitude is arbitrarily taken as G, the gain of the amplifier in one of the switching modes. The discriminator gain function may be expanded into the following Fourier series:

$$f_d(t) = \frac{4G}{\pi}\left(\sin \omega_c \cdot t + \frac{1}{3}\sin 3\omega_c \cdot t + \frac{1}{5}\sin 5\,\omega_c \cdot t + \ldots\ldots\right)$$

If the discriminator input is again:

$$v_i(t) = F(t)\cdot \sin(\omega_c \cdot t - \beta)$$

then the output is given by the product of these two and may be expanded into the form:

$$v_o(t) = G\cdot F(t)\left\{\cos\beta\left[\frac{2}{\pi} - \frac{4}{\pi}\left(\frac{1}{3}\cos 2\,\omega_c \cdot t + \frac{1}{15}\cos 4\omega_c \cdot t + \frac{1}{35}\cos 6\,\omega_c \cdot t + \ldots\ldots\right)\right]\right.$$
$$\left. - (\sin\beta)\cdot\left(\frac{8}{\pi}\right)\left[\frac{1}{3}\sin 2\omega_c \cdot t + \frac{2}{15}\sin 4\,\omega_c \cdot t + \frac{3}{35}\sin 6\,\omega_c \cdot t + \ldots\ldots\right]\right\}$$

The useful part of the output is F(t) multiplied by the time invariant part of the series:

$$\text{useful } v_o(t) = \frac{2}{\pi}\cdot G \cdot \cos\beta \cdot F(t)$$

with all other terms contributing only to the ripple. Here again it may be seen that the input modulation F(t) emerges unchanged by the process of demodulation. Also the effect of β on the gain is identical for that found already for the half-wave discriminator. In fact, comparison of the equations for useful $v_o(t)$ for the half-wave and full-wave discriminator shows that, except for a factor of two, the only difference between the output of the full-wave and half-wave circuits is that in the former, the fundamental frequency component of ripple is absent. Hence the table of harmonic magnitudes constructed for the half-wave discriminator applies also to the full-wave discriminator if $G\cdot F(t)$ is set $= \frac{1}{2}$ and the first harmonic be zero.

The table may again be used to find the approximate rms value of the ripple voltage and to estimate the effect of β on it. Using only the harmonics up to and including the eighth, it is found that for $\beta = 0^\circ$ the rms value of the ripple is 0·216 volts, while for $\beta = 90^\circ$ it goes up to 0·473 volts. (Note that these values are computed with $G\cdot F(t) = \frac{1}{2}$ in order to get a direct comparison with the half-wave circuit). It is clear that for $\beta = 0^\circ$ the ratio of the rms ripple voltage to useful output is less than half as large in the full-wave as it is in the half-wave circuit. This is due to the absence of the fundamental, which is responsible for the major share of the ripple voltage in the half-wave discriminator. It is also apparent that the effect of phase shift between signal carrier and reference is more serious in increasing the ripple, and it is somewhat more important here to keep this phase angle low.

The amount of filtering required to reduce the ripple output following a full-wave discriminator to a permissable amount is much less than that needed to produce an equal attenuation of the ripple produced by a half-wave circuit. This is due to the smaller rms value of the ripple and to the fact that the lowest ripple harmonic is the second harmonic rather than the fundamental of the carrier frequency. For an example, assume that a double-section RC-filter, producing an attenuation proportional to the square of the frequency, is employed as a filter. Then for the same reduction in ripple relative to d.c. output, the filter used with the full-wave circuit may have about three times the passband (as defined by the reciprocal of the RC time constant of the two filter sections) of the filter used with the half-wave circuit. The superiority of the full-wave discriminator in causing reduced attenuation and phase lag of desired signal information is therefore evident.

In order to find the maximum frequency of the input modulation F(t) that can be handled without ambiguity by a full wave discriminator, let: $F(t) = V \cdot \sin \omega_s \cdot t$

the equation for $v_o(t)$ becomes:

$$v_o(t) = G \cdot V \left\{ \cos\beta \left[\frac{2}{\pi} \sin \omega_s \cdot t + \frac{2}{3\pi} \sin (2\,\omega_c - \omega_s) - \frac{2}{3\pi} \sin (2\,\omega_c + \omega_s)\, t \ldots\ldots \right] - \sin\beta \cdot \frac{4}{\pi} \left[\frac{1}{3} \cos (2\omega_c - \omega_s)\, t - \frac{1}{3} \cos (2\omega_c + \omega_s)\, t + \ldots\ldots \right] \right\}$$

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As before the first term of the series,

$$G \cdot V(\cos \beta) \frac{2}{\pi} \sin \omega_s \cdot t$$

represents useful signal, with all other factors being ripple, and it can be seen that the lowest harmonic of the ripple is $(2\omega_c - \omega_s)$. The upper limit on ω_s is reached when this lowest ripple frequency coincides with the signal frequency:

$$\omega_s = 2\omega_c - \omega_s$$

$$\omega_s = \omega_c$$

Thus the absolute upper limit is equal to the carrier frequency, a limit twice that of the half-wave discriminator. As was mentioned in connection with the half-wave circuit, the actual useful upper limit is normally considerably less than the theoretical maximum because of practical difficulties of filter design. However, if these difficulties are assumed to be identical in the two cases, the fact that the practical upper-frequency limit of the full-wave discriminator is twice as high as that of the half-wave discriminator is still true.

It should be pointed out that the result concerning the maximum theoretical frequency usable with a full-wave discriminator depends on the fact that the input signal was assumed to have a sinusoidal carrier. If the carrier were an accurate square wave, identical with the one assumed for the discriminator demodulating function, the process of demodulation would have subjected the signal to two successive multiplications by a square wave. This, at least theoretically, would be equivalent to multiplication by a constant, so that there would be no upper frequency limit and incidentally, no ripple in the output. While this result is primarily of theoretical interest, owing to the difficulty of generating the required accurate square waves, it does point out the desirability of square-wave modulation and demodulation in general.

As discussed in connection with the half-wave discriminator, it is possible to ~~the~~ take advantage of the fact that the ripple is concentrated in relatively narrow bands around the harmonics to design an improved type of ripple filter incorporating band-rejection filter sections. The use of this type of filter is of considerably greater practical importance with the full-wave than with the half-wave discriminator. For the half-wave discriminator the extra complication involved hardly seems justified, since the discriminator itself is not optimum and there is little point in trying to improve an inferior circuit when a better circuit is available. This argument does not apply, however, to the full-wave discriminator, for it represents the best that can be achieved without increasing the circuit complexity many-fold. Since the lowest harmonic of the ripple is the second and since the next one (the fourth) is almost five times smaller, a relatively large improvement in filtering efficiency is obtained by use of a single rejection filter to remove the band of frequencies around the second harmonic. All other harmonics can be removed easily by a standard RC low-pass filter and, as mentioned in connection with the half-wave discriminator, this part of the filter attenuates only the higher frequencies and has therefore a minimal effect on the signal band. The rejection filter must attenuate effectively all the frequencies in a band that is centred around the second harmonic of the carrier and has a width or bandpass equal to twice the maximum frequency expected in the signal envelope; ie:

bandpass is from $(2\omega_c - \omega_s)$ to $(2\omega_c + \omega_s)$

The improved performance possible with the full-wave discriminator is obtained at the expense of additional circuit complexity. In order to get accurate full-wave output, the two halves of the circuit must be carefully balanced, that is, the gain must be the same no matter which triode pair is conducting. If this is not the case, every other pulse will be larger than the two adjacent pulses and the output will then contain a component at the fundamental of the carrier frequency.

Other Discriminator Circuits

The two discriminator circuits described have been presented primarily as examples of practical circuits giving a half-wave or full-wave output. The theory outlined in these circuits is intended to explain the basic functions that a discriminator circuit is intended to perform, and also to outline the problems involved in this type of circuitry. Other circuits operating on the same principles may however be devised and may be ~~more~~ preferable, depending on the application. Among these other types of discriminators are the single-ended triode discriminator, the half-wave cathode follower discriminator, half-wave diode discriminators, electromechanical demodulators, the clamping discriminator, and the ring demodulator. A discussion of these circuits and other similar circuits is given by Greenwood, Holdman, and MacRae, "Electronic Instruments" - Radiation Lab Series Vol. 21 McGraw-Hill Book Co. 1948 pp. 383-386.

The ring demodulator circuit mentioned above will be discussed here in some detail as it is a simpler circuit to construct than the full-wave discriminator previously discussed and yet fulfills the intended purpose with the same degree of satisfaction. This circuit will be employed in the experimental work in connection with this dissertation.

Ring Discriminator

Diode discriminator circuits are sometimes preferred over triode circuits since the problem of drift is considerably reduced. For this reason diode discriminators are often used particularly in systems having sufficient a.c. amplification and therefore not requiring additional d.c. amplifiers.

The ring discriminator as shown in fig. 26 is a full-wave discriminator. Its operation, which is typical of all diode discriminators, may be explained as follows. The reference signal is assumed to be very much larger than the input signal therefore, for the half cycle during which point b is positive and point d negative, the lower two diodes conduct and the upper two are blocked. Hence there is an electrical connection between the output terminal and point c, the lower end of the input transformer.

The usual ring discriminator circuit has the output lead connected directly to the center-tap of the reference input transformer secondary. This center-tap is not usually sufficiently accurately located in the usual type of transformer used for this purpose and hence a potentiometer is shown in the circuit with the output lead attached to its center-tap. The potentiometer resistance must be sufficiently high so as not to draw excessive power from the reference signal source. This type of arrangement allows accurate balancing of the discriminator without an accurately center-tapped transformer secondary.

In the operation of the circuit, if the potentiometer center-tap is correctly placed so that the voltages that appear across the two halves of the potentiometer are equal, and if the impedances of the circuits b-c and c-d are exactly the same, there will be no voltage difference between point c and the

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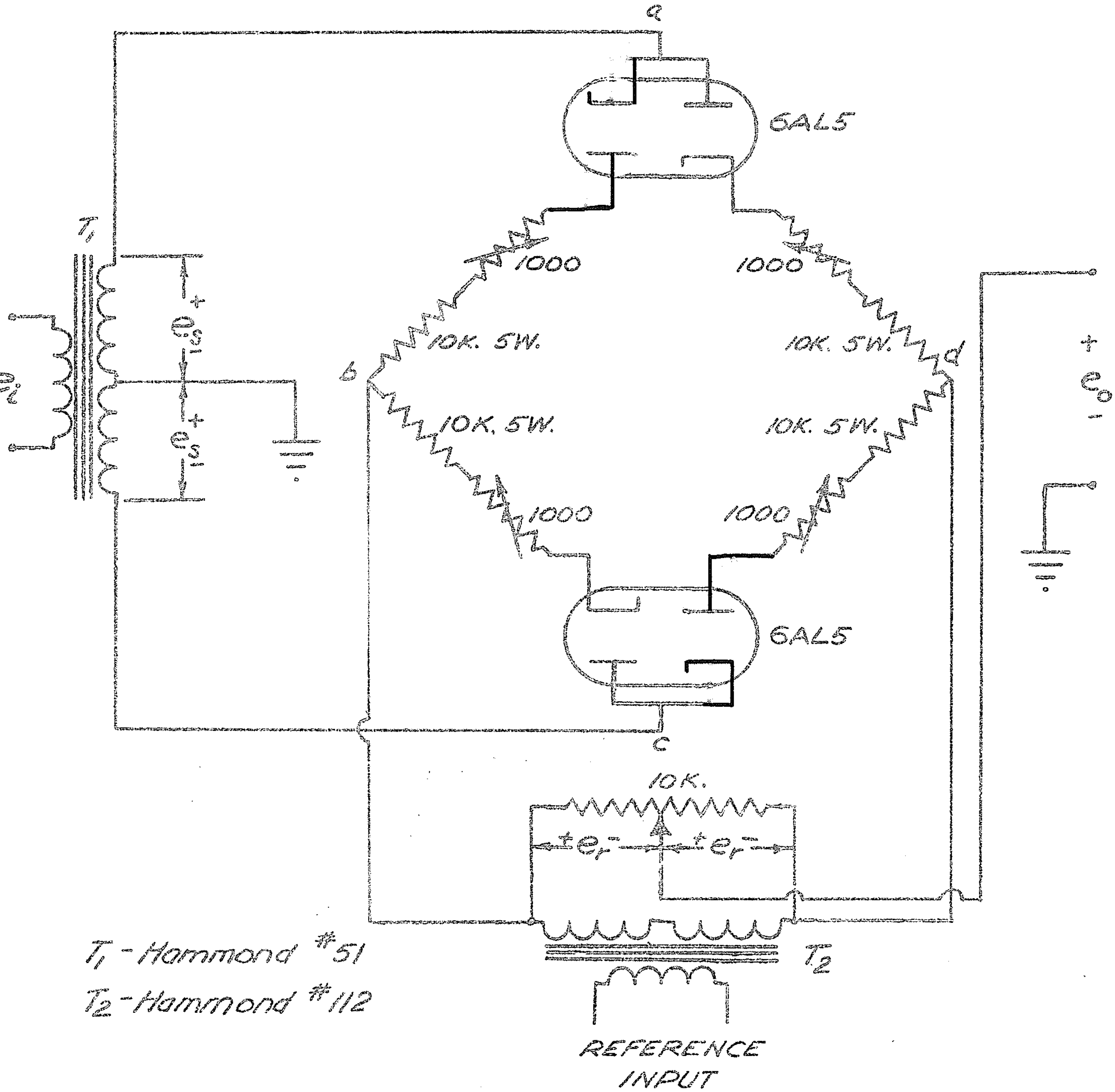

RING DISCRIMINATOR

fig. 26.

output. The output voltage will therefore be the same as that of point c. During the next half cycle, point b is negative and d is positive. Thus the ~~upper~~ lower two diodes conduct and the output is connected to point ~~a~~ c. Thus the circuit acts as a switch connecting the output either to point a or point c in synchronism with the reference signal. The resistances in the arms are required to limit the currents flowing in the diodes and to equalize the impedances of the four arms.

From the description it should be clear that the operation of the circuit consists of multiplying the input signal by a square wave of unit amplitude, if the turns ratio of the input transformer is unity. Hence the analysis of the full wave discriminator is directly applicable, and the output waveshapes observed under various operating conditions are shown in fig. 24 page 94. It follows that the lowest ripple frequency is the second harmonic of the carrier and that the maximum envelope frequency of the input is the carrier frequency.

In practice, inaccuracies between the bridge arms and inaccuracies in the location of the center-tap of the input transformer result in output waveshapes that depart considerably from the ideal forms shown in fig. 24. The effect of these unbalances is that the output contains a

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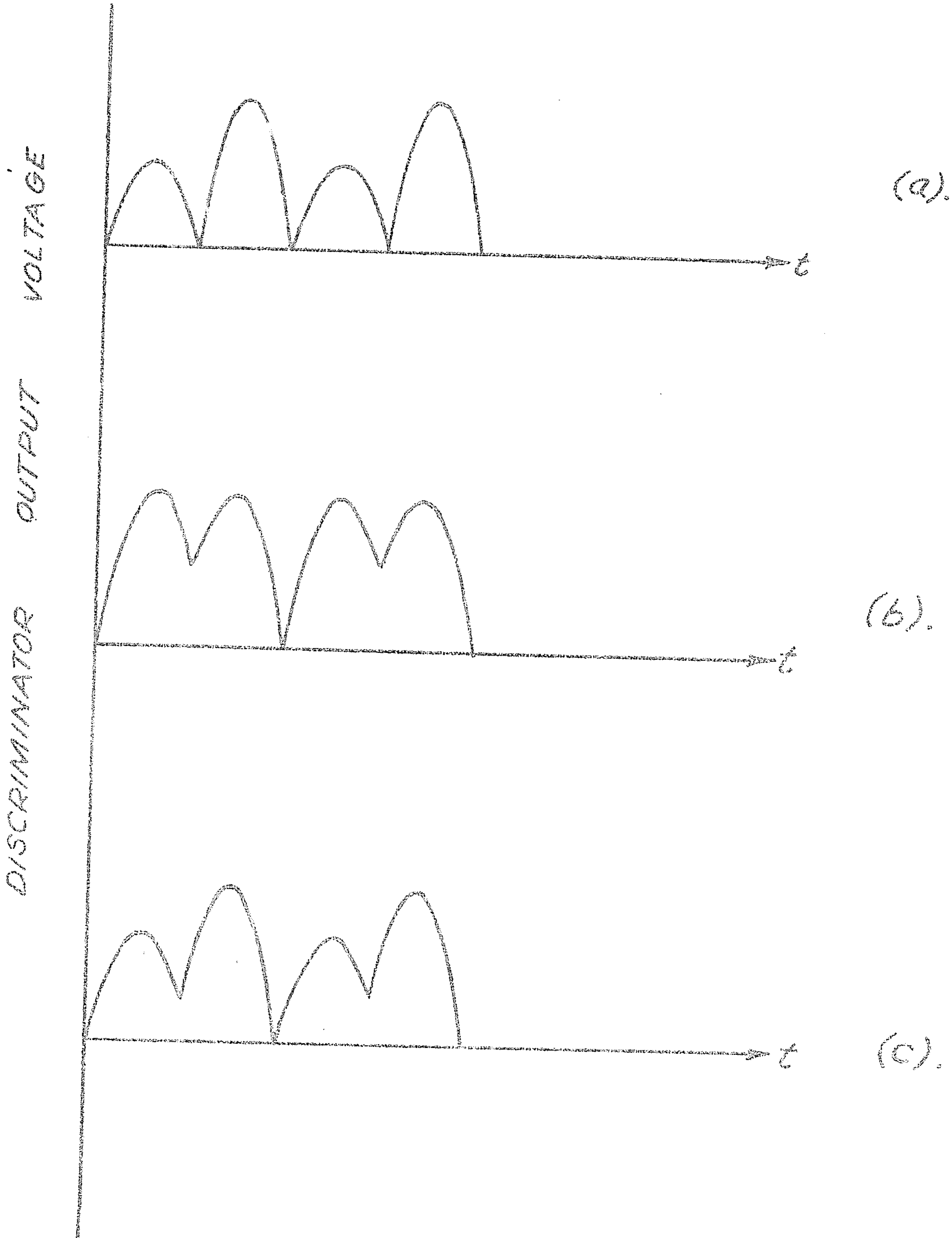
component at fundamental frequency, so that a filter with a reduced bandpass, is required to smooth out the ripple. Since this to some extent defeats the purpose of the full wave discriminator, some effort to balance the circuit properly is usually justified. The various adjustments required for this are best understood by considering the various output waveforms in some detail.

Typical waveforms observed when the signal is in phase with the reference are shown in fig. 27. The voltage of fig. 27(a) may be obtained by adding a sine wave, at fundamental frequency and in phase with the reference to a true full-wave rectified signal like that shown in fig. 24(a). It can be shown, by means of elementary circuit analysis, that this type of output results from inaccurate positioning of the center-tap ~~of the center-tap~~ on either the potentiometer or input transformer or from inequalities in the resistors used in the bridge arms. If the center-tap of the input transformer is incorrectly located then no compensating adjustment of either the resistance or the reference center-tap can be made to reduce this unbalance except at one signal amplitude. On the other hand, if the signal transformer is not at fault, then it should be possible by adjustment of any of the four resistors to produce a true full-wave rectified output. It may be found, however, that there may be some d.c. output for zero input. If this undesirable, then



OUTPUT WAVEFORMS RESULTING from CIRCUIT UNBALANCE

Fig 27

another adjustment must be provided. If the center-taps cannot be moved, one of the resistors in circuit b-a-d and one in circuit b-c-d of fig. 26, must be made adjustable to permit both zero d.c. output for no input and a balanced full-wave output to be achieved. It is always possible however to move the center-tap on the potentiometer but not usually the input transformer tap.

The waveform in fig. 27(b) results from the addition of a full-wave rectified voltage and a sine wave at fundamental frequency but 90° out of phase with the reference. This effect cannot be produced by resistance unbalance but may be caused by unequal capacitances between the two halves of the reference winding and ground. This in turn may be caused by a reference transformer that does not have an electrostatic shield between its primary and secondary windings. The effect may be simulated by a small capacitor connected from either point b or point d to ground of fig. 26. The remedy is to use a transformer with a Faraday shield or to add compensating capacitors between points b or d and ground. In fig. 27(c) is shown the result when both the effects described occur simultaneously. Both resistive and capacitive adjustments must then be made to convert the output to the balanced full-wave form.

In addition to the balancing problem, another point of interest in this circuit is the maximum input signal for which the circuit will still function properly. Let e_o be the output voltage, and e_a, e_b, e_c and e_d be the voltage of points a, b, c, and d respectively all with respect to ground. The voltage across each half of the reference input potentiometer is termed e_r. This maximum input signal can be determined by considering the upper two diodes completely open circuited when the lower two conduct. This leaves a simple situation with voltage $2e_r$ applied across b-c-d and hence causing a current to flow in this circuit.* The same current will flow in b-c as in c-d and therefore if the resistances of both these arms are equal the voltages appearing across them will be equal. In traversing from the centertap of the potentiometer connected across the secondary of T_2 to point c the sum of the voltages is $+ e_r - e_r = 0$, hence

$$e_o = e_r - e_r + e_c$$

$$e_o = e_c$$

e_c however is exactly the voltage across one half of the signal input transformer secondary thus

$$e_c = -e_s$$

Similarly e_a is equal to the voltage across the other half of the secondary:

$$e_a = e_s$$

Therefore:

$$e_o = e_c = - e_a$$

*(e_o open circuited)

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and the voltage across arm a-b becomes

$$2e_a - e_r$$

In the same manner the voltage across arm a-d is given by $(2e_a + e_r)$. The reference voltage has the polarity indicated since the lower rectifiers are conducting, hence if e_a is positive the voltage across the rectifier in arm a-b becomes positive when e_a exceeds $\frac{e_r}{2}$ and the rectifier conducts. For negative e_a the voltage across the rectifier arm a-d becomes negative when e_a exceeds $\frac{e_r}{2}$ and this rectifier conducts. Since neither of these rectifiers should conduct when the lower set conducts, this represents improper operation. To obtain proper operation it is therefore necessary that

$$e_r > 2e_a \quad \text{or} \quad e_r > 2e_s$$

It is noted that this computation has also given the maximum back-voltage on the rectifiers. The back voltage across the rectifier in arm a-d e_{ad} is:

$$e_{ad} = 2e_a + e_r$$

$$= 2e_s + e_r$$

now if $e_r = 2e_s$

$$e_{ad} = 4e_s = 2e_r$$

For $e_r > 2e_s$ the back voltage will exceed $2e_r$.

CHAPTER 8

OUTPUT FILTER NETWORKS

Output Filter

The output from the discriminator is in the form of a half-wave or full-wave rectified signal with varying polarity. To obtain the useful signal component from this wave it is necessary to employ a filter at the discriminator output. The signal containing the desired information will have a frequency somewhere in the range from d.c. to the frequency corresponding to the maximum angular rotation or oscillation of the synchro generator rotor. As explained previously it is desirable, from a practical standpoint, to keep this maximum rotation or oscillation frequency safely below the point where any attenuation of the useful signal might occur due to the action of the filter. To obtain maximum usefulness from the system it is therefore desirable to design the filter system so that the point at which attenuation of the useful signal begins is at as high a frequency as possible. This will allow a maximum range of angular ~~rotation~~ velocity or oscillation for the generator rotor.

It was found in the analysis of the full-wave discriminator that the maximum allowable useful signal frequency was ω_c the carrier frequency. This brings up a point with respect to the design of the filter. The filter must be capable of attenuating the harmonic components while not attenuating the actual useful signal. Since the lowest harmonic of the ripple is $(2\omega_c - \omega_s)$, where ω_s is the signal frequency, the filter must be able to distinguish between ω_s and this $(2\omega_c - \omega_s)$ frequency, but if $\omega_s = \omega_c$, as should theoretically be possible, the lowest harmonic of the ripple would have the same frequency as the useful signal and the filter would be unable to separate them. Thus ω_s must be less ω_c by an amount entirely dependent on the sharpness of the filter. A filter with a steep leading edge attenuation characteristic will therefore provide a higher filtering efficiency than a filter with an attenuation characteristic which did not have as steep a leading edge.

In the experimental work here a full-wave discriminator was employed. Since this type of discriminator theoretically eliminates the fundamental, the filter design should only be concerned with the suppression of signals in bands centered at the second, fourth, and higher harmonics of the carrier frequency. Since the lowest frequency component of the ripple is the second harmonic and since the next one, the fourth, is almost five times smaller, a relatively high filtering efficiency will be obtained by using a single rejection filter to remove the band of frequencies around the second harmonic, coupled with the use of a standard low-pass filter to attenuate the higher frequencies. This relatively high filtering efficiency is mainly due to the attenuation characteristic of the band rejection filter which shows a much steeper leading edge than the standard low-pass filter and thus allows for a wider useful signal bandwidth.

The low-pass filter should start to attenuate at a frequency near where the band rejection filter attenuation begins so as to insure maximum attenuation of the higher frequency harmonics.

In actual construction of a system such as this, some unbalance will almost always be present in the discriminator and consequently some fundamental frequency component may occur in the output. If it is desired to eliminate this fundamental component an extra band rejection filter may be used at the fundamental frequency, or since there would be no band of frequencies to be eliminated any simple null network which would attenuate the first harmonic frequency alone would suffice. This would consequently reduce the maximum allowable ω_s to the point where the null network used to eliminate the fundamental had no attenuating affect on the useful output signal.

Selection of the Type of Band Rejection Filter

The type of band rejection filter chosen for a particular application depends on several factors.As explained previously the filter should be designed to attenuate frequencies in the band $(2\,\omega_c - \omega_s)$ to $(2\,\omega_c + \omega_s)$ introducing as small as possible an effect below this region. There are many types of filters such as R-C networks, crystal filters, mechanical filters, and certain networks employing operational amplifiers to list a few, which are capable of producing the desired effect. The choice of type depends mainly on the application and its governing specifications. In many applications a simple null network based on a bridged - T or twin - T design will often be sufficient and hence the expense of a more elaborate filter can be avoided. A description of the theory and design procedure for the twin - T network will be included here since this network will be employed in the experimental work.

R.C. Twin-T Filter as a Band Rejection Filter

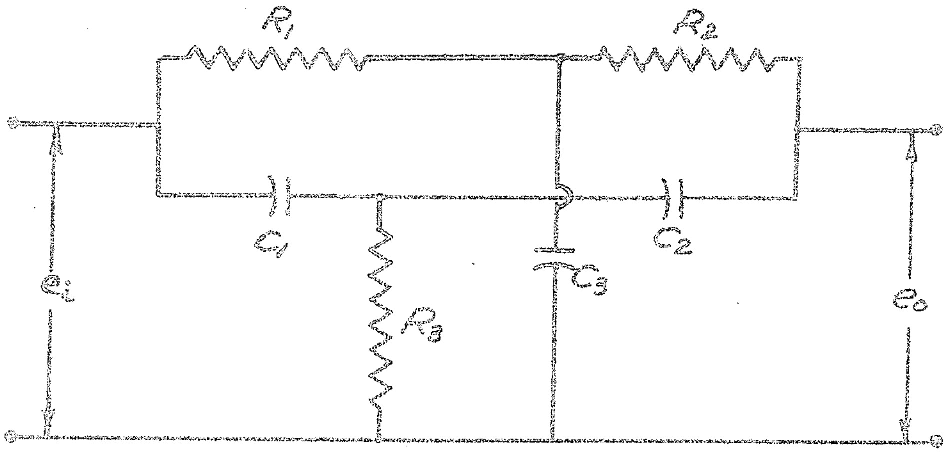
The twin-T network is shown in fig. 28. The transfer function of the unloaded filter is:

$$\frac{e_o}{e_i} = H(s) = \frac{N(s)}{\Delta(s)} \tag{1}$$

where
$$N(s) = \left[R_1R_2R_3C_1C_2C_3\, S^3 + (R_1 + R_2)\, R_3C_1C_2\, S^2 + R_3\,(C_1 + C_2)\, S + 1\right] \tag{2}$$

and
$$\Delta(s) = \left[R_1R_2R_3C_1C_2C_3\, S^3 + \left\{(R_1 + R_2)\, R_3C_1C_2 + (R_2 + R_3)\, R_1C_2C_3 + R_1R_3C_1C_3\right\} S^2 + \left\{R_1\,(C_2 + C_3) + R_2C_2 + R_3\,(C_1 + C_2)\right\} S + 1\right] \tag{3}$$

To use this network as a band rejection filter there must be a frequency at which there is a null in the corresponding



TWIN - T NETWORK

Fig. 28

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harmonic response function. The numerator of the transfer function is a cubic polynomial and therefore has three zeros. For a null at ω_r, two of the zeros must be $j\omega_c$ and $-j\omega_c$ and N(s) must have the form:

$$N(s) = \left[1 + \frac{S}{j\omega_r}\right]\left[1 - \frac{S}{j\omega_r}\right]\left[1 + \frac{S}{\gamma}\right] = \frac{S^3}{\gamma\omega_r^2} + \frac{S^2}{\omega_r^2} + \frac{S}{\gamma} + 1 \qquad (4)$$

where γ is the third zero, which must be negative and real. Comparison of the two equations for N(s) indicates that

$$\frac{1}{\omega_r^2} = (R_1 + R_2)\, R_3 C_1 C_2 \qquad (5)$$

$$\frac{1}{\omega_r^2} = \frac{R_1 R_2 R_3 C_1 C_2 C_3}{R_3 (C_1 + C_2)} \qquad (6)$$

The second relation is obtained by dividing the coefficient of S^3 by that of S. Equating these last two expressions and cancelling common factors gives for the null condition

$$\frac{C_3}{C_1 + C_2} = \frac{R_3 (R_1 + R_2)}{R_1 R_2} = n \qquad (7)$$

where n is any real positive number. It can be shown that n has an optimum value depending on the ratio of R_1 to R_2 and C_1 to C_2. If the network is symmetrical, that is, with $R_1 = R_2$ and $C_1 = C_2$, it will be found that the optimum value for n is 1.

If the condition that a null exists in the corresponding harmonic response function of the transfer function is met, then the expression for H(s) can be simplified considerably.

To do this it is convenient to let

$$C_2 = \alpha C_1 \qquad R_2 = \beta R_1 \tag{8}$$

Substituting for R_2 in equation (7) yields

$$R_3 = \frac{n \beta}{1 + \beta} \cdot R_1 \tag{9}$$

and using this result and equation (8) in equation (5) it is found that

$$R_1 C_1 = \frac{1}{\omega_r} \sqrt{\frac{1}{\alpha \cdot \beta \cdot n}} \tag{10}$$

Finally the frequency variable S is again normalized with respect to the rejection frequency ω_r by defining

$$p = \frac{S}{\omega_r} \tag{11}$$

Substitution of all these results into equation (1) yields, after some algebraic manipulation, the normalized transfer function

$$H\left(\frac{S}{\omega_r}\right) = H(p) = \frac{1 + p^2}{1 + Kp + p^2} \tag{12}$$

where $K = \dfrac{n(1 + \alpha) + \alpha(1 + \beta)}{\sqrt{\alpha \cdot \beta \cdot n}}$ (13)

Thus, for a given null frequency the circuit behavior is completely determined by n, α, and β, the latter two defining the ratios C_2/C_1 and R_2/R_1 respectively.

The sharpness of the null at the rejection frequency may be defined in a number of ways, but a convenient one is to consider the slope of the magnitude ratio at the null as a measure of sharpness. Accordingly the optimum value of K is defined as the value giving the steepest slope of the magnitude function at $\omega = \omega_r$. By differentiation of $\left|H\left(\frac{j\omega}{\omega_r}\right)\right|$ and then setting $\omega = \omega_r$ it is found that the slope at $\omega/\omega_r = 1$ is $-2/K$. Hence the optimum value of K is the smallest realizable value. In determining this value it must be kept in mind that α, β, and n must be positive real numbers. It is convenient to express K as the sum of two terms:

$$K = \left[\sqrt{\frac{n}{\alpha\beta}} + \sqrt{\frac{\alpha\beta}{n}}\right] + \sqrt{\frac{\alpha}{\beta}}\left[\sqrt{n} + \frac{1}{\sqrt{n}}\right] \qquad (14)$$

The two terms indicated by the parentheses are both positive and K is therefore minimized by minimizing each one separately. The first term has the form $\left(x + \frac{1}{x}\right)$. By differentiation this function is easily shown to be minimum for $x = 1$ hence one condition for minimum K is

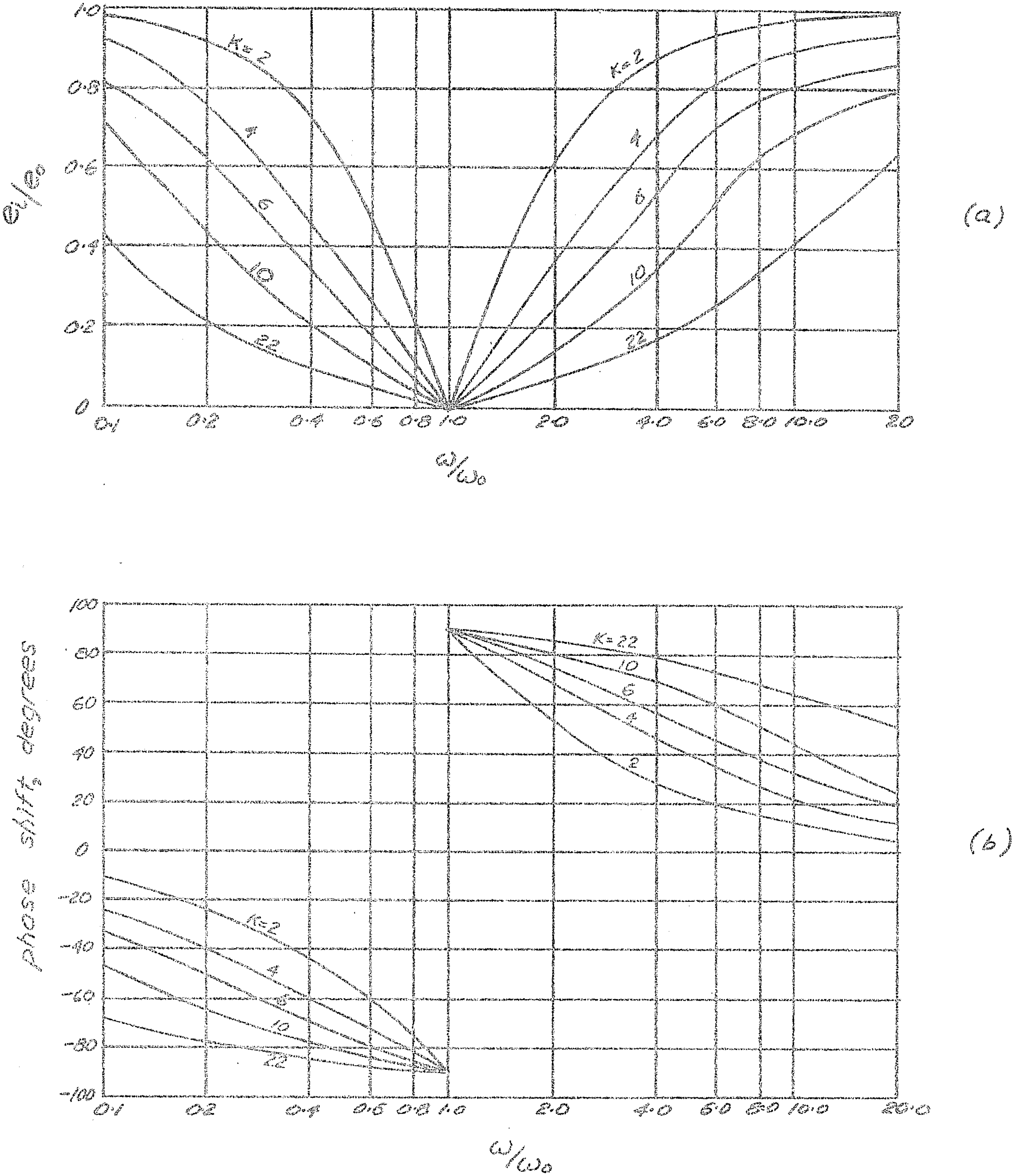
$$\frac{\alpha\beta}{n} = 1$$

$$\text{or} \qquad n = \alpha\beta \qquad (15)$$

Substituting this value of n into the expression for K gives

$$K = 2 + \alpha + \frac{1}{\beta} \qquad (16)$$

It is clear that the minimum value of K is 2 and that it is approached by making α as small as possible and β as



CHARACTERISTICS OF THE TWIN-T

fig. 29

large as possible. In practice, if the twin-T drives a grid of a vacuum tube, R_2 should normally not exceed 1 megohm. Also, since R_1 should be about ten times as large as the source impedance of the input, a practical minimum on R_1 is about 10,000 ohms. Hence a practical maximum value for β is about 100. Similar considerations limit α to about 0·01. Hence the practical minimum for K is of the order of 2·02.

The twin-T is often built symmetrically, that is, with $\alpha = \beta = 1$. From equation (15) it is known that for minimum K, $n = \alpha\beta = 1$. For these conditions $K = 4$, or twice its optimum value. As can be seen in fig. 29, as K increases the filter produces a decidedly larger phase lag at frequencies below the rejection frequency than the optimum value does. It is also noted that for frequencies exceeding the null frequency the phase of the harmonic response function is a lead angle.

Band Rejection Filter Design

In theory the use of a full-wave discriminator totally eliminates all fundamental carrier frequency components from the output signal of the discriminator. The major ripple component is therefore the second harmonic. Since unbalance and other practical defects in the discriminator are usually present to at least some degree, some

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fundamental will exist in the output signal. As was explained in the section on discriminators this effect can be kept to a minimum by careful balancing and aligning of the unit and this minimum should be sufficiently small to not require filtering for most normal applications. In designing a filter to be used for experimental work in connection with this dissertation a band rejection filter will be employed to attenuate the bandwidth of ripple frequencies around the second harmonic with only a standard low pass filter being used to remove the fourth and higher harmonics.

In feedback control systems it is usually desirable to minimize all phase lags added to the system. In using a twin-T network as a band rejection filter it is therefore desirable to use as small a K value as can be practically employed so as to minimize the added phase lag in the signal bandwidth. The K value must be also selected so as to provide sufficient attenuation over the entire ripple bandwidth around the second harmonic. This is important since the sharpness of the filter does increase with decreasing K. In the experimental work here a maximum upper limit of roughly 1000 oscillations per minute modulating frequency would easily include all expected situations. Thus the bandwidth of the rejection filter should attenuate to a negligible amount frequency components from about

780 to 820 cycles per second, the carrier frequency being 400 c.p.s. The plot of attenuation against frequency for the twin-T network shows that any K value down to K = 2 would suffice and therefore a K value of 2·1 will be used.

To insure that R_1 is much greater than the impedance of the source feeding the filter let $R_1 = 10,000$ ohms. The parameters defining the filter performance should be chosen such that X_{c_1} at 800 c.p.s. will also equal 10,000 ohms for the same reason for which R_1 is made that value. Let $\alpha = \cdot 05$ and $\beta = 20$. From equation (15), for minimum K, $n = \alpha\beta = 1$. Now going back to equation (10)

$$C_1 = \frac{1}{\omega_r R_1}\sqrt{\frac{1}{\alpha \cdot \beta \cdot n}}$$

$$= \frac{1}{6{\cdot}28 \times 800 \times 10000}$$

$$\doteq 0{\cdot}02 \text{ mfd}$$

which has a reactance of 10,000 ohms at 800 c.p.s.

$$C_2 = \alpha C_1 = \cdot 05 \times \cdot 02 = 0{\cdot}001 \text{ mfd}$$

$$R_2 = \beta R_1 = 20 \times 10,000 = 200,000 \text{ ohms}$$

$$C_3 = n(C_1 + C_2) = \cdot 021 \text{ mfd.}$$

$$R_3 = \frac{nR_1R_2}{R_1+R_2} = \frac{10,000 \times 200,000}{210,000} = 9,540 \text{ ohms}$$

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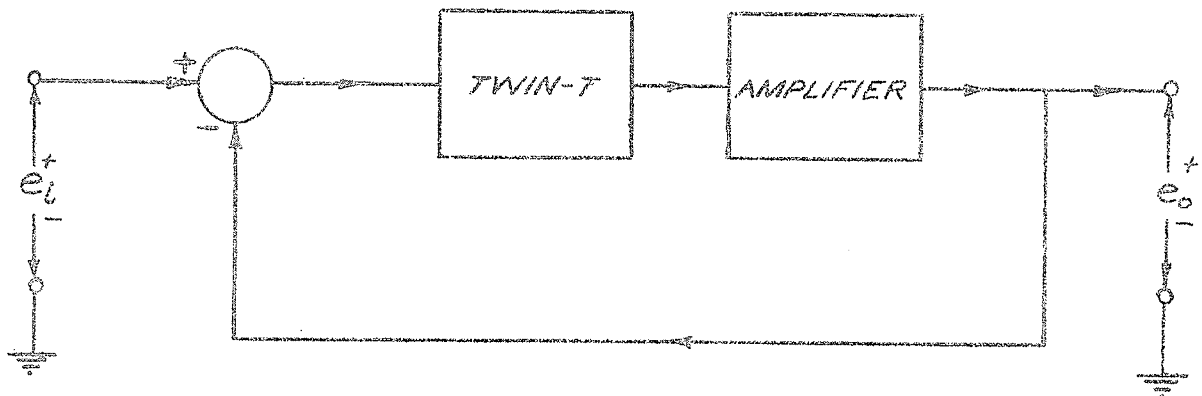
In using the twin-T network as a band rejection filter, even if a fairly sharp filter is achieved by using a relatively optimum K value, the resulting unit will still not allow full use of the advantages made possible by using a full-wave discriminator. Theoretically the 400 cycle carrier frequency system can make use of signal frequencies which approach 400 c.p.s. The band rejection filter should not attenuate below this point then but unfortunately even a twin-T with the optimum K value of 2 will attenuate below this frequency with the transmission factor still being 3 db down at 40% of the null frequency. Obviously then much better band rejection filters can be designed if the system warrants such. The twin-T filter designed here and used in the experimental work was however quite adequate for the purpose ~~though~~ since the highest expected signal frequencies would not exceed 20 c.p.s. and the filter experimentally proved to be quite flat up to 50 c.p.s.

The performance of the twin-T network can be greatly improved with respect to the sharpness of the null by using it in a circuit known as a Rejection Amplifier. The block diagram of this is shown in fig. 30. The operation of this circuit is most easily explained by determining the overall transfer function e_o/e_i. Application of standard techniques and the transfer function for the twin-T gives:



REJECTION AMPLIFIER

fig. 30

$$\frac{e_o}{e_1} = \frac{G}{G+1} \cdot \left[\frac{p^2 + 1}{p^2 + \frac{K}{G+1} \cdot p + 1} \right]$$

where G is the amplifier gain. Comparison with the original twin-T transfer function (equation 12) indicates that the effect of the circuit is essentially to change the K in the denominator of the transfer function to $K/(1 + G)$. It was shown previously that the slope of the magnitude of the harmonic response function at the null was $-2/K$ for the twin-T; hence it is clear that the action of the amplifier is to make the slope steeper, and therefore to sharpen the null.

Low-Pass Filter

To eliminate the ripple components remaining after attenuation of the second harmonic all that is required is a simple low pass filter which will effectively attenuate the fourth and higher harmonic components without introducing attenuation or detrimental phase shifts into the signal passband. This filter may be used in tandem with the band rejector filter without isolation between the two provided the low-pass unit does not provide too small a load impedance to the twin-T network and thus impair the performance of the twin-T. A simple lag network will be sufficient for this purpose and with a resistance value of 1 meg ohm will provide no loading effect on the twin-T. Using a break frequency of 1000 c.p.s. provides an approximate 2:1 attenuation of the fourth harmonic without introducing any appreciable phase lag into the signal passband.

CHAPTER 9

BASIC EXPERIMENTAL WORK

Tests with a Basic Synchro Pair

Since the prime use of synchro devices in control systems is in the detection of shaft angular position error and the subsequent production of an error signal proportional to this, experimental work in connection with this dissertation should begin with a look at this basic employment of the device. To do this, a synchro generator and a synchro control transformer were connected together as shown in fig. 2 page 10. The units were mounted on the synchro test bench described in the appendix. The specifications of the units employed are:

General Electric Synchro Generator
Model 2J1F1 - Serial No. 433691
Primary voltage 115 v.
Secondary phase voltage 57·5 v.
400 cycle.

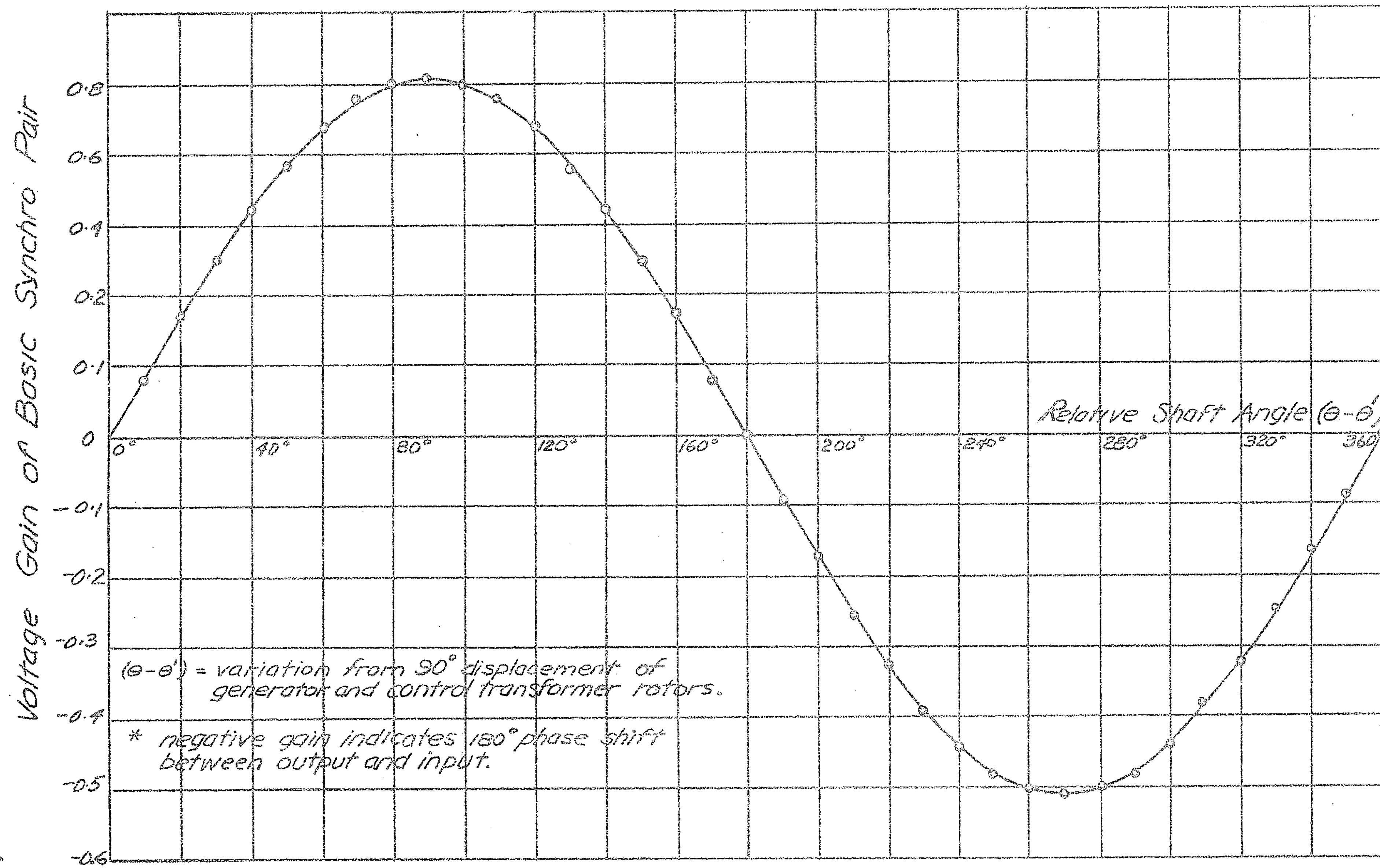
General Electric Synchro Control Transformer
Model 2J1G1 - Serial No. TB3
Stator line-to-line voltage 57·5 v.
400 cycle.

The pair was aligned for the test by setting the control transformer so that its dial plate registered "zero" degrees. The synchro generator rotor was then positioned at the location where the output signal from the control transformer

........./127b.

rotor was a minimum. This minimum control transformer output should theoretically have been zero but due to effects discussed under "Synchro Errors" the minimum output obtainable was 0·001 volts. The dial plate of the generator was then positioned to zero degrees while the generator rotor was held at the minimum output position. The test to verify the relation between the control transformer output voltage and the relative angular position of the two shafts, for a constant voltage input to the synchro generator, was carried out by fixing the control transformer at the zero degree position and then obtaining the output voltage from various positions of the generator shaft. In relation to figure 8, page 21b this constitutes setting θ' equal to zero and then varying θ. The results of the test are shown in the accompanying characteristic, fig. 31. The theoretically predicted sinusoidal variation of output with shaft relative angular displacement was obtained.

Although the minimum obtainable control transformer output voltage was 0·001 volts rather than the zero volts, the minimum output was taken as zero to facilitate plotting of the gain characteristic. The gain characteristic has been plotted in such a way as to indicate the 180° phase shift that occurs when the synchro generator and control transformer rotors are between 180° and 360° apart. This indication is given by plotting the gain as a negative value.



GAIN CHARACTERISTIC OF BASIC SYNCHRO PAIR

fig. 31

for gains occurring with a 180° phase shift. Although the results of this test are rather easily described and displayed,, the obtaining of these results and those in the following test took a great deal of time and effort.

Synchro System Using a Differential Generator

This basic synchro system is now enlarged by inserting a differential generator between the units of the original basic pair. This test is intended to experimentally verify the theoretical development showing that the relative angle between the synchro generator and the control transformer is shifted by an amount equal to the angle introduced by the differential generator. Tests were made for three differential generator angles. For each test it is seen from the results shown in fig. 32, that the relative angle between the basic pair units is actually shifted by the differential generator angle.

General Electric Selsyn Differential Generator
Model No. 2J1H1 - Serial 6225688
57·5 - 57·5 volts
400 cycles.

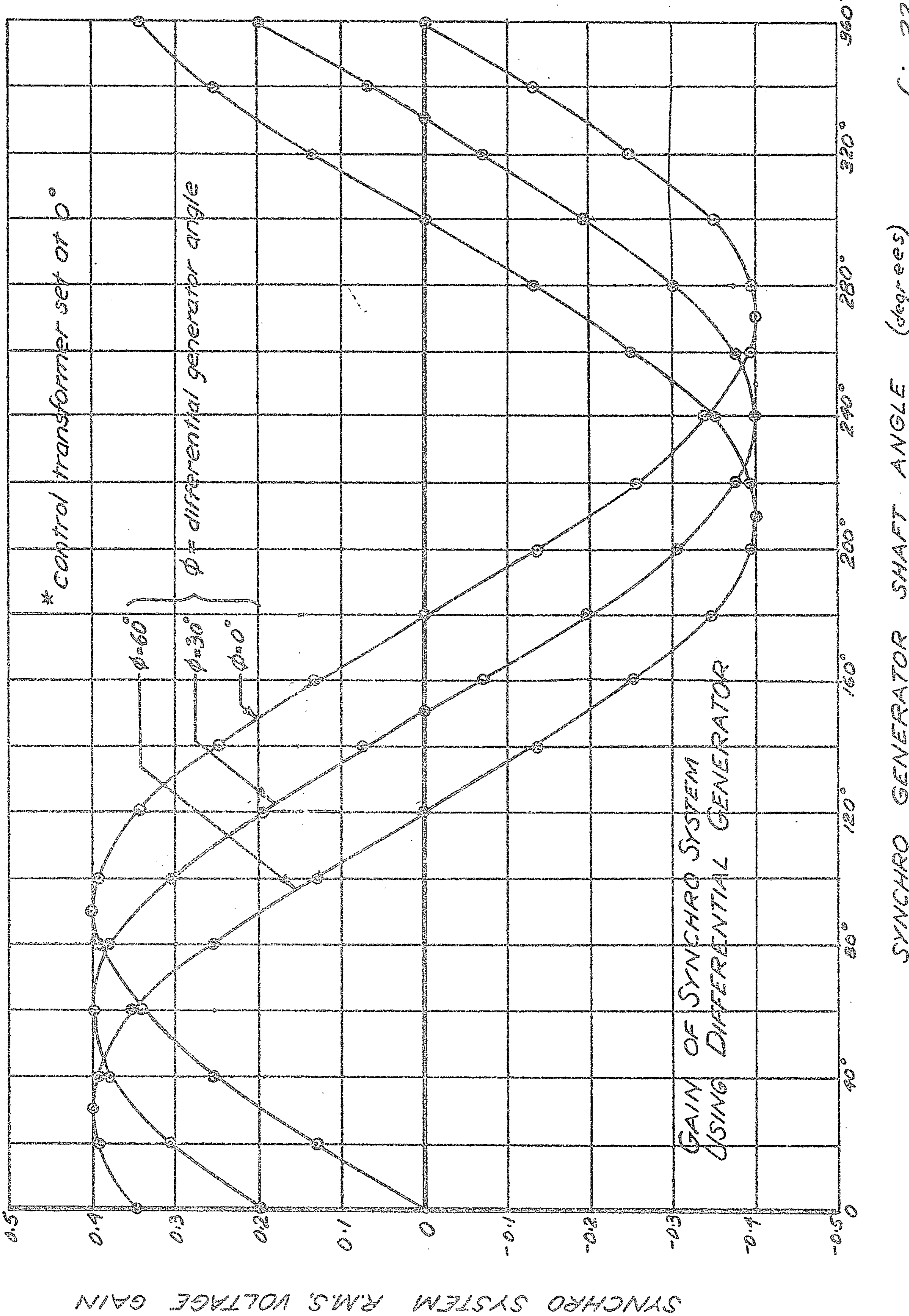

*control transformer set at 0°
φ = differential generator angle
φ=60°
φ=30°
φ=0°
GAIN OF SYNCHRO SYSTEM
USING DIFFERENTIAL GENERATOR
0.5
0.4
0.3
0.2
0.1
0
-0.1
-0.2
-0.3
-0.4
-0.5
0
40°
80°
120°
160°
200°
240°
280°
320°
360°
SYNCHRO GENERATOR SHAFT ANGLE (degrees)
SYNCHRO SYSTEM R.M.S. VOLTAGE GAIN

fig. 32

Velocity Detection

From the previous theoretical work it was known that continuous rotation of the synchro generator rotor with respect to the control transformer rotor would produce an output signal from the control transformer which would have the form of an amplitude modulated wave with the frequency of modulation being directly related to the relative rotation speed of the rotors. A similar situation occurs when the basic system is enlarged to include a differential generator. If the rotors of the synchro generator and control transformer are kept stationary and if the rotor of the differential generator is driven in continuous rotation the output from the control transformer will be the same as for the first case discussed above. This last situation has not been worked out analytically here but can be done without too much difficulty in a manner similar to that used in analyzing the operation of the basic pair in continuous rotation. The result was verified experimentally however.

The original interest in pursuing this line of investigation was to determine the type of characteristics that could be obtained from a system in which one synchro, the synchro generator, was used as a position indicator so that its error signal would be proportional to its relative shaft position, and a second synchro, the differential generator,

........../132 a.

was used as a velocity indicator, that is, its effect on the overall error signal would be proportional to the velocity of its rotor. Thus the output of the control transformer in this three-unit system would be relative to the position of the synchro generator shaft and also to the angular velocity of the differential generator shaft. As was stated above this was the original intent in pursuing this line of reasoning and was simply an idea which seemed worthy of further investigation. It was soon found, through experimental procedures, that such a synchro system did not produce the type of result as originally postulated. Checking back on the analytical development for the effect of the differential generator shows why the reasoning was invalid. For a synchro system consisting of a synchro generator, a differential generator, and a control transformer, the voltage output from the control transformer rotor is given by:

$$E_{out} = E_m \cdot \sin(\theta - \theta' \pm \phi) \cdot \sin \omega_c \cdot t$$

where $\theta - \theta'$ represents the relative shaft positions of the synchro generator and control transformer, and ϕ is the angle of the shaft position of the differential generator with respect to a fixed reference position. With continuous rotation of either the rotor of the synchro generator or the rotor of the differential generator, the following result will occur:

$$(\theta - \theta' \pm \phi) = (\theta_o + \omega_m \cdot t)$$

......../132 b.

where θ_o represents the sum of $(\theta - \theta' \pm \dot{\phi})$ at $t = o$ and ω_m is the angular shaft velocity of the continuously rotating rotor. Thus, regardless of which rotor is continuously rotating, the resultant output from the control transformer will appear as a carrier at frequency ω_c modulated by a frequency ω_m. The position of the non-continuously rotating shaft simply determines the value of θ_o, that is the resultant angle at $t = o$. Thus the original desire for a system which would produce an error signal which would depend on the velocity of one shaft and the static position of another did not materialize but it did serve a worthwhile purpose in that its investigation did lead to certain other interesting possibilities. It might be wise to interject at this point that this dissertation is intended solely as an investigation of the characteristics of synchro devices and consequently the intended applications for some of the characteristics described will not be too greatly elaborated upon.

As noted above, the investigation of the original intent did lead to other paths of thoughts which seem more likely to produce more useful results. By experimental observation it was found that continuous rotation of both the synchro generator and the differential generator produced an interesting effect. If the two units were driven at the

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same angular speed but in opposite directions of rotation, the output signal from the control transformer was an unmodulated constant ampltitude carrier frequency signal. The fact that the direction of rotation of the rotors is opposite is solely attributed to by the construction of the differential generator used in these tests. Returning to the equation for transformer rms output voltage,

$$E_{out} = E_m \sin (\theta - \theta' \pm \phi)$$

shows the system to be operating and producing an error signal exactly the same as the basic synchro pair produces. Considering the angle of control transformer rotor position θ' to be at 0°,

$$E_{out} = E_m \sin (\theta \pm \phi)$$

Now the situation just described obviously represents the case where $(\theta \pm \phi)$ remains constant and hence the output voltage remains constant. If the rotors of the two generators begin to rotate at different speeds while the control transformer rotor remains fixed, the output will be a carrier frequency signal modulated at a frequency corresponding to the difference in angular velocities of the two generator rotors. If all three rotors are continuously rotated the output will be a function of all three rotor velocities but will still appear as a modulated wave. To outline this reconsider the original output voltage equation for this system:

$$E_{out} = E_m \sin (\theta - \theta' \pm \phi) \sin \omega_c \cdot t$$

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Let $(\theta \pm \phi) = \omega_o t$ and $\theta' = \omega_r \cdot t$. Also let θ_o be the sum of $(\theta - \theta' \pm \phi)$ at $t = o$.

Therefore

$$E_{out} = E_m \sin (\omega_o \cdot t - \omega_r \cdot t + \theta_o) \cdot \sin \omega_c \cdot t$$

If the time reference is chosen so that $\theta_o = 0$,

$$E_{out} = E_m \sin (\omega_o - \omega_r) t \quad . \sin \omega_c \cdot t$$

Thus the frequency of modulation will be proportional to the difference between the relative angular speed of two generator shafts (ω_o) and a reference speed (ω_r). This therefore presents the possibility for the control of the difference in speed between two shafts, one connected to the synchro generator and the other connected to the differential generator. The reference speed is of course obtained from the control transformer drive. It would also be possible to switch the drives connected to any of the units and still produce the same result. To make use of this characteristic however a unit must be included which would produce a useable signal proportional to the modulating frequency for it is the modulating frequency which indicates the error.

Further Experimental Work with Continuous Rotation of the Generator Rotor

The previous work with the basic synchro pair involving continuous rotation of the generator rotor brought up a point which, although it belongs more under the heading of general modulation theory, does have some relation to the topic of this dissertation. With the generator rotor of a basic pair running in continuous rotation relative to the control transformer rotor, and with the output signal from the control transformer applied to the input of the discriminator, the discriminator output has the form of the wave shown in fig. 33. This waveform is easily accounted for in terms of the action of the discriminator which provides a full-wave rectified positive output for one half of the full revolution of the relative synchro shaft position angle ($\theta - \theta'$) and a full-wave rectified negative output for the other half of the revolution. Since the rotors make one complete revolution for every period corresponding to the continuous rotation frequency the discriminator output will alternate from positive to negative along the time axis. The situation shown in fig. 33 has the number of carrier frequency half-waves an integer multiple of the modulating frequency. This has been done simply to facilitate drawing and is not necessarily the general case. This means that the position of the carrier relative to the envelope wave

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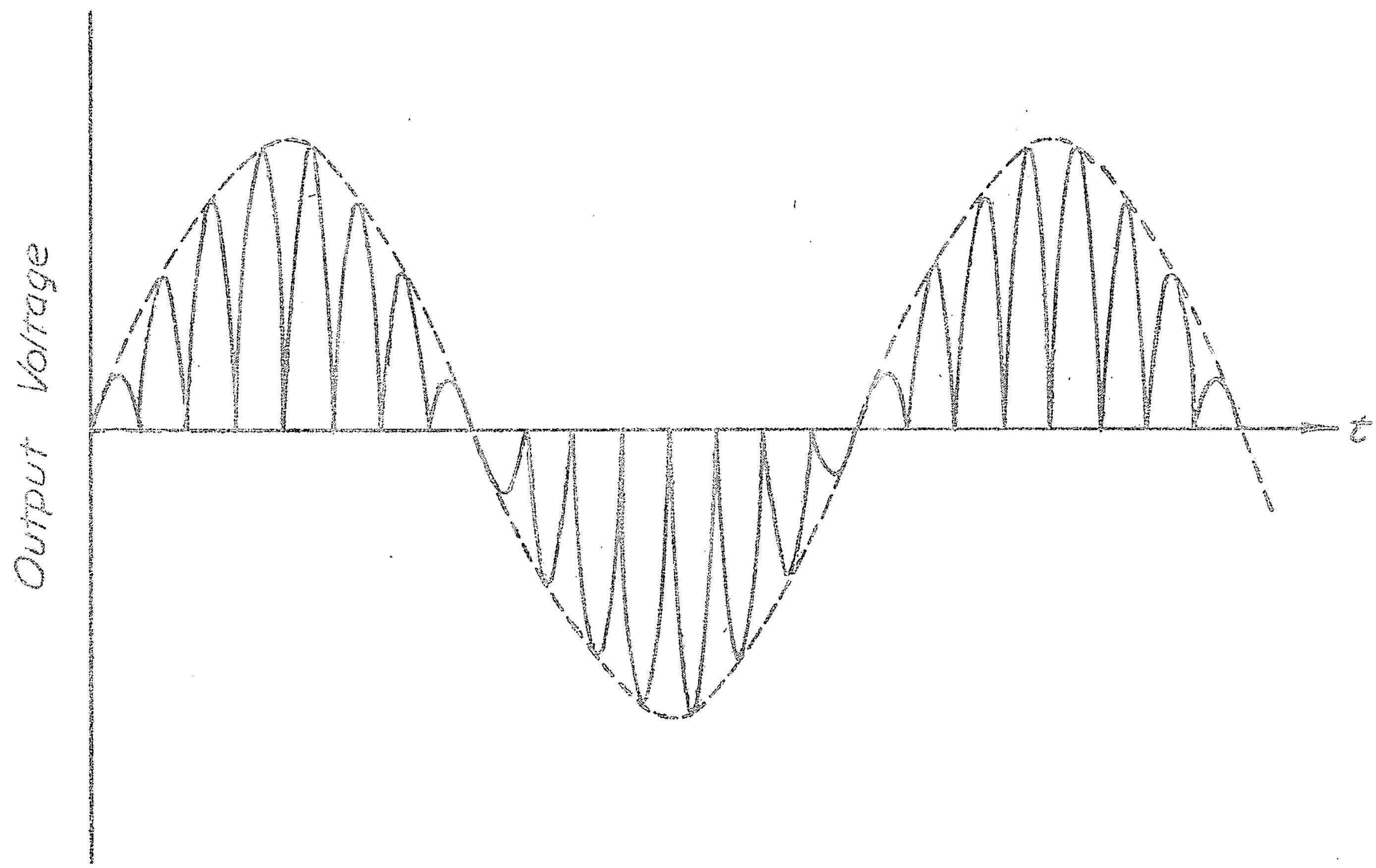

DISCRIMINATOR OUTPUT for CONTINUOUS ROTATION of GENERATOR ROTOR

fig. 33

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could change from one cycle of the envelope wave to the next. This makes analytical treatment of the waveform not impossible but very awkward and hence experimental determinations become more practical. The problem mentioned at the beginning of this item is to determine what relation if any the d.c. component of this waveform, over a time base of one half-period of the modulating frequency, has to the modulating frequency. Obviously if the carrier frequency is much greater than the modulating frequency the d.c. component of a complete cycle of the entire wave will be zero and hence any attempt to measure the d.c. component by using the output of the discriminator would yield no useful results. If however a simple full-wave rectifier is used instead of the discriminator the output will appear as in fig. 34 and useful measurements will be possible.

The simple full-wave rectifier circuit shown in fig. 35 was built and the output of the control transformer was fed into it. A d.c. voltmeter was placed across R_L to measure the d.c. voltage component in the output. For very low modulating frequencies no results were obtainable as the meter tended to follow the instantaneous d.c. level and hence no steady reading was obtained. As the modulating frequency was increased by increasing generator rotor speed of rotation the meter tended to measure the average d.c. level of the cycle. Once the modulating frequency had passed the point

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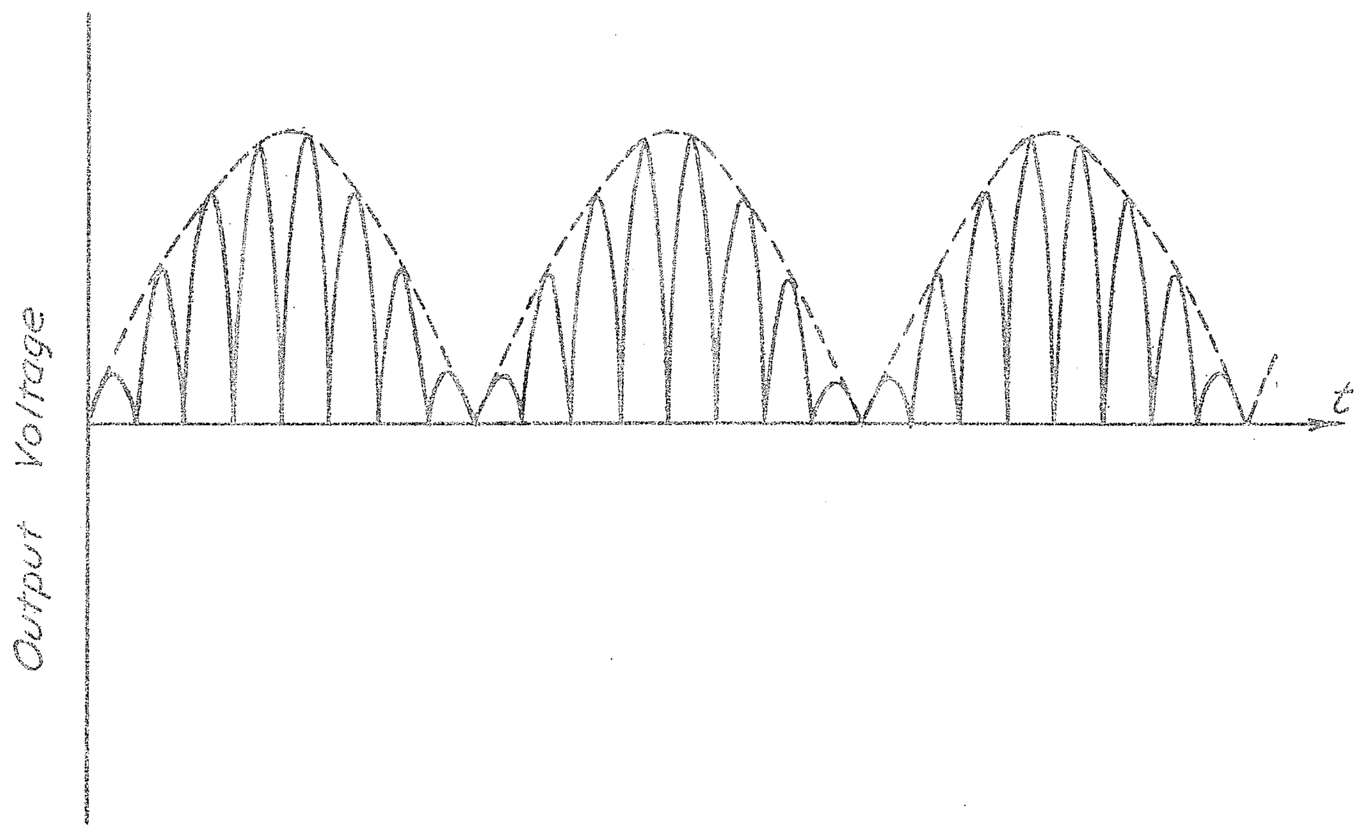

FULL-WAVE RECTIFIER OUTPUT for CONTINUOUS ROTATION of GENERATOR ROTOR

fig. 34

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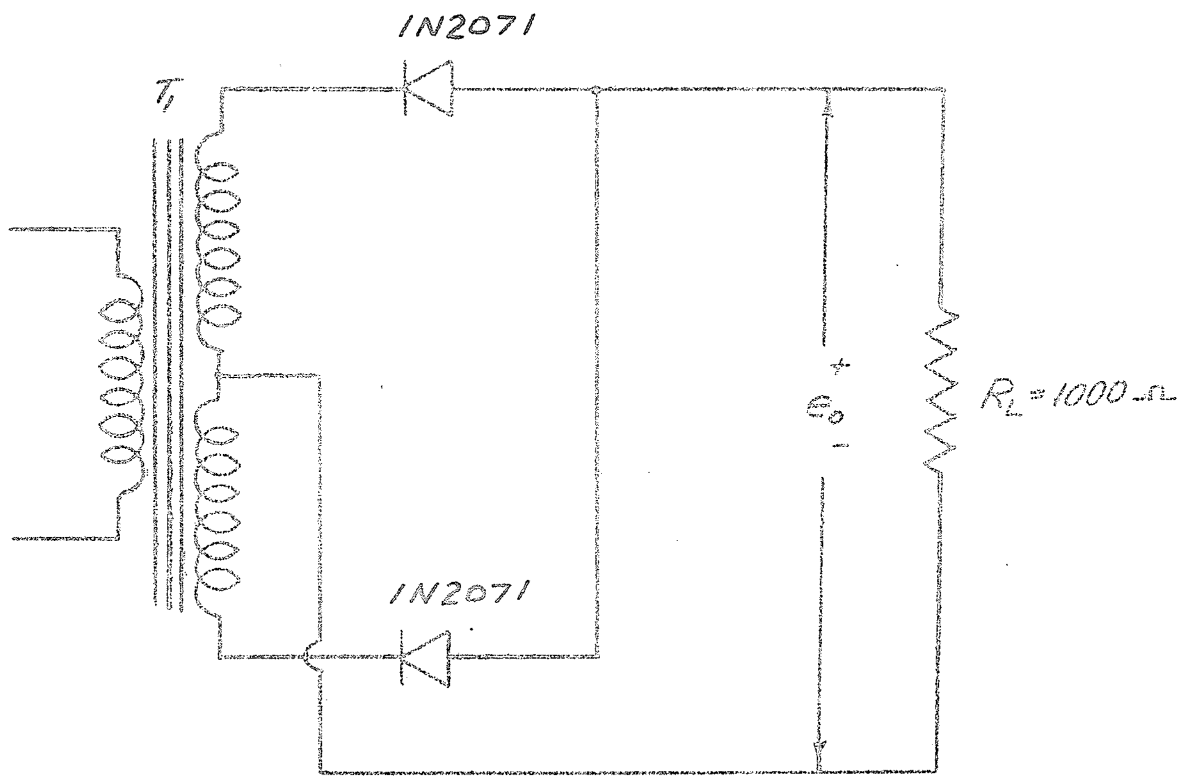

T_1 - Hammond Transformer #51

FULL-WAVE RECTIFIER CIRCUIT

fig. 35

where a steady meter indication was obtained the d.c. level indicated remained relatively independent of the modulating frequency. Some minute change in the level might have been present but this change was so small that no possible application could have been made of it with respect to a control system. This of course is the result which this test was designed to determine and although the answer to the problem turned out to be much simpler than was anticipated it never the less does provide an answer to the original question.

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CHAPTER 10

THE HARMONIC RESPONSE FUNCTION OF THE BASIC SYNCHRO SYSTEM

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The harmonic response function of the basic synchro system (fig. 36) relates the output of the output filter to a sinusoidal shaft position input to the rotor of one of the synchros. For convenience of description it will be assumed here that the rotor of the control transformer will remain fixed in position and the generator rotor only will be rotated. This is an arbitrary choice for descriptive purposes only as it is immaterial whether the generator rotor or control transformer rotor rotates or whether they both rotate as long as the relative motion is sinusoidal. The effect of the output filter will also be considered here as this is necessary to obtain the final single frequency sinusoidal output signal.

The diagram of the basic synchro pair is indicated schematically in fig. 37. It will be assumed in this analysis that the windings of the synchro stators are symmetrically located about the periphery of the stator and that the flux fields produced by currents in the windings of the stators can be represented by vectors spaced at 120° with respect to each other. As in any transformer equivalent circuit, the transformer secondary can be approximately represented by an equivalent generator in series with the internal impedance of the windings such as is shown in fig. 38.

The instantaneous voltages produced by the equivalent voltage generators are:-

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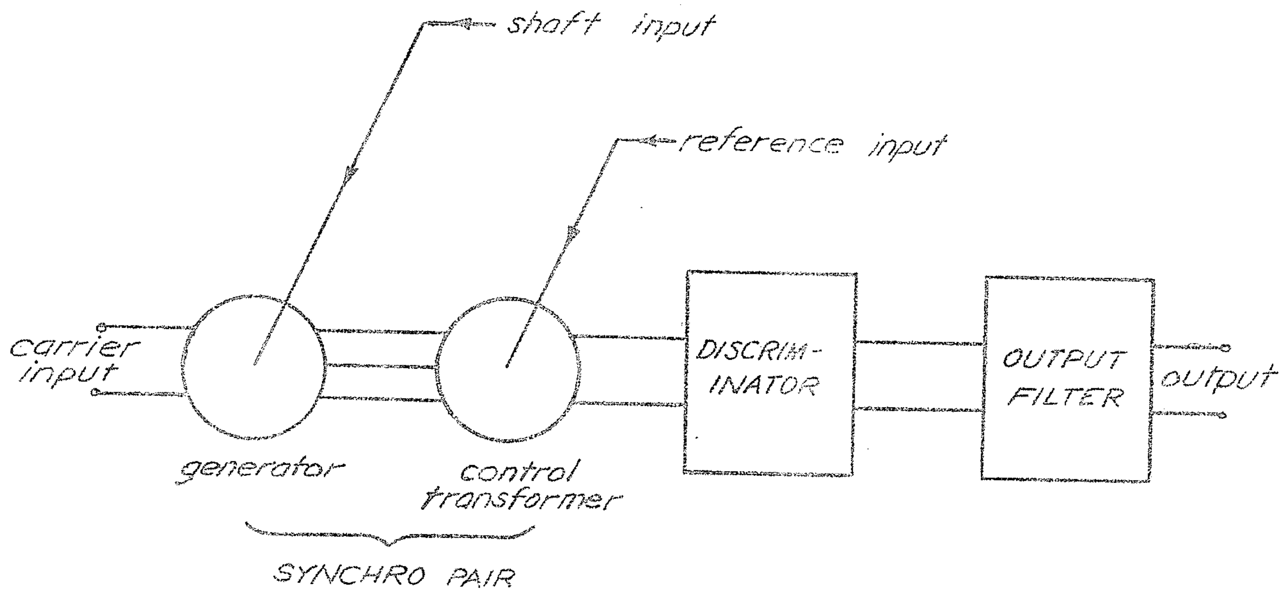
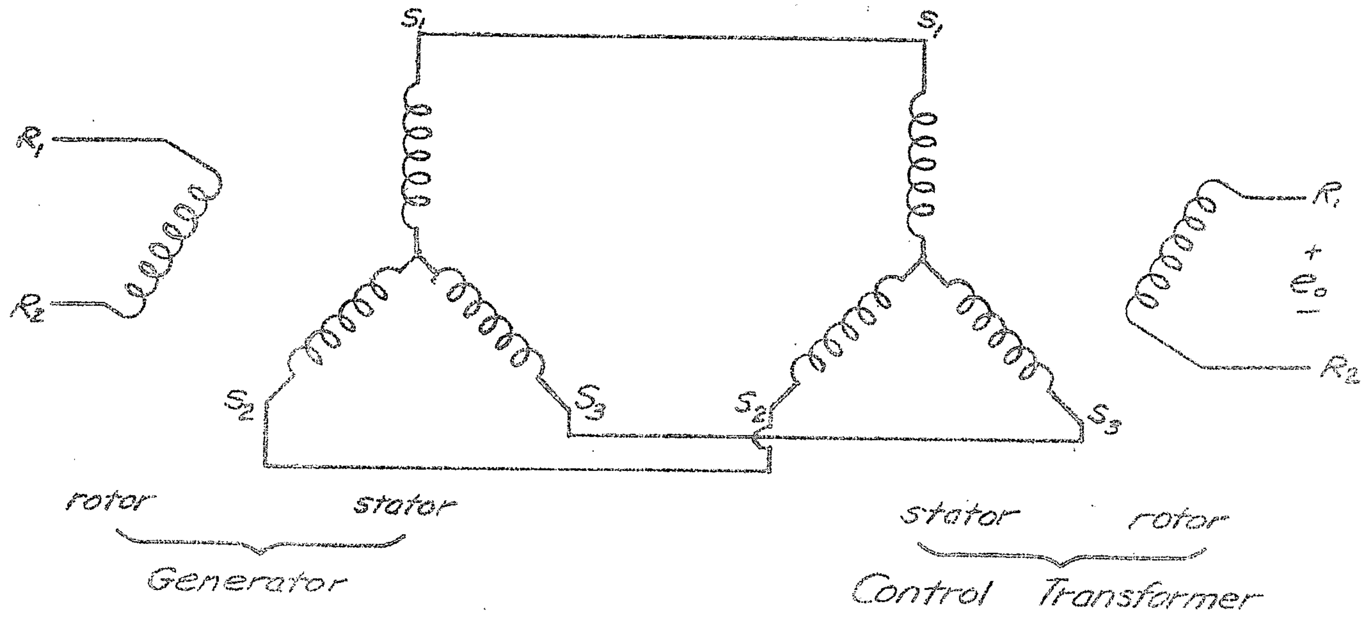


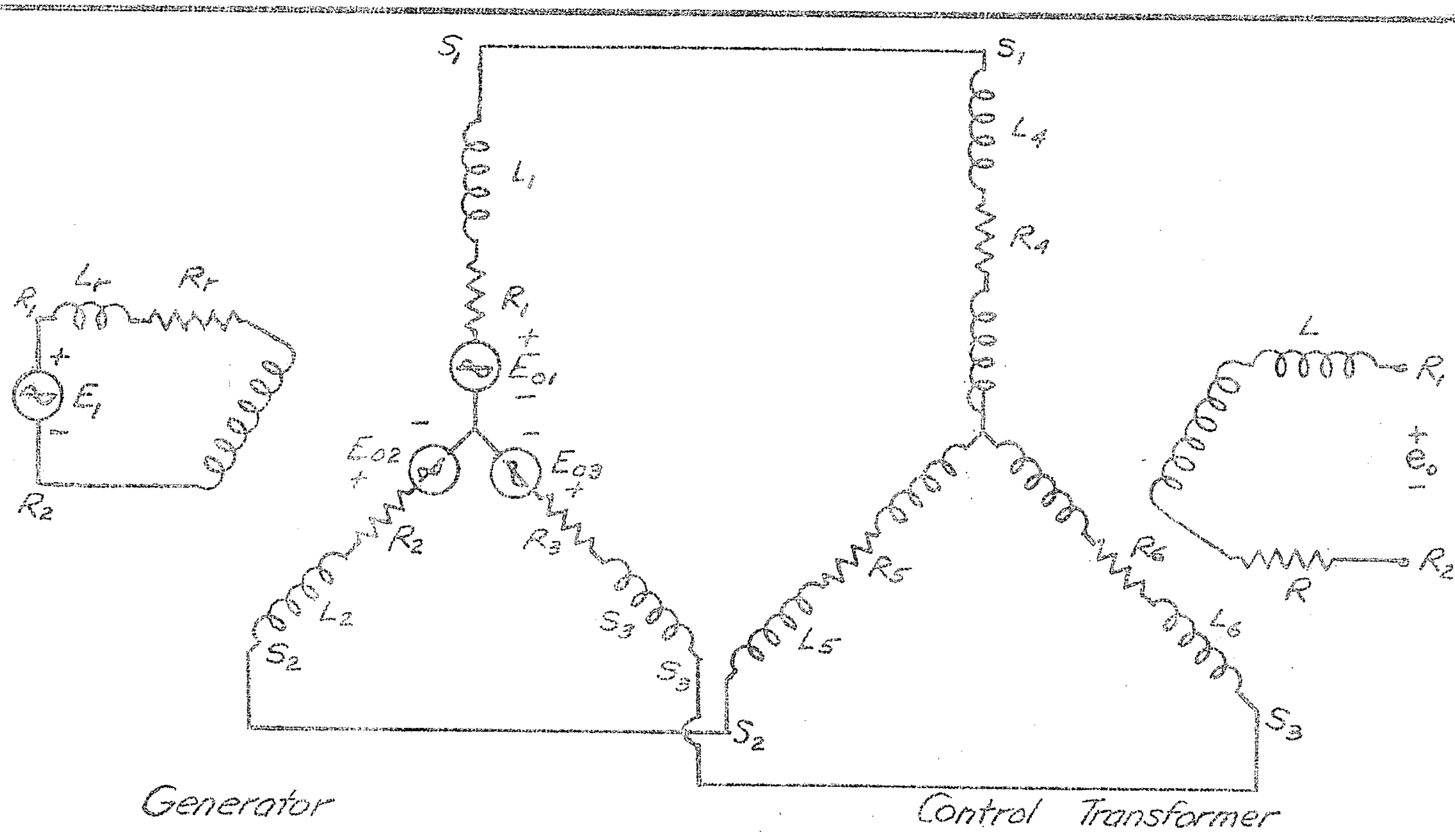
BASIC SYNCHRO SYSTEM

fig. 36



BASIC SYNCHRO PAIR

fig. 37



APPROXIMATE EQUIVALENT CIRCUIT of the BASIC SYNCHRO PAIR

fig. 38

$$E_{01} = E_1 \cdot K_1 \cdot \cos\theta \cdot \sin(\omega_c \cdot t + \psi_1)$$

$$E_{02} = E_1 \cdot K_2 \cdot \cos(\theta - 120^\circ) \cdot \sin(\omega_c \cdot t + \psi_2)$$

$$E_{03} = E_1 \cdot K_3 \cdot \cos(\theta - 240^\circ) \cdot \sin(\omega_c \cdot t + \psi_3)$$

where K_1, K_2, and K_3 are attenuation factors to account for voltage drops due to leakage flux core losses, and variation in turns ratios between primary and secondary in the units, and ψ_1, ψ_2, and ψ_3 represent the phase shifts occurring between E_1, and the equivalent circuit generator voltages. The attenuation factors and phase shifts should be very nearly the same although for an exact analysis it must be remembered that unbalances can exist between them. As discussed in a previous section, these attenuation factors might vary for different positions of the rotor since reluctance paths can change as the rotor is moved. It will be assumed here though that these unbalances and changes are negligible.

To determine the harmonic response function from angular shaft displacement to output voltage of the filter, it is necessary to determine the gain in terms of magnitude and phase shift. The magnitude of the gain from shaft input to discriminator output is easily measured by experimentally measuring the d.c. output component from the discriminator for a finite displacement between the generator and control transformer rotors. The test must be conducted in such a manner however that the maximum angular displacement

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of the shaft will be sufficiently small that approximately linear operation of the synchro pair will occur. The gain of the output filter can be measured separately. To arrive at the gain magnitude expression analytically necessitates knowledge of the various attenuation factors involved, the values of the internal resistances, self and mutual inductances of the coils, as well as other factors. These could all be measured directly or deduced by a series of measurements, however to determine all of these factors would involve much more effort than to simply measure the gain magnitude straight away and therefore it is probably wiser and more accurate to measure the gain magnitude directly rather than attempt an analytic approach. If however it is desired to determine the effects caused by variation of coil impedances, or some other change, analytic development might be worthwhile.

To determine the phase shift presents a slightly more complex problem. Experimentally this could be determined by oscillating the rotor of the generator sinusoidally, ensuring that the amplitude of oscillation does not exceed the limits necessary to produce approximately linear operation, and comparing the phase of the signal from the output filter with the phase of the rotor position wave. This necessitates construction of a device to impart sinusoidal oscillation to the generator shaft and hence if some other method of evaluating the phase shift could be employed it might be preferable.

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It is an alternative method to this direct procedure for phase measurement that this particular section of the work is intended to develop.

Indirect Method of Obtaining Synchro System Phase Shift

Referring back to the previous analytical development for continuous oscillation of the generator rotor it is seen that the voltages of the circuit will have the form of a carrier signal modulated sinusoidally at a frequency equal to the frequency of oscillation of the generator rotor. The amplitude of the modulation will be dependent on the degree of maximum shaft displacement from the center of oscillation. It is now assumed that the modulating effect on the carrier will be in time phase with the sinusoidal oscillation of the generator rotor. Since there is no time delay here there will also be no phase shift from shaft oscillation to modulation effect of the voltages E_{01}, E_{02}, and E_{03} of the equivalent circuit of fig. 38. All of these voltages will be in time phase since they are all produced from the same excitation and since it has been assumed that there will be no time delay between the shaft oscillation and the resulting effect produced in the synchro generator secondary coils. To provide some justification for this assumption consider the picture of what is physically happening in the generator. As the rotor turns through an arc $\Delta\theta$ the magnetic field

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created by the currents flowing in the rotor winding automatically, and with no lag, follows the motion of the rotor. This shift in position of the rotor flux field is instantaneously reproduced in effect in the amplitude modulation of the voltages of the equivalent circuit generators since there is no device present to create delay. The voltages of each of the generators of the equivalent circuit will now have the waveform of a sine-wave modulated carrier with the variation in carrier amplitude being in phase with the rotor position waveform. From the analysis of the synchro pair for continuous oscillation of the generator rotor the equivalent circuit voltage generators will have an instantaneous output voltage expressed by:

$$e_{out} = E(\sin \omega_s \cdot t)(\sin \omega_c \cdot t)$$

for oscillation about the correspondence point and for small oscillation amplitudes. Applying the appropriate trigonometric relation yields:

$$e = \frac{E}{2}\left[\cos(\omega_c - \omega_s) - \cos(\omega_c + \omega_s)\right]$$

showing that only the two sidebands exist and no carrier signal which is the form of a suppressed carrier system. If the oscillation had not been centered at the correspondence point a term of the carrier frequency would also be present. The equivalent generator voltages will produce currents which will consequently set up a flux field in the control transformer thus producing a rotor output voltage of the form of

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a modulated carrier. If the sideband frequencies are not too far removed from the carrier frequency the phase shift from the equivalent generators to the output of the control transformer encountered by each sideband frequency component will not differ appreciably from the phase shift that would be encountered by a carrier frequency signal. It will be assumed here that the sideband frequencies do not differ greatly from the original carrier frequency so that this assumption regarding phase shift will be valid. For some idea of the worth of this assumption consider the case of a 400 cycle carrier frequency system. This frequency corresponds to a radian freq. of 2512 radians/second. In normal control systems, oscillation frequencies would normally not be exceptionally high and probably at the most would not exceed 2000 oscillations/minute which corresponds to 209.1 radians/second. This is still less than 10% of the carrier frequency hence the reactance effects on the sidebands would not be appreciably different from the same effect on a carrier frequency signal.

The output signal from the control transformer is fed into the discriminator ~~whose~~ the output of which then goes into the output filter. The filter is designed to pass signals in the range from d.c. to the highest expected frequency of oscillation and to attenuate all higher frequencies. By the action of the discriminator the waveform being fed into the input of the filter will contain a component corresponding to the frequency of oscillation of the generator rotor and it is

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this component that will be passed by the filter. Since the reactances of the synchros and the filter will cause some phase shift of the signal information which they are transferring, the output signal from the output filter will not be in phase with the oscillation of the generator rotor. It is the magnitude of this phase shift which it is desired.

From the equivalent circuit generators to the output of the control transformer all information content is transmitted in amplitude modulated carrier form in a wave consisting solely of the upper and lower sidebands, if oscillation occurs about the correspondence point, or a wave containing also a carrier frequency component if oscillation is about a point slightly removed from the correspondence position.

Consider the effect of a reactive network with harmonic response function denoted by $H(j\omega)$, on an amplitude modulated wave. It can be shown that* the following relationships depict the time axis "delays" of the carrier component and the envelope outline of an amplitude modulated wave due to the effects of $H(j\omega)$.

$$\text{carrier delay} = \left. \frac{-\operatorname{Arg} H(j\omega)}{\omega} \right|_{\omega = \omega_c}$$

$$\text{envelope delay} = - \left. \frac{d\operatorname{Arg} H(j\omega)}{d\omega} \right|_{\omega = \omega_c}$$

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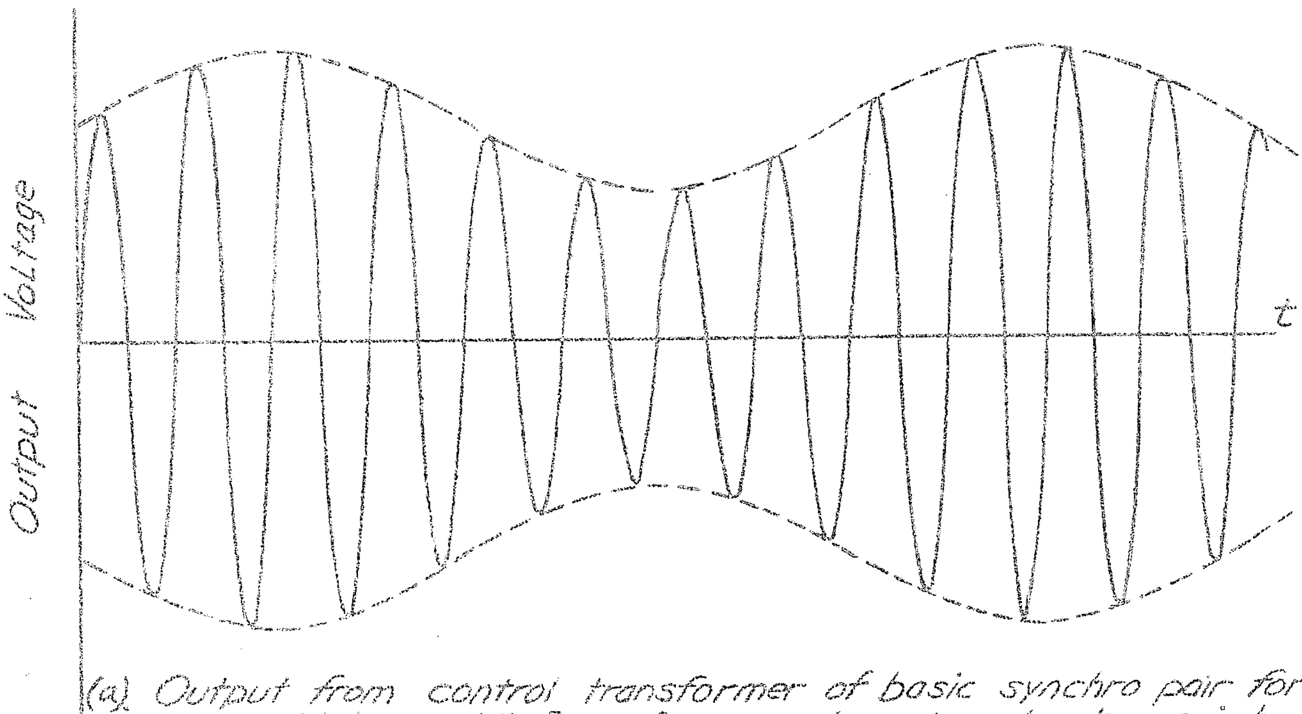
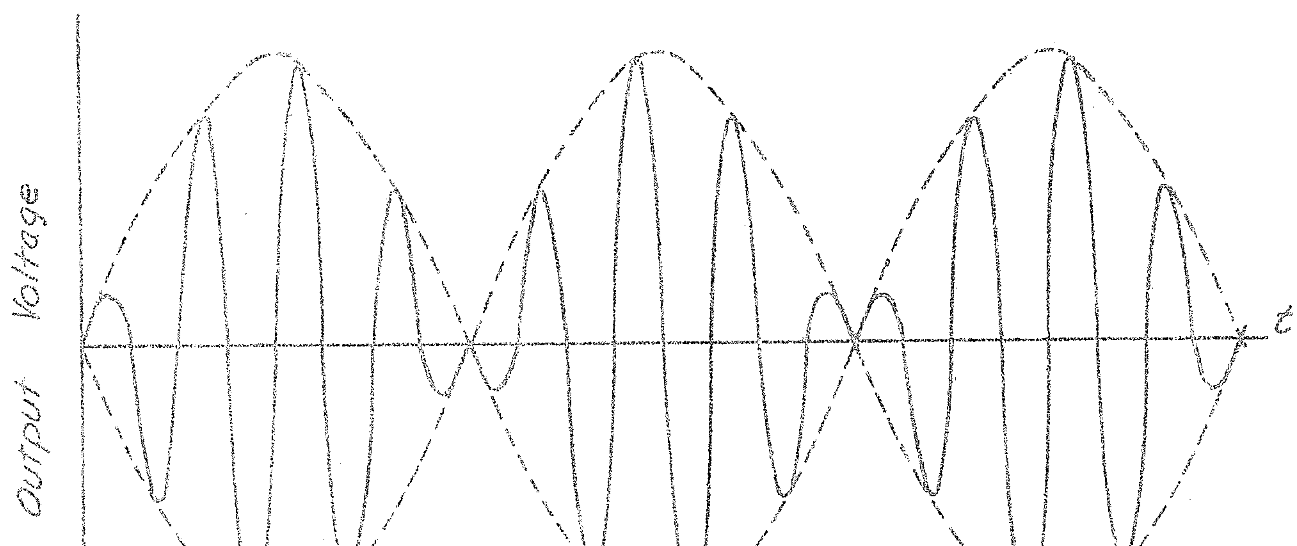
* Mason and Zimmerman - "Electronic Circuits, Signals, and Systems" (p. 367)



(a). Output from control transformer of basic synchro pair for sinusoidal oscillation of generator rotor about a point away from the correspondence point. ($\Delta\Theta > \sigma_m$)



(b). Output from control transformer of basic synchro pair for sinusoidal oscillation of generator rotor about the correspondence point.

fig. 39

Assume $H(j\omega)$ has the form:

$$H(j\omega) = \frac{A_o}{1+j\dfrac{\omega}{\omega_o}}$$

therefore carrier delay, t_c, $= \dfrac{-1}{\omega_c} \cdot \tan^{-1} \dfrac{\omega_c}{\omega_o}$

Now, if ω_o is much larger than ω_c, such as would the case in a normal synchro system, t_c can be reduced to:

$$t_c \doteq \frac{-1}{\omega_o}$$

and $\quad \theta_c = \omega_c \cdot t_c \doteq -\dfrac{\omega_c}{\omega_o}$.

envelope delay, $t_s = \dfrac{-d}{d\omega} \tan^{-1} \left.\dfrac{\omega}{\omega_o}\right|_{\omega=\omega_c}$ -

$$\theta_s = t_s \cdot \omega_s = -\frac{\omega_o \cdot \omega_s}{\omega_c^2 + \omega_o^2} \doteq -\frac{\omega_s}{\omega_o} \quad \text{for } \omega_o \gg \omega_c$$

where ω_s is the modulating frequency. If these two expressions for phase shift are now written as a ratio the following expression is obtained:

$$\theta_m \doteq \frac{\omega_s}{\omega_c} \cdot \theta_c .$$

The action of the discriminator, including the output filter, is to select from the output of the control transformer a component at frequency ω_s. If the action of the discriminator and filter is considered quite closely it will be seen that the component which is selected will be equivalent, except for amplitude, to the signal which would be obtained if the envelope of the waveform had been detected. Hence θ_s, the angular phase shift in the envelope wave from the equivalent circuit generators to the output of the control transformer rotor, will be the contribution to the total phase shift of the signal at frequency ω_s, between the sinusoidal rotor position input and the output of the filter, due to the reactances between the equivalent circuit generators and the output of the control transformer.

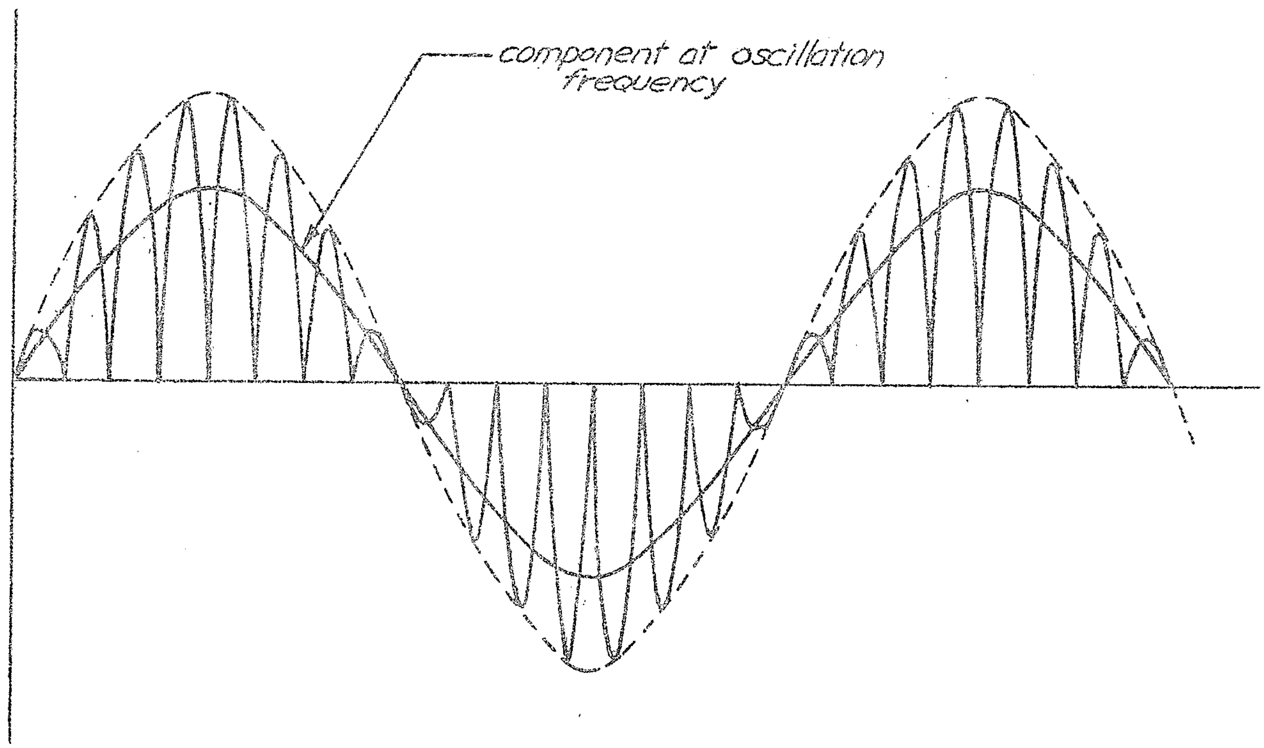
Phase Shift Introduced by the Discriminator

Considerable effort was put forth in the section dealing with the theoretical operation of the discriminator to develop two expressions which will be quite valuable at this point. These are:

$$v_o \doteq \frac{G}{\pi} \cdot \cos\beta \cdot F(t) \qquad \text{and}$$

$$v_o \doteq \frac{2G}{\pi} \cdot \cos\beta \cdot F(t)$$

which are the expressions for the useful output voltages from the half-wave and full-wave discriminators respectively. The



DISCRIMINATOR OUTPUT for CONTINUOUS SINUSOIDAL OSCILLATION of GENERATOR ROTOR SHOWING the COMPONENT at the OSCILLATION FREQUENCY
(OSCILLATION OCCURRING ABOUT the CORRESPONDENCE POINT)

fig. 40

input to the discriminator is assumed to be an amplitude-modulated wave with F(t) being the modulating function, that is, the input to the discriminator is given by:

$$v_i(t) = F(t) \sin (\omega_c \cdot t - \beta)$$

where ω_c is the carrier frequency and β is the phase shift occurring between the signal input to the discriminator and the reference input. From this it is noted that the original modulating function F(t) emerges from the discriminator multiplied only by a constant without any phase lags or amplitude changes depending on frequency. It is concluded from this therefore that the process of demodulation theoretically does not introduce any phase lags (or leads) into the signal channel. In practice however the signal input transformer will introduce some small shift into the wave components due to its internal reactances. The amount of this shift however should be quite small since the load impedance across the secondary of the transformer will be quite high. Following the same reasoning as used with regard to the shift due to the reactances of the synchro pair, the phase shift introduced between the output of the filter and the rotor position waveform will be given by:

$$\theta_{c_T} \cdot \frac{\omega_s}{\omega_c}$$

where θ_{c_T} is the phase shift of the carrier frequency signal across the input transformer. θ_{c_T} as explained above will never exceed more than a few degrees.

Phase Shift Across the Output Filter

The shift introduced into the useful signal by the output filter is simply the phase shift of the filter at frequency ω_s. This can be measured directly and will probably be found to far exceed the phase shifts introduced by the other units of the system.

Total Phase Shift

In the previous discussion it was shown that there was no phase shift between the waveform of actual shaft oscillation and the modulating component effecting modulation of the equivalent generators waveforms. Applying this then, the total phase shift from rotor shaft oscillation to discriminator output will be equal to the carrier frequency phase shift between the equivalent generators and the output of the control transformer, multiplied by the ω_s/ω_c factor, the carrier frequency phase shift across the discriminator input transformer, again multiplied by the ω_s/ω_c factor, plus the phase shift at frequency ω_s occurring across the output filter.

$$\theta_{s\ \text{total}} = \left[\left(\theta_c \ \begin{matrix}\text{(from equivalent generators to output} \\ \text{of control transformer)}\end{matrix}\right) \cdot \frac{\omega_s}{\omega_c}\right] + \left[\begin{matrix}\text{carrier frequency signal phase shift} \\ \text{across discriminator input transformer}\end{matrix} \cdot \frac{\omega_s}{\omega_c}\right] + \left[\begin{matrix}\text{oscillation frequency signal phase} \\ \text{shift across output filter}\end{matrix}\right] \; . \; .$$

The problem is now to measure the carrier frequency signal phase-shift from the equivalent generators to output of control transformer. The total phase shift of a carrier frequency signal from generator rotor input to control transformer output is a straight forward measurement. If the phase shift from generator input to the equivalent generator waveform is know it can be subtracted from the overall phase shift to yield the desired angle.

Measurement of Carrier Phase Shift

The measurement of the phase shift from synchro generator input to control transformer output presents nothing special or new. The phase measurement from generator input to equivalent circuit generator does present some problems. If it is assumed that the attenuation factors and phase shifts between the rotor of the generator and each of the stator coils which form the secondary of the transformer are equal the problem of measuring these factors is reduced. The following equations show the relation between the input voltage to the generator rotor and the secondary coil voltages.

$$e_{os_1}(t) = E_1 \cdot K_1 \cdot \cos(\theta) \cdot \sin(\omega_c \cdot t + \psi_1)$$

$$e_{os_2}(t) = E_1 \cdot K_2 \cdot \cos(\theta - 120^\circ) \cdot \sin(\omega_c \cdot t + \psi_2)$$

$$e_{os_3}(t) = E_1 \cdot K_3 \cdot \cos(\theta - 240^\circ) \cdot \sin(\omega_c \cdot t + \psi_3)$$

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In terms of these equations equal attenuation factors would mean K_1, K_2, and K_3 all equal and similarly equal phase shifts between rotor and stator coils would mean ψ_1, ψ_2, and ψ_3 would all have the same value. Since, in the construction of the synchros great care is taken to place the coils symmetrically about the stator and to make their physical characteristics as nearly equal as possible these attenuations and phase-shift factors usually have only negligible differences but assurance of this for every precise analysis cannot be assumed until some measurements are actually made. For all normal intents and purposes though they can be assumed equal as was done in the development of the basic theory of the synchro system earlier in this dissertation. If the connection at the center of the Y-connected secondary is accessible, the measurement of the phase-shift between the rotor and each of the windings individually is possible. If not, the deduction of the phase-shift between rotor voltage and secondary coil voltage must be done using the results of a series of measurements made using the three accessible secondary coil leads only. Regardless of which method is used though, the measurements must be ~~done~~ made so as to leave the secondary effectively open circuited so as not to allow the internal impedances of the windings to effect the measurements since it is the shift from generator rotor input to equivalent circuit generator only which is desired.

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If it is found that the phase angles (ψ) do have slightly different values a difficulty will exist in choosing the angle which properly represents the desired phase relation for the equivalent circuit generators since there will not be three corresponding phase angles to measure at the output of the control transformer. It is therefore necessary to take an average value of these angles as the appropriate phase shift. Further complexity in this matter is not deemed necessary due to the small magnitudes of the errors that will probably be involved.

Experimental Determination of a Harmonic Response Function

In an attempt to verify some of the techniques for determining the phase shift involved in the harmonic response function of the basic synchro pair certain experiments were performed involving both the direct and indirect techniques. The following sections outline the types of tests made, their results, and the meaning of these results in relation to the developed theory. Phase shift measurements were made in these tests with an Advance Electronics Co. Vectorlyzer (type 202).

The Effect of the Synchro Pair on a Carrier Frequency Signal

Measurements were made using a basic synchro pair consisting of a generator and a control transformer. The specifications of the units used are:

General Electro Synchro Generator
Model 2J1F1 Serial No. 474886
Primary voltage 115 v.
Secondary voltage 57·5 v.
400 cycle.

General Electric Synchro Control Transformer
Model 2J1G1 Serial No. 6374555
Stator line-to-line voltage 57·5 v.
400 cycle.

The synchro generator was modified slightly by bringing out a sixth lead from the unit. This lead was from the neutral point of the Y-connected stator windings and was used to make single coil measurements on the secondary of the generator. The Vectorlyzer was used to measure the phase shift from the generator input to each coil of the secondary with the coils open-circuited. It was found that the amount of this shift was very small and quite negligible with respect to the shift occurring in the rest of the synchro. This means that to all intents and purposes the generators of the equivalent circuit (fig. 38 page 144) will be in phase with the carrier input to the synchro generator rotor. The carrier frequency signal phase shift from the equivalent circuit generators to the output of the control transformer will then be equal to the shift from the carrier input to the generator rotor to the output of the control transformer. The amount of this phase shift was measured at 3° for the synchro pair used in these tests.

Discriminator Phase Shift

The phase shift of a 400 cycle signal across the input transformer of the ring discriminator was found to be quite negligible and hence the total phase shift across the discriminator was assumed to be zero.

Output Filter Tests

The band-rejection, low-pass filter combination described previously and used in some of the test work here did suffice for most purposes since it eliminated to an adequate degree most of the harmonic content in the output of the discriminator, however it still left some unwanted ripple in the output which was especially undesirable for measurement of phase shifts. The situation could be improved by use of a more elegant filter design using the same band-rejection, low-pass combination with possibly ~~and~~ additional rejection filter to eliminate the fundamental frequency ripple

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component. In the determination of the synchro pair phase shift though it is not absolutely necessary to retain a wide useful signal bandwidth since only a relatively small range of frequencies is required to provide the necessary information, therefore a different filtering network can be used. In the tests in connection with this part of the work oscillation frequencies will never exceed 15 c.p.s. and hence the use of a simple R-L-C low-pass filter with a flat response from d.c. to about 50 c.p.s. would be sufficient and would eliminate the expense of the more elaborate combination type filter described above. The filter used in the tests is shown in fig. 45 page 187, and was designed as described in the appendix under "A Low-Pass Filter Design". The phase shift vs. frequency characteristic of this filter is included with the design to allow computation of the synchro pair phase shift by the analytical or indirect method.

Total Phase Shift for the Basic Synchro System

To obtain the total phase shift from sinusoidal synchro generator rotor position to the output signal from output filter for the test set-up used here it is necessary to consider only the shift introduced by the synchro pair and the output filter since there is no discriminator phase shift. For an example a rotor oscillation frequency of 15 c.p.s. will be used. From previous meaurement it is known that the phase shift from generator input to control transformer output is 3° for the 400 cycle carrier signal. The amount of shift between the 15 c.p.s. input and output waves due to the synchro reactances will then be:

$$\theta_s = \frac{\omega_s}{\omega_c} \cdot \theta_c = \frac{15}{400} \times 3 \cdot 0^\circ = 0 \cdot 1125^\circ \quad \text{or}$$

$$\theta_s = 6 \cdot 75 \text{ minutes.}$$

The phase shift across the output at 15 c.p.s. was measured to be $8 \cdot 0^\circ$ lag. The output signal from the filter will therefore lag the rotor position input sinusoid by $8 \cdot 11^\circ$. Excluding the effect of the output filter would leave only about a tenth of a degree shift between the 15 c.p.s. component of the discriminator output and the 15 c.p.s. position input wave. If the output filter could be designed to introduce a negligible phase lag into the useful signal bandwith then the harmonic response function from position input to filter output would be essentially just a gain constant.

Total Harmonic Response Function

The total harmonic response function consists of the gain magnitude and the phase shift. From previous measurements it is known that the rms voltage gain of the basic synchro pair specified earlier is 0·0833 per degree of shaft rotation in the linear portion of the gain vs. relative shaft displacement characteristic as shown in fig. 31. Assuming a 115 volt input to the synchro generator yields a gain of 9·6 volts rms/degree from shaft angular rotation to control transformer output. The gain magnitude of the discriminator was measured at 0·732 volts d.c. output/volt rms input. Thus the overall gain from the difference in angular position between the synchro generator and control transformer rotors to the discriminator output is 7·0 volts d.c./degree and the total harmonic response function is given by:

$$\text{h.r.f.} = \frac{7 \cdot 0}{1 + j\frac{f}{7650}}$$

where the break frequency of 7650 c.p.s. is deduced from the 15 c.p.s. phase shift of $0{\cdot}1125^{\circ}$. To complete the harmonic response function of the synchro system this above expression must be multiplied by the h.r.f. of the output filter.

Direct Measurement of the Harmonic Response Function

To measure the harmonic response function of the synchro pair and discriminator by direct measurement is in theory a much simpler method than the indirect one just described however in actual practice the construction of the equipment necessary for such a measurement is considerably more costly and difficult. To obtain a shaft input function which will have a sinusoidal input with respect to time necessitates the construction of a mechanical oscillator. There are several types of devices which may be used as mechanical oscillators and the actual design used will depend on the accuracy desired in the waveform, the frequencies of oscillation which the unit must produce, and on the equipment available to build the oscillator. The device selected for the experimental work here is described in the Appendix under "Mechanical Oscillator". To obtain a comparison of the waves between which it is desired to measure the phase shift, both of these waves must be in electrical form ~~so as~~ to be able to use the Vectorlyzer. To obtain an electrical signal in phase with the sinusoidal shaft input a potentiometer was directly coupled to the shaft of the synchro generator on the mechanical oscillator. No normally detectable phase shift will occur between the shaft position sinusoid and the a.c. component of the potentiometer output. The control transformer of the synchro pair was mounted on the face-plate of the

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mechanical oscillator and was equipped with a heavily damped turning knob on its rotor shaft to allow the relative angle between the rotors of the generator and control transformer to be set at any desired value. In determining the harmonic response function this relative shaft angle is set so that oscillation will occur entirely within the linear region of the relative shaft-angle voltage-output characteristic of the synchro pair.

The equipment was set up with the output of the control transformer being fed into the discriminator and subsequently into the output filter. The output of the filter was placed across one of the inputs of the Vectorlyzer and the signal from the potentiometer across the other input.

Considerable noise was present in the signal from the potentiometer and this introduced a very large error factor into the phase measurement. This was eliminated by shunting the two output leads from the potentiometer with a 0·1 mfd. condenser. Some discrepency in the waveform produced by the mechanical oscillator was found and was due to inaccuracies in the manufacture of the bearing surfaces at the pivot points. The signal from the output filter contained some 400 cycle ripple which although it was small in magnitude compared to the oscillation frequency signal, was sufficient to introduce some error into the phase angle measurement.

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Actual phase angle measurements showed a phase difference between the potentiometer output and the signal from the output filter of approximately 8° for a 15 c.p.s. oscillation frequency. Exact measurements however were not obtained due to instability in the Vectorlyzer and also due to fluctuations in speed of the mechanical oscillator drive. This 8° is close to the phase shift value obtained by the previous method where it was found to be $8{\cdot}11^{\circ}$. The inaccuracies in the measurements made using the mechanical oscillator arrangement far exceeded one-tenth of a degree, the expected shift across the actual synchro pair, and hence exact verification of the previous method was not accomplished.

To accurately measure the phase shift of the harmonic response function down to a prevision of at least one-tenth of a degree requires a mechanical oscillator which will produce a sinusoidal waveform with a negligible degree of error. The design used here is quite adequate provided that the pivot points are constructed with bearings which allow only a negligible amount of undesirable play and provided the drive motor is sufficiently large so as to provide a fly-wheel effect and hence even out any fluctuations in speed due to the variation in loading caused by the linkages. If the motor is too small variations in speed and hence discrepencies in the waveform produced will result. The desired accuracy of measurement also cannot be obtained unless all ripple components of the signal from the discriminator are suppressed to a ~~very~~ negligible amount by the output filter. This could be adequately accomplished by using the R-L-C filter used here with a band rejection filter in a Rejection Amplifier to remove the 400 cycle ripple component. Another difficulty encountered was the inability to obtain results over a range of oscillation frequencies. This was due to the fact that the maximum obtainable oscillation speed of the mechanical oscillator (15 c.p.s.) was just equal to the minimum frequency for which the Vectorlyzer could be used to measure phase shift accurately. To obtain more general results the mechanical oscillator should be constructed to close enough tolerances to allow for oscillation up to 25 c.p.s. These improvements in the mechanical oscillator and filter system were not carried out here due to limitations on construction facilities, cost, and time, and because it was felt that the returns from such efforts would not fully justify the output necessary to produce such a device.

Artificial Verification of the Indirect Method

To verify the indirect method, that is the method of obtaining the phase shift contribution to the harmonic response function due to the reactances of the synchros by measuring the corresponding carrier frequency signal phase shift across the synchro pair and then multiplying by the

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ratio of ω_s / ω_c to obtain the desired θ_s value, a subterfuge can be used to produce larger phase shifts in the amplitude modulated signal components between the input to the synchro generator and the output from the control transformer and hence make the θ_s value, which is directly proportional to the phase shift in these components, greater. This method entails the introduction of a phase shifting network between the synchro generator and control transformer of the mechanical oscillator system. In this way a much larger phase shift can be introduced into the components of the amplitude modulated signal than would normally be introduced by the reactances of the synchros alone and subsequently produce a proportionately larger phase shift of the rotor oscillation frequency signal between the output of the filter and the generator rotor position input. The use of a filter design which would produce sufficiently small phase shifts at the oscillation frequency so that the contribution due to the reactances of the synchro pair becomes an appreciable portion of the total phase shift would also be desirable.

The verification of the indirect method is basically just the verification of the fact that when a low frequency signal is used to amplitude modulate a much higher carrier frequency signal and the resultant wave passed through a phase shifting network followed by a detector to extract the low frequency component, the phase shift existing between

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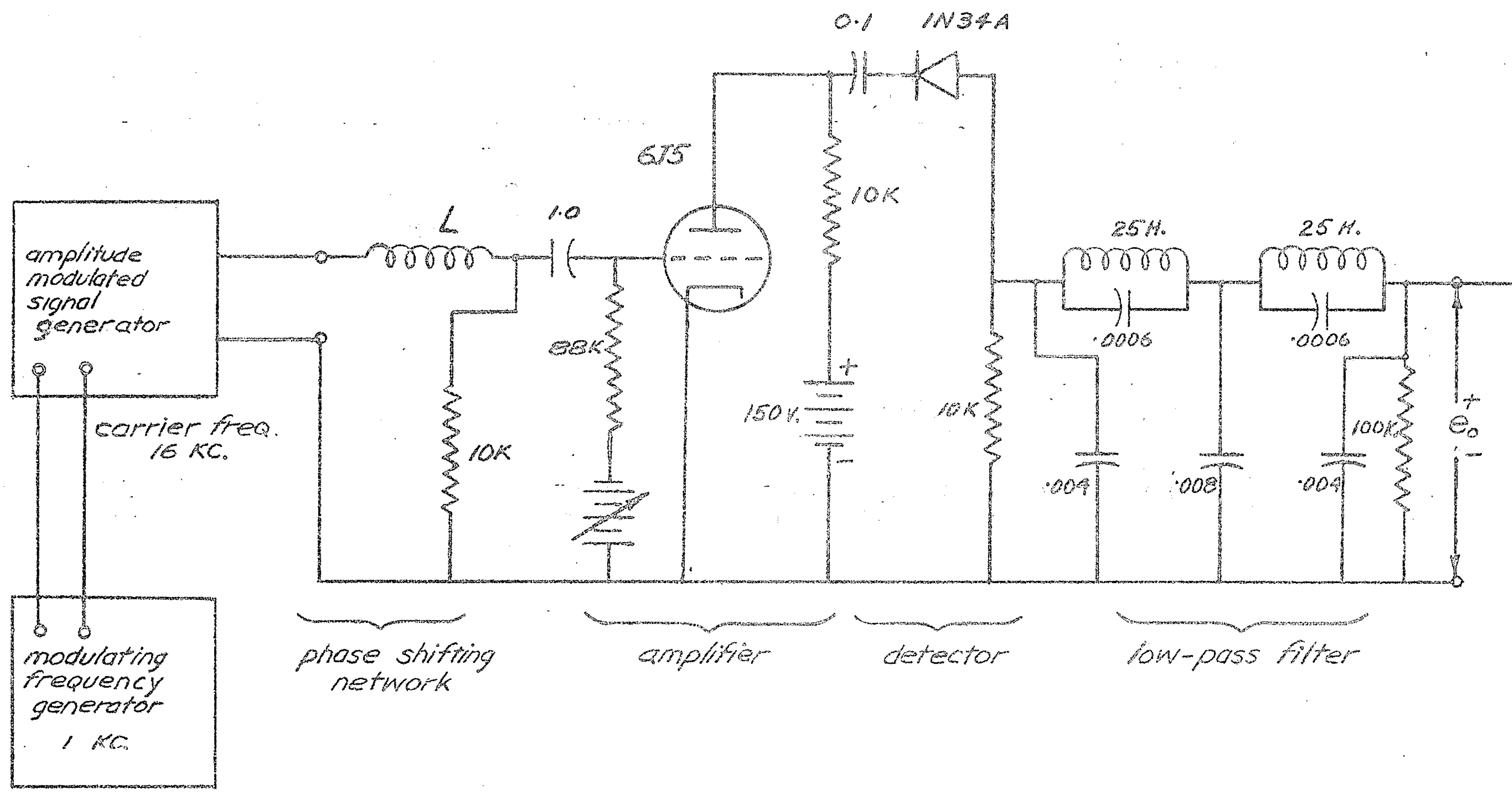
the final and original low frequency signals will be directly proportional to the shift of a carrier frequency signal across the same network with the constant of proportionality being the ratio of the modulating frequency to carrier frequency. The carrier frequency must be much greater than the modulating frequency to insure that the sideband components of the modulated wave undergo very closely the same phase shift as a carrier signal. Since this is the case a further simplification with respect to verifying the indirect method can be made since the equipment necessary to verify the above principle is much more easily set up than a mechanical oscillator system and also more precise results are obtainable.

Experimental Verification of the Modulation Theory Principle

To verify the modulation theory principle outlined above the network shown in fig. 41 was set up. A carrier frequency of 16 KC was used and a modulating frequency of 1000 cycles. The phase shift occurring between the final 1000 c.p.s. output signal and the output from the low frequency oscillator was measured without the R-L phase-shifting circuit in the network and then with this circuit inserted. This method eliminated the need for computing the phase shifts that occur in the amplifier output network and the output filter. Insertion of the phase-shifting circuit caused the



NETWORK for VERIFICATION of MODULATION PRINCIPLE

fig. 41

total phase shift between the low-frequency oscillator and the signal from the output filter to lag by an additional 5° over the shift occurring without this circuit. At a frequency of 16 KC the phase shift of a constant amplitude signal across the R-L phase-shifting network was 81°. The modulation theory principle would therefore predict a phase-shift change in the 1000 cycle modulating signal from the low frequency oscillator to the output of the filter of:

$$\frac{1000}{16000} \times 81^{\circ} = 5 \cdot 0^{\circ}$$

with the insertion of the phase-shifting network. Thus there is agreement between the theoretically derived value and the experimental results thereby giving some verification to the derived theory.

In performing these tests it was found that very close attention had to be paid to the signal levels used. With the phase-shifting network in the circuit a large attenuation of the signal from the signal generator was provided and hence a level of about 100 m.v. maximum peak to peak appeared at the amplifier input. (A percent modulation of 60 was used). Without the phase-shifting network the level reaching the amplifier input was considerably higher and caused distorted waveforms in the output of the amplifier and detector circuit. To overcome this it was necessary to reduce the input level until the amplifier input was approximately the same as it

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<u>NOTE</u>: Further deliberations and calculations since the preparation of the manuscript have shown an error to have been present in the experimental work just described. In the derivation of the approximate equation for the phase shift of the envelope outline of an amplitude modulated wave due to the effects of a reactive network, (p. 152), the development was based on the assumption that the break frequency of the network was much larger than the carrier frequency of the system. This led to the expression

$$\theta_s \doteq \theta_c \cdot \frac{\omega_s}{\omega_c}$$

which was employed in connection with the experimental work. The H.R.F. of the L-R phase-shifting network used in the test was determined experimentally and is given by:

$$\text{H.R.F.} = \frac{G}{1+j\dfrac{\omega}{1\cdot 77 \times 10^4}} \qquad \omega_0 = 1\cdot 77 \times 10^4 \text{ rad./sec.}$$

In the process of setting up the test the carrier frequency was unknowingly selected quite a bit higher in value than the ω_0 of the reactive network. (ie: $\omega_0 = 1\cdot 77 \times 10^4$ rad./sec., $\omega_c = 10 \times 10^4$ rad./sec.). This was necessary in order to achieve a high value for the phase shift in the carrier wave, however in obtaining this large θ_c value the basic assumption on which the approximate relation being tested was based, was seriously violated and hence the correlation of the theory and practice obtained in the tests was purely accidental and grossly in error. The actual value of the

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phase shift of the envelope waveform as calculated using exact methods* is 2·0°.

The discovery of this error brings out the point in that to experimentally verify the approximate relationship for θ_s would necessitate using a carrier frequency which was much less than the break frequency of the network being used to provide the phase shift. This would mean that the resulting carrier phase shift would be small and since the modulating frequency is usually much less than the carrier frequency, the relative magnitudes of the phase shifts which would be encountered in an experimental verification would probably be in the neighbourhood of, or even less than, the accuracy of the measuring devices and hence suitable verification of the approximate form would be very difficult.

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* Mason and Zimmerman - "Electronic Circuits, Signals, and Systems" (p.367)

had been in the previous test.

Additional Observations

An unexpected result was uncovered in the process of setting up the test to verify the modulation theory principle. It was first thought that a simple way to obtain a large phase shift of the modulated wave components was to pass this signal through a simple single-stage grounded-cathode amplifier thus imparting a 180° shift to the modulated wave components and thereby producing a resultant shift in the modulating signal of

$$\frac{1000}{16000} \times 180^\circ = 11 \cdot 25^\circ$$

between the low frequency oscillator and the output of the filter. The experimental results however showed either a 0° or 180° shift of the 1000 c.p.s. wave depending on the polarity of the detector diode. Examination of the theory behind the situation showed that although similar 180° phase shifts can be introduced into a constant amplitude, constant frequency signal by both a simple grounded cathode amplifier and also a linear passive reactive lumped-parameter network, the effects resulting from the operation of each of these networks on an amplitude modulated signal will not be the same.

Looking first at the effect of a simple grounded cathode amplifier stage on an amplitude modulated signal, consider that a wave such as shown in fig. 42a. is fed into the input of such a stage. Due to the basic action of the amplifier the output will be an exact reproduction of the input except for a polarity reversal and an appropriate gain factor. The output will therefore appear as in fig. 42b. From this it is seen that the envelope of the output wave will not have any time axis displacement with respect to the envelope outline of the input wave. This means of course that the amplifier has not affected the wave in any way that will result in a phase shift in the modulating frequency signal between the oscillator producing the modulating frequency and the output of the detector, due to the amplifier.

Consider the same wave being fed into a passive reactive filter which would produce a 180° shift of a carrier frequency signal. Since it is assumed that the sideband frequencies are quite close to the carrier frequency the phase shifts induced into the sideband components will also be approximately 180°. The simplest way of explaining the process here is to consider the input wave shown in fig. 42a. to be broken down into its three constant amplitude components, one at the carrier frequency and the other two at their respective sideband frequencies. These are shown in fig. 43 with the difference in frequencies between the components exaggerated for descriptive purposes only. Since the frequencies and therefore the phase shifts through the reactive network

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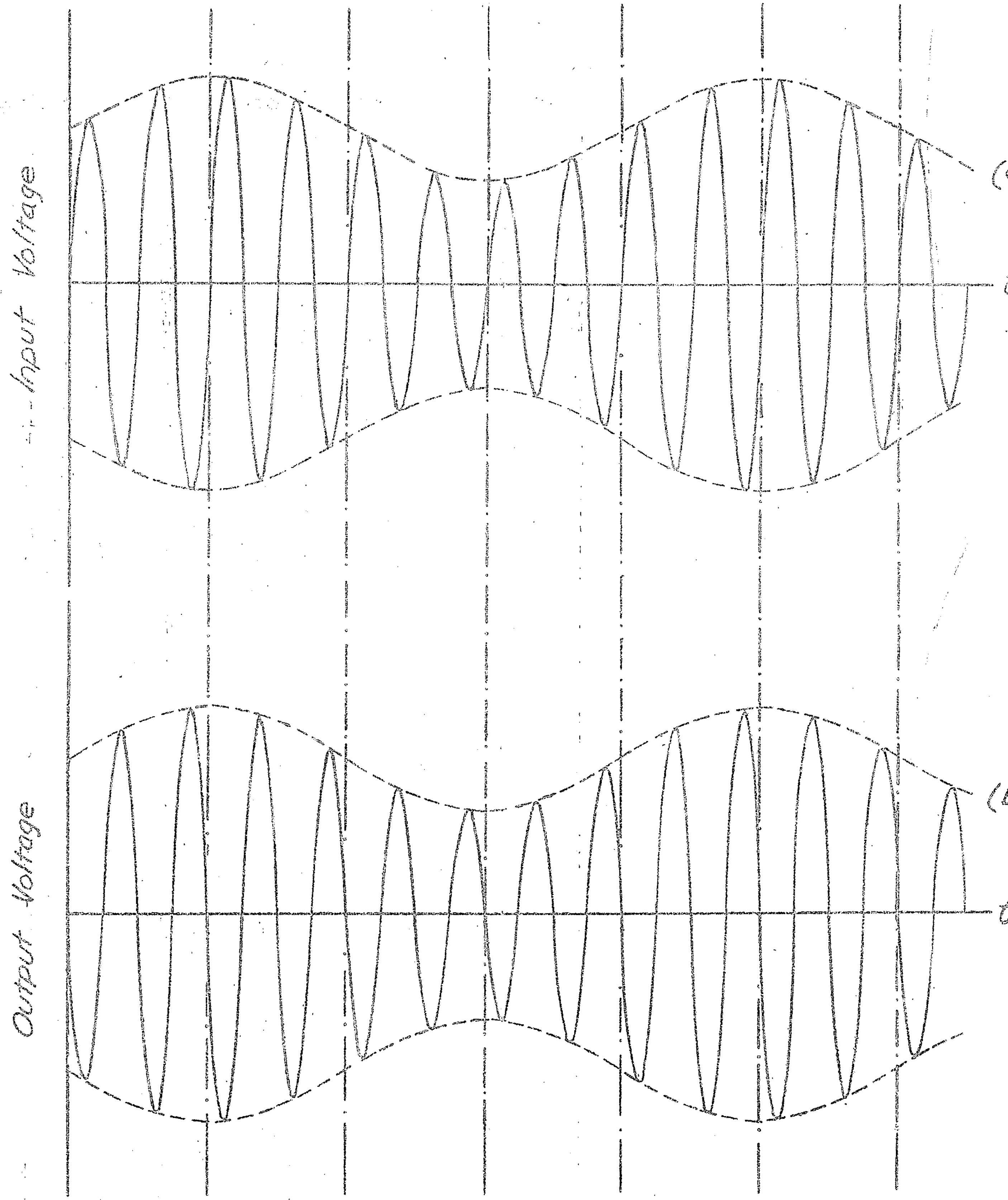
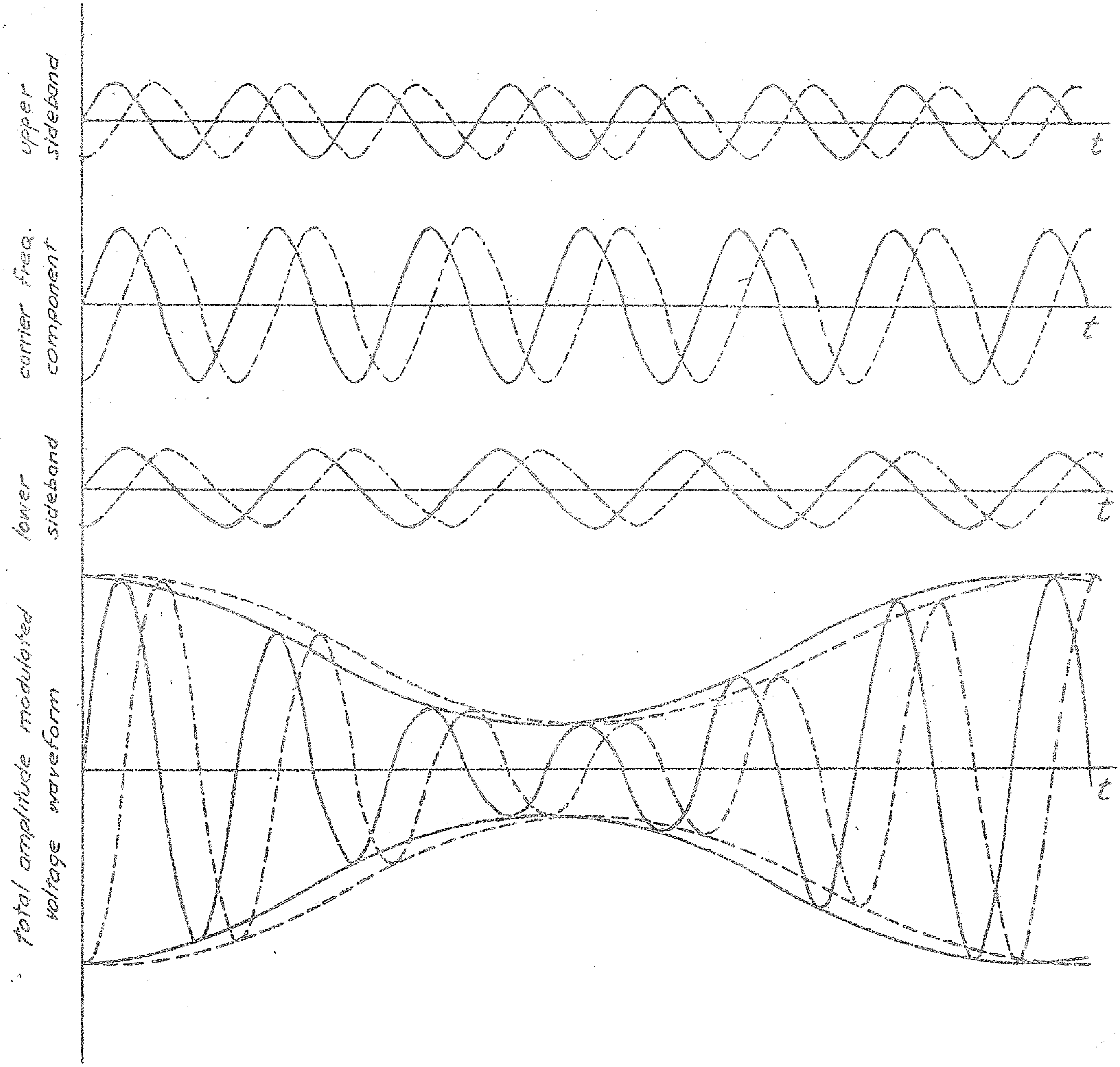


Effect of a single stage grounded-cathode amplifier on an amplitude modulated signal.

fig. 42



The solid-line waveforms are the total amplitude modulated wave and its corresponding Fourier components as they appear at the input to the passive reactive network. The dotted-line waveforms illustrate the total wave and its components as they appear at the output of the network. It is assumed that each component is shifted by approximately 90° by the reactive network.

ILLUSTRATION of the MODULATION THEORY PRINCIPLE

Fig. 43

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for each of the frequency components will be in actuality just about equal, the time axis displacement of the components will be almost equal. Recombination of the shifted components will show that the envelope outline of the output has a displacement relative to the envelope outline of the input wave and hence when the modulating frequency component is recovered after detection and filtering there will be a displacement between it and the oscillator output, where the modulating signal was originally produced, due to the effects of the filter. The amount of this angular shift will be the ratio of the modulating to carrier frequencies times the carrier signal phase shift across the filter. To illustrate this time axis displacement in the outline of the envelope wave consider fig. 43. This shows the input wave being broken down into its three frequency components. These components are also shown as they would appear after each undergoes an approximate 90° phase shift due to the effects of a passive filter. These are shown dotted. Recombining these shifted components shows the amplitude modulated wave to be approximately shifted along the time axis as illustrated by the dotted envelope outline. It is felt that showing only a 90° phase shift and allowing the reader to extrapolate the idea to a 180° shift provides for a clearer illustration of the phenomena. It is possible to obtain an ~~wrong~~ incorrect picture of the situation if one attempts to illustrate immediately a 180° shift without first regarding the effects caused by smaller shifts.

This then shows the variation in effects of the single stage grounded-cathode amplifier and the passive filter network on an amplitude modulated signal and also provides an explanation of the results which occurred in the first attempts to design a phase shift network to allow experimental verification of the modulation theory principle.

A Further Thought

After completing the theoretical development and associated experimental work in connection with the preceding section the thought occurs as to what the effect would be in regard to the phase shift caused by a linear passive filter between the output of the filter and the original signal from the modulating frequency oscillator output if the carrier frequency and the produced sideband frequencies were not close together. Considering the signal in terms of its Fourier components this would mean that the phase shifts imparted to each component by the filter would be distinctly different and hence the time axis shift of each would have corresponding variations. Recombination of the components as they appear at the output of the filter would probably not yield a displacement of the envelope outline relative to the outline of the input wave with as simple a relation to the carrier

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shift as was found with the case where all frequency components were close together. The investigation of such a situation would probably be quite informative but unfortunately would tend to stray from the main theme of this dissertation and hence will be left here simply as an interesting thought.

A P P E N D I X

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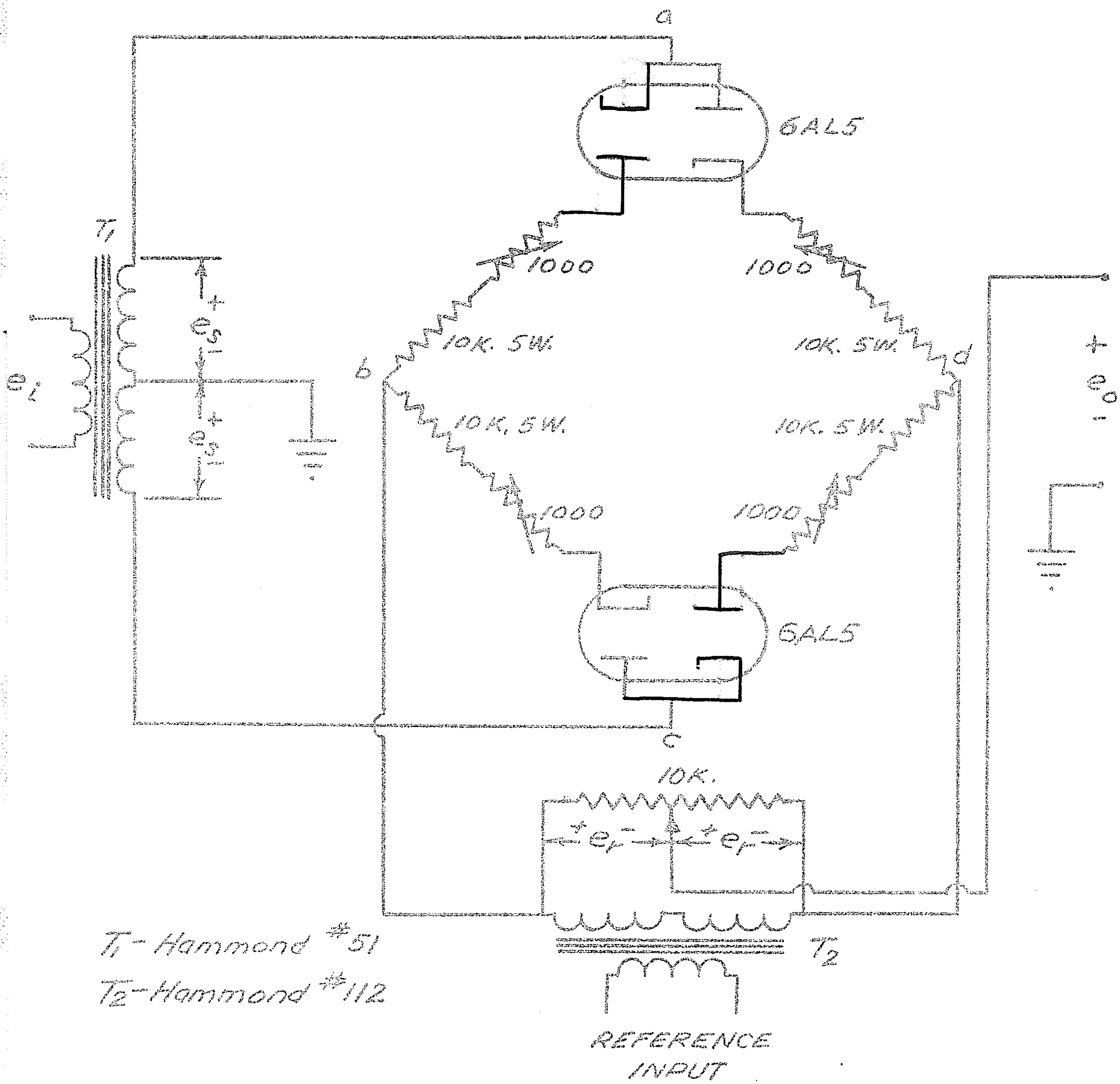
Ring Discriminator Design and Construction

The final discriminator design is shown in fig. 26. Experimentally it was found that the maximum rms signal voltage, which is the input signal to the discriminator, was approximately 60 volts consequently $e_s \doteq 60$ v. This was for a 400 cycle supply voltage of 120 v. The reference voltage was therefore 120 v. and $e_r = 180$ v. Thus e_r was greater than the necessary value of twice e_s. The value of fixed resistance used in the arms of the bridge was 10K ohms. This value was higher than originally intended but was found necessary in order to limit the current in the primary and secondary of the reference signal transformer. Using such high resistances had the favourable effect of swamping out unbalances in the diode forward resistances. 1000 ohm potentiometers were also used in each arm of the bridge in order to allow for final balancing of the arm resistances.

Considerable difficulty was encountered in first aligning the unit. This was traced to a slight inaccuracy in the center-tap location on the secondary of the reference transformer. To rectify this the resistance network shown in the schematic was placed across the transformer secondary and the output lead taken from the center-tap of the potentiometer. This provided means for accurate balancing of the output signal. Any inaccuracy in the placement of the center-tap on the primary of the input signal transformer



RING DISCRIMINATOR
(figure 26 repeated)

secondary was not great enough to prevent balancing of the output and consequently no correction instigated.

The discriminator is aligned by adjusting the potentiometers on the front of the unit. Four of these are the 1000 ohm arm-balancing resistors and the fifth is the potentiometer used to provide the accurate center-tap across the secondary of the reference input transformer. These are all adjusted until the balanced full-wave rectified output waveform is obtained. The alignment is carried out with the filter <u>not</u> connected to the discriminator and with the oscilloscope leads placed across the discriminator output.

A plot of the d.c. component of output voltage from the discriminator with respect to rms input voltage is shown in the following figure. The curve is linear over most of its extent as would be expected since the d.c. component of a full wave rectified wave does vary linearly with the amplitude of the wave. Some non-linearity is however found in the curve once the rms input voltage exceeds 35 volts. This might be due to tube-effects although no specific answer for this can be given at this time since this effect was not appreciably detrimental to the operation of the system to warrant investigation to determine the cause.

The d.c. component was measured by placing a d.c. voltmeter across the output terminals of the discriminator.

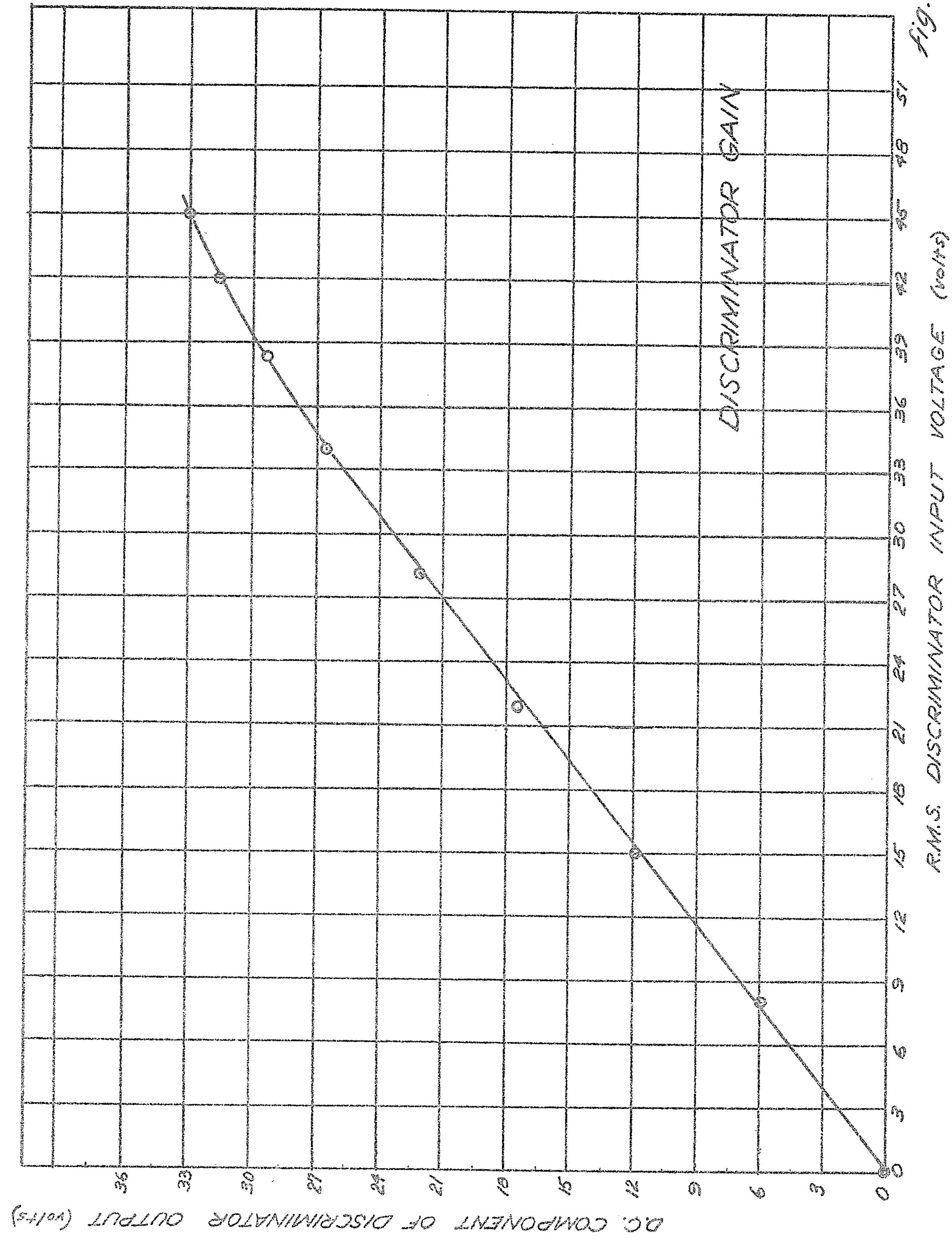

DISCRIMINATOR GAIN
R.M.S. DISCRIMINATOR INPUT VOLTAGE (volts)
0 3 6 9 12 15 18 21 24 27 30 33 36 39 42 45 48 51
D.C. COMPONENT OF DISCRIMINATOR OUTPUT (volts)
0 3 6 9 12 15 18 21 24 27 30 33 36

fig. 44
KAI

The output filter was not used in this measurement and consequently the gain factor of the discriminator and filter combined will be slightly lower than the values shown in this curve due to attenuation across the filter.

A Low-Pass Filter Design

If the maximum frequency of generator rotor rotation or oscillation is much less than the carrier frequency then adequate filtering can be obtained in certain instances from a low-pass filter design* which displays a reasonably sharp leading edge on its attenuation characteristic. In connection with the experimental work here on the harmonic response function of the basic synchro pair the maximum oscillation frequency was only 15 c.p.s. and the carrier frequency was 400 c.p.s. and hence this type of filter was suitable for this test.

$$L = \frac{Z_o}{\pi f_c} \qquad C = \frac{1}{\pi \cdot f_c \cdot Z_o} \qquad Z_o = \sqrt{\frac{L}{C}} = \pi \cdot f_c \cdot L$$

$$L_1 = mL \qquad C_1 = \frac{1 - m^2}{4m} \cdot C \qquad C_2 = mC$$

choosing $f_c = 300$ c.p.s. (point of initial attenuation)

$f_{\infty} = 600$ c.p.s. (point of maximum attenuation)

$L = 25$ henrys

* Westinghouse Industrial Electronics Reference Book Page196, filter no. 3

yields $$m = \sqrt{1 - \left(\frac{f_c}{f_{co}}\right)^2} = 0.866$$

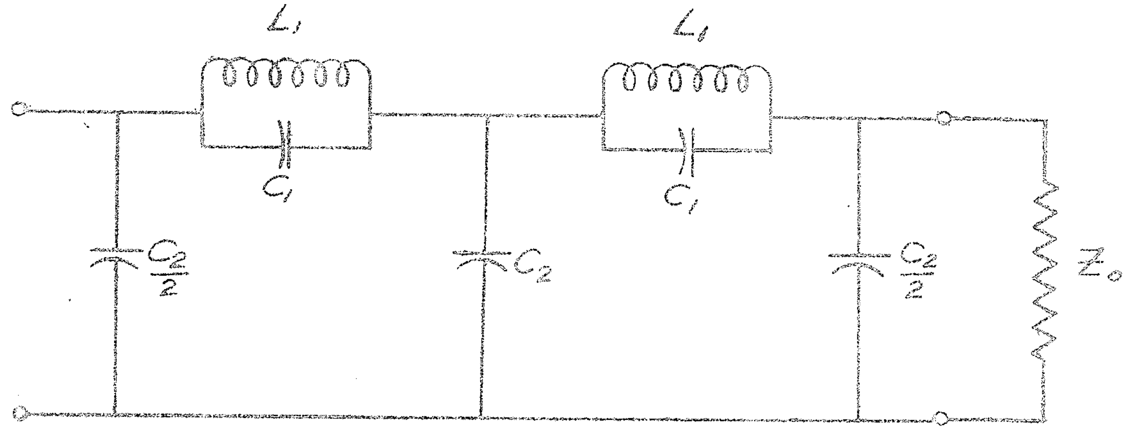
$$L_1 = \frac{L}{m} = 28.9 \text{ henrys}$$

$$Z_o = 27{,}200 \text{ ohms}$$

$$C = 0.04 \text{ mfd.}$$

$$C_1 = 0.0029 \text{ mfd.} \doteqdot .003 \text{ mfd.}$$

$$C_2 = 0.0345 \text{ mfd.} \doteqdot 0.035 \text{ mfd.}$$



LOW PASS FILTER

fig. 45

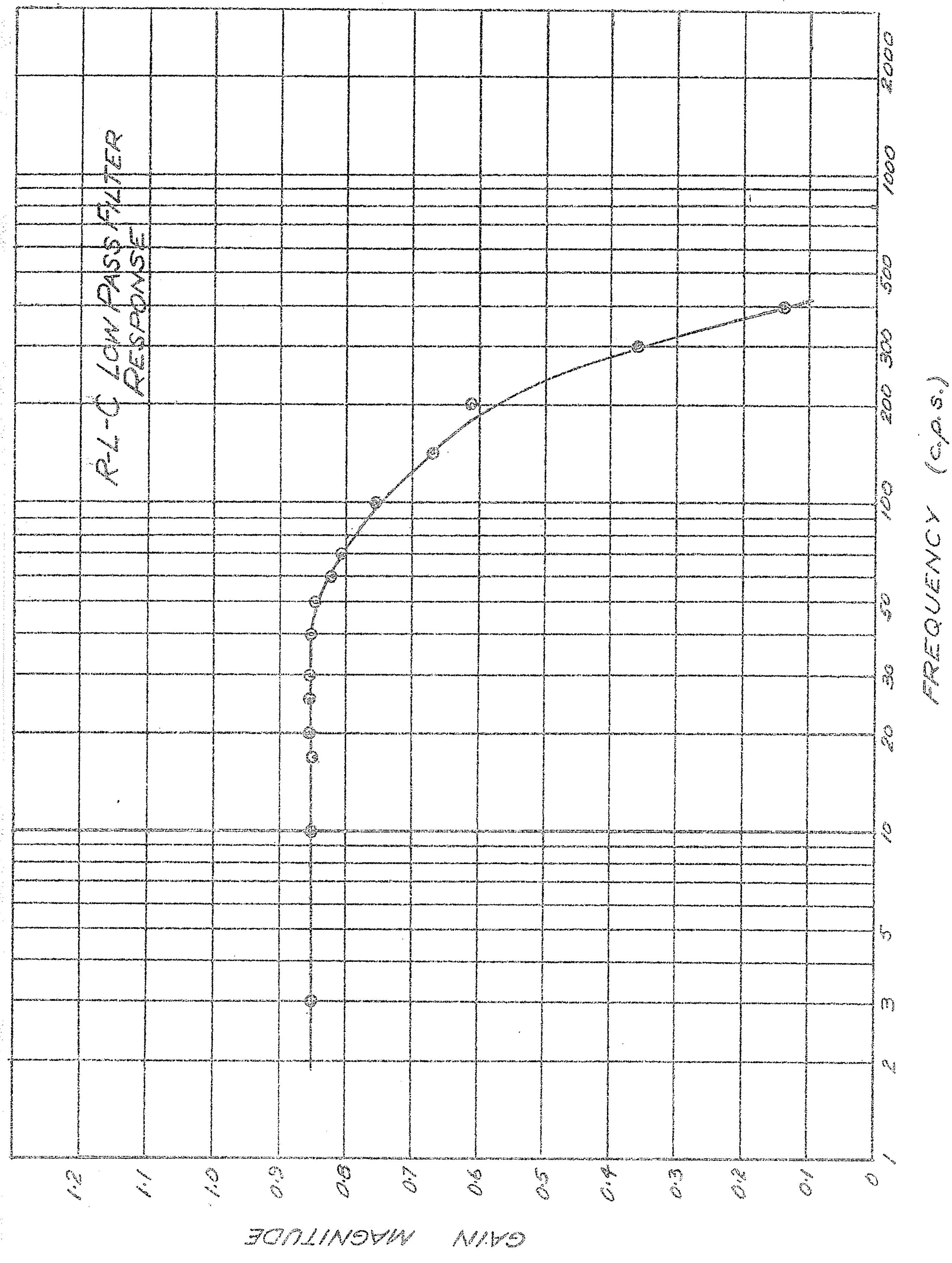

R-L-C LOW PASS FILTER RESPONSE
FREQUENCY (c.p.s.)
GAIN MAGNITUDE
1 2 3 5 10 20 30 50 100 200 300 500 1000 2000
0 0.1 0.2 0.3 0.4 0.5 0.6 0.7 0.8 0.9 1.0 1.1 1.2

Fig. 46

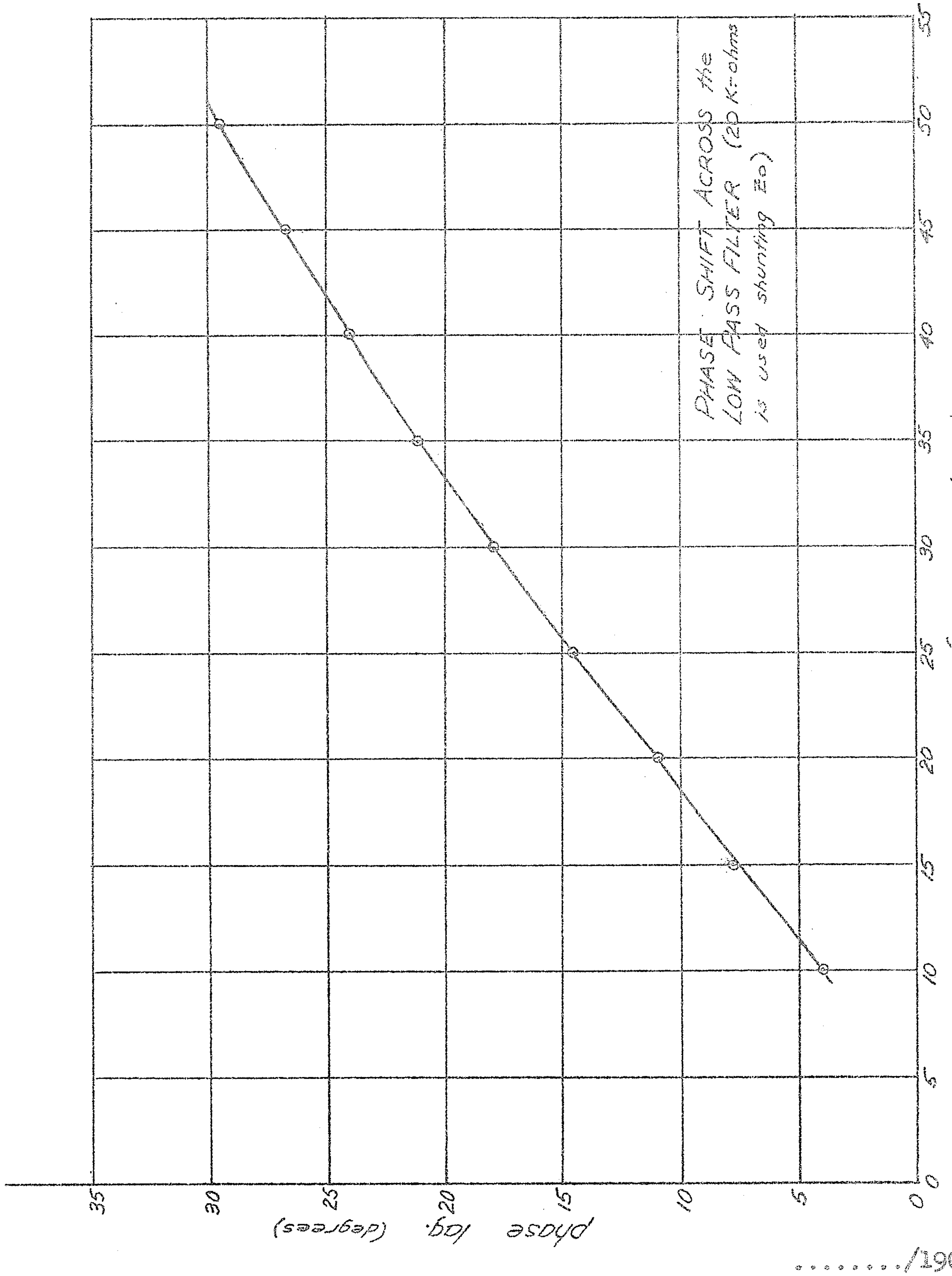

PHASE SHIFT ACROSS the
LOW PASS FILTER (20 K-ohms
is used shunting Zo)
Frequency (cps.)
phase lag. (degrees)
0 5 10 15 20 25 30 35 40 45 50 55
0 5 10 15 20 25 30 35

fig. 47

Synchro Test Bench

A synchro test bench was constructed and was equipped with a synchro generator, a differential generator, and a control transformer. Provision was made in the interconnecting wiring to insert the differential generator into the system simply by moving an 8-pin plug from the "differential-generator-out" socket to the "differential-generator-in" socket. Sockets on the front of the unit were installed for the 400 cycle input to the generator and for the output signal from the control transformer.

Each of the units was accompanied by a small d.c. motor rated at 250 r.p.m. at 28 volts supply voltage. These were coupled to the synchros through a system of drive-wheels and were intended to serve as drive motors in experimental work involving continuous rotation of the synchros. The motors, although slightly underpowered, did work well enough to provide good results in certain continuous rotation experiments. The drive-motors were installed with a device which allowed them to be easily detached from the synchro shaft when it was desired to run static positioning tests with the synchros.

Dial plates were constructed and mounted on each shaft to provide a means for measuring the amount of angular rotation of the synchro rotors. An attempt was made to

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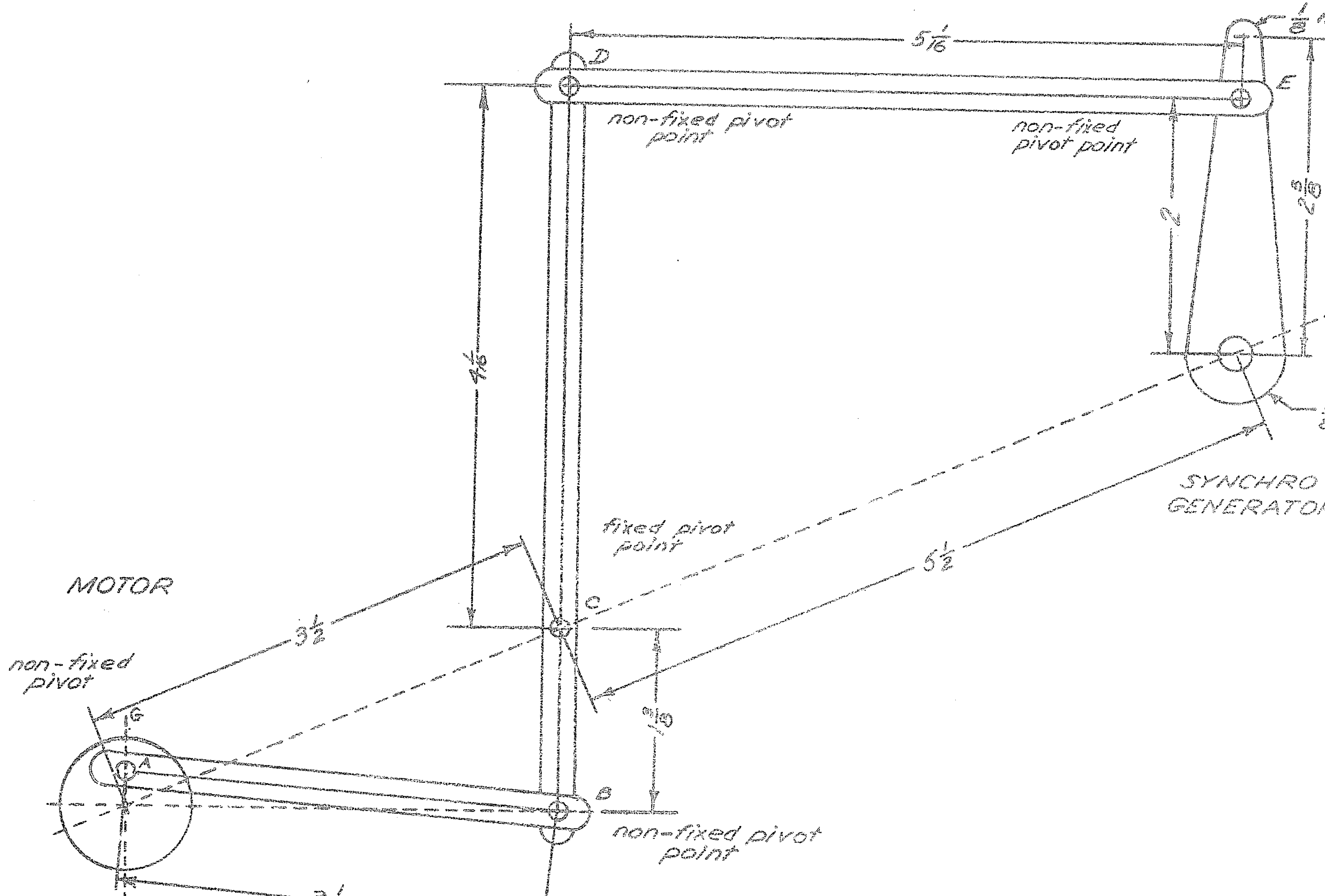
construct verniers to accompany these dials so that sufficiently accurate measurements could be made which would yield experimental results regarding the errors inherent in the synchros. This was found to be an almost impossible job using only ordinary drafting equipment as the errors in the construction of the vernier were greater than the errors which the device was intended to measure and hence no error measurements were made. The dial plates however were accurate enough for the remaining tests.

Equipment Cabinet

The discriminator, output filter, and power supply control panel were all built into an equipment cabinet. The control panel is fairly straight forward in arrangement but some description of it will be included to facilitate possible future use. The two switches marked "alternator" are used in starting and stopping the 400 cycle supply and operate as described under POWER SUPPLY. The 400 cycle voltage controls consist of a coarse control (toggle switch) and a fine control (potentiometer). The remainder of the 400 cycle equipment on the panel is self-evident. A 6·3 volt pilot lamp is connected to the filament supply and a 28 volt pilot lamp is connected across the 28 volt d.c. feed from the alternator.

Mechanical Oscillator

The mechanical oscillator was constructed to be used in the direct determination of the harmonic response function of the basic synchro pair. It was designed to impart a sinusoidal oscillation to the rotor of the synchro generator with an amplitude of oscillation of 20°. This figure of 20° was chosen since it kept the operation of the synchro pair within a region which provided reasonably linear synchro operation. The oscillator drive motor was supplied from a variac to allow speed control of the motor and hence control of the frequency of oscillation of the rotor. A diagram of the linkages is shown in fig. 48. This mechanism was arrived at after several attempts at other designs. The first design attempted was to connect the synchro rocker-arm to the rotating motor driver wheel with a straight connecting rod of length L. If R is the radius of rotation of the motor pilot point and if L is much greater than R then the desired sinusoidal oscillation of the synchro will be obtained. This type of mechanism however introduces a type of error into the waveform which is rather undesirable. This error causes the sine wave to become distorted and spreads one-half of the wave over a base length which is greater than 180° and the other half a corresponding amount less than 180°. This error approaches zero only as L becomes much greater than R. A sample design was tried with an L/R ratio of 13.4 and even



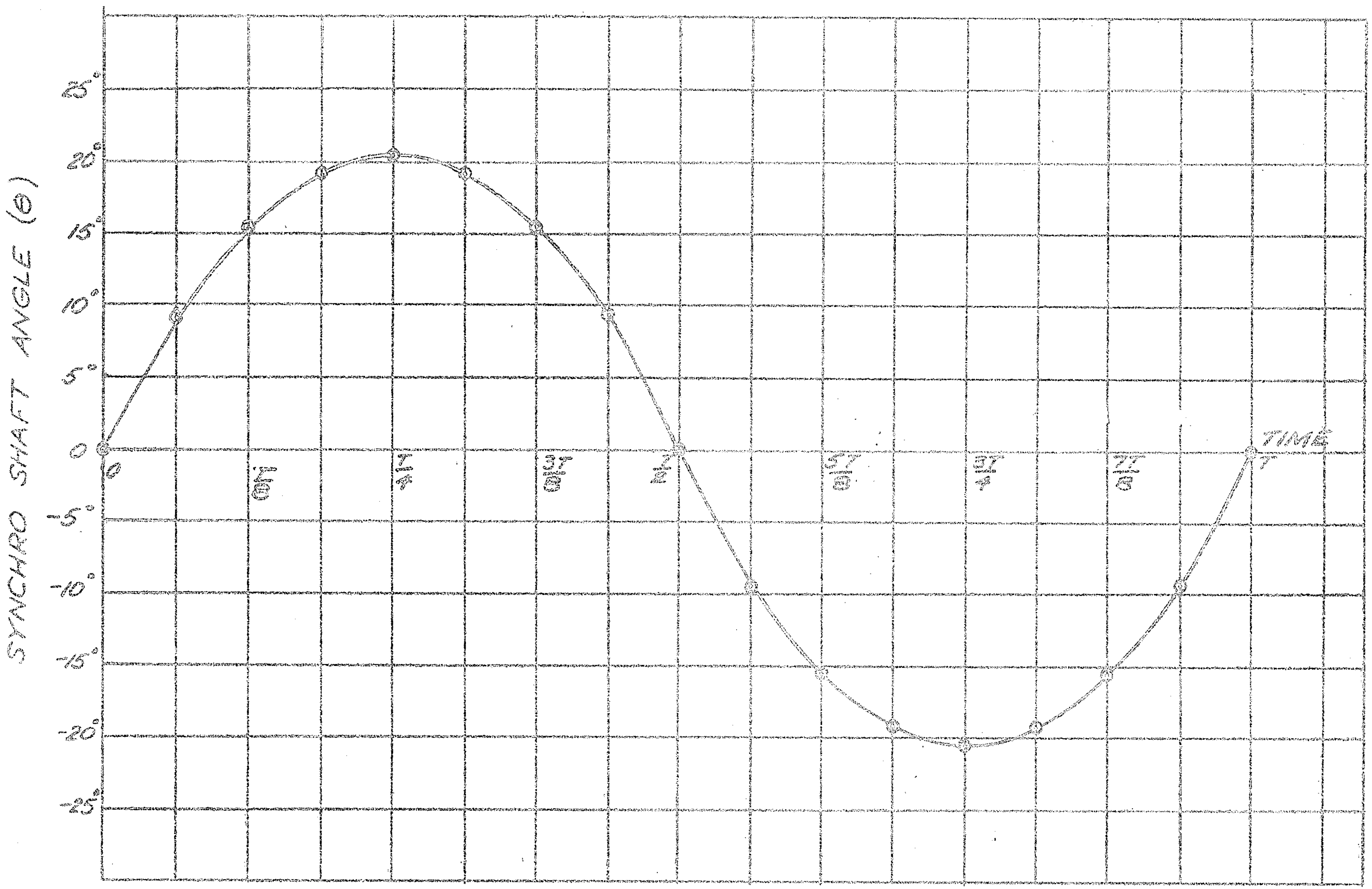
radius of rotation of pivot = $\frac{1}{4}$"
wheel diameter = 1"

LINKAGE LAYOUT for MECHANICAL OSCILLATOR

fig. 48

with this ratio one half of the sine wave had a base length of 225° and the other 135° which consequently gave considerable waveform distortion. To use this type of system and obtain satisfactory waveforms would mean using connecting rods which would be much longer than would be feasible for the frequencies of oscillation which it was desired to obtain.

The design finally decided upon was one step from the simple single connecting rod arrangement. The locations of the pivot points are instrumental in eliminating certain errors in the production of sinusoidal oscillation of the synchro rotor. The center of the motor shaft, fixed pivot point C, and the center of the synchro rotor are all located on the same center-line. With the linkages in the positions shown pivot point B is so located that a line drawn through the center of pivot B, and normal to the center-line between B, C, and D, will pass through the center point of the motor shaft. Line FG is parallel to line BCD. This location of B eliminates the variation in base length for the two halves of the sine wave since 180° rotation of the motor shaft from the position shown will place all linkages except for BA in exactly the same position as shown in the diagram. The center of the pivot at D is located so that the line DE will be normal to the line from E to the center of the synchro shaft for the position of the linkages shown.



MECHANICAL OSCILLATOR SHAFT
POSITION WAVEFORM

fig. #9

The operation of the device is based on the principle that the vertical component of the position of point A follows a simple harmonic motion pattern with respect to time for constant speed rotation of the motor shaft. This sinusoidal motion is very closely imparted to point B, the error being the difference between arc length and the desired straight line motion of point B. This error is relatively small as maximum angular amplitude of linkage BC motion is 10°. The motion of point B is reproduced by point D and thus the movement of D with respect to time is nearly sinusoidal. This causes the synchro shaft to have a sinusoidal oscillation with only a slight additional error being introduced by the fact that the synchro connecting arm pivot distance, that is the distance from the center of the synchro rotor to point E, is smaller than pivot distance DC. The total error between the actual synchro angular motion and the desired sinusoid will always be less than about 5% which is well within tolerable limits.

A potentiometer was directly coupled to the synchro shaft to enable an electric signal to be produced which would be in phase with the shaft oscillation and thus facilitate measurement of the synchro pair phase shift.

400 Cycle Motor-Alternator Power Supply

The synchro units employed in the experimental work in this project were taken from a dismantled aircraft bombsight. To supply the necessary 400 cycle power a General Electric aircraft motor-alternator (model 5AS131JJT1) was employed. The specifications for this unit are as follows:

General Electric Motor-Alternator

Model 5AS131JJT1 - Type AS1

Input - 27 v. d.c. - 100 amps (maximum)

- rpm 8000

Output -

-115 volts - 400 cycles - 1500 VA

-1 phase - p.f. 0·9

Serial No. XA 13016

The unit contains a fairly complicated switching, system and voltage regulating system and so a brief description of the operation of this system will be included to facilitate future use of this power supply. All references to the circuit are made with respect to the diagram of the circuit in fig. 50 page 205.

Power Input

The ~~27 volts~~ d.c. input power to the unit is ~~made~~ fed through the two heavy contact posts on the end of the unit, the heavily insulated and guarded post bwing the positive terminal.

Power Output

The a.c. output power is taken from terminals F and E of the amphenol plug. A d.c. output is also provided in this plug with the 27 volts d.c. being across terminals C and G, C being positive and G being solidly grounded. This d.c. output is rated at 15 amps continuous duty with an allowable 30 amps peak output. The maximum a.c. current is approximately 15 amps.

Switching

The switching circuit utilizes two switches, one a continuous contact switch and the other a "hold" type switch. To start the motor-alternator the continuous contact switch, between A and B of input plug, is closed. Since B is connected through a 10 amp fuse directly to the d.c. input supply, closing this switch places 27 volts d.c. on terminal A. When the switch between H and A is held closed, 27 volts appears across the coils of the primary contactor and also the four-level relay thus causing these to close. Both are normally-open type relays. This then allows power to be fed through the contacts of the primary contactor to the d.c. motor shunt field and the d.c. motor armature. The armature is grounded at this time through a series resistance. The d.c. motor will

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now start to rotate with the initial surge of armature current being limited by the armature series resistance. The secondary contactor, which in the meantime is still open, has one side of its coil connected to 27 volts d.c. and the other side to one of the top contactors on the four level relay. Since this four level relay is now closed, the lead from the secondary contactor coil is directly connected to the ungrounded side of the series armature resistance. Since the current through this resistor will be relatively high thus causing a high voltage drop across the resistor until the armature has built up speed, the voltage across the secondary contactor coil will be very small initially. To explain this further, consider $E_{s.c.}$ as the voltage across the secondary contactor coil, and I_a as the armature current.

$$E_{s.c.} = 27 - I_a \cdot (R \text{ series})$$

Since the drop across the armature circuit is initially small as compared to $I_a \cdot R$ series, $E_{s.c.}$ will be initially very small. As motor speed builds up, back emf builds up and I_a will decrease thus allowing $E_{s.c.}$ to increase. When $E_{s.c.}$ reaches the required level, the secondary contactor will close, thus shorting out the series resistor and therefore allowing the motor to attain full speed. The switching procedure is then:

to start - 1. close switch AB

2. press switch HA until secondary contactor closes, then release.

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to stop - 1. open switch AB.

In starting, the switch between H and B need only be held closed long enough to allow the primary contactor and the four level relay to close. This is because once the 4 level relay is closed the d.c. voltage is applied across its own coil and the coil of the primary contactor through the second level (from the top) contacts and thus the short across H-A which originally supplied this conducting path is no longer necessary. It should be noted here though that a simple double-throw double-pole switch could also be used in this system. The switch would have to be a continuous contact type that would continually short A to H and B in the ON position and break these contacts in the OFF position. Leaving the short between H and A would have no detrimental effects to the operation of the unit as it would simply by-pass the closed second level contacts of the 4 level relay. The first switching scheme was included here, even though it is more complicated than the second system, to explain the originally intended use of the unit with respect to switching.

Radio Noise Filter

A radio noise filter is provided through which the d.c. power for the motor armature, motor field, and the a.c. field circuit is filtered. This is used to help eliminate any frequency components which might find their way into the synchros and be a source of error as discussed under the section on discriminators.

D. C. Motor Circuit

The feed for the d.c. motor comes through the main circuit breaker, through the primary contactor, to the d.c. by-pass radio noise filter. The shunt field circuit is fed from the output of the filter. The field coil has a non-linear Glo~~w~~bar resistor and a 2 ohm variable resistor in series with it, the variable resistor being used to control the current level. The armature is also connected to the output of the filter. Two brushes are used in parallel on the input and two on the output. The output of the armature circuit is grounded either directly or through the series resistor depending on the position of the secondary contactor as discussed previously.

A.C. Generator Circuit

The rotor circuit provides the d.c. field for the generator. The d.c. current is taken from the output of the noise filter, through a series variable resistance in the voltage regulator section, and fed to the rotor by slip rings or "collectors" as G.E. prefers to call them. The output of the rotor is grounded through the contact of the secondary contactor and therefore, until the secondary

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contactor is closed no d.c. field will be present in the generator and hence no a.c. power can be drawn from the unit. The stator provides the a.c. output to the load. This output appears across pins F and E of the output plug. A capacitor is used between each a.c. output terminal and ground to by-pass any high-frequency components that may have found their way into the 400 cycle output. A similar capacitor is also used on the input terminal to the rotor. Physically these condensers are the grey can type units located immediately below the Amphenol output plug.

A voltage stabilizing circuit is included in the a.c. circuit, Hence a portion of the a.c. stator output voltage is tapped off and applied across opposite corners of a full-wave bridge rectifier. The output of the rectifier has across it one side of the stabilizing transformer in series with the control coil of the voltage regulator. The other side of the stabilizing transformer has one end grounded through the contacts on the secondary contactor and the other end attached to the positive d.c. input to the rotor field. It therefore directly shunts the rotor circuit. The d.c. resistance of the rotor circuit and the transformer winding is seven ohms. The operation of the voltage regulating unit, which includes the bridge rectifier, seems fairly clear. If the a.c. output voltage increases above a reference level the rectified d.c. current out of the bridge rectifier, which feeds the control coil of the voltage regulator increases. This increases the resistance of the variable resistor in the regulator thus decreasing the rotor voltage and consequently reducing the a.c. output voltage.

Voltage Control

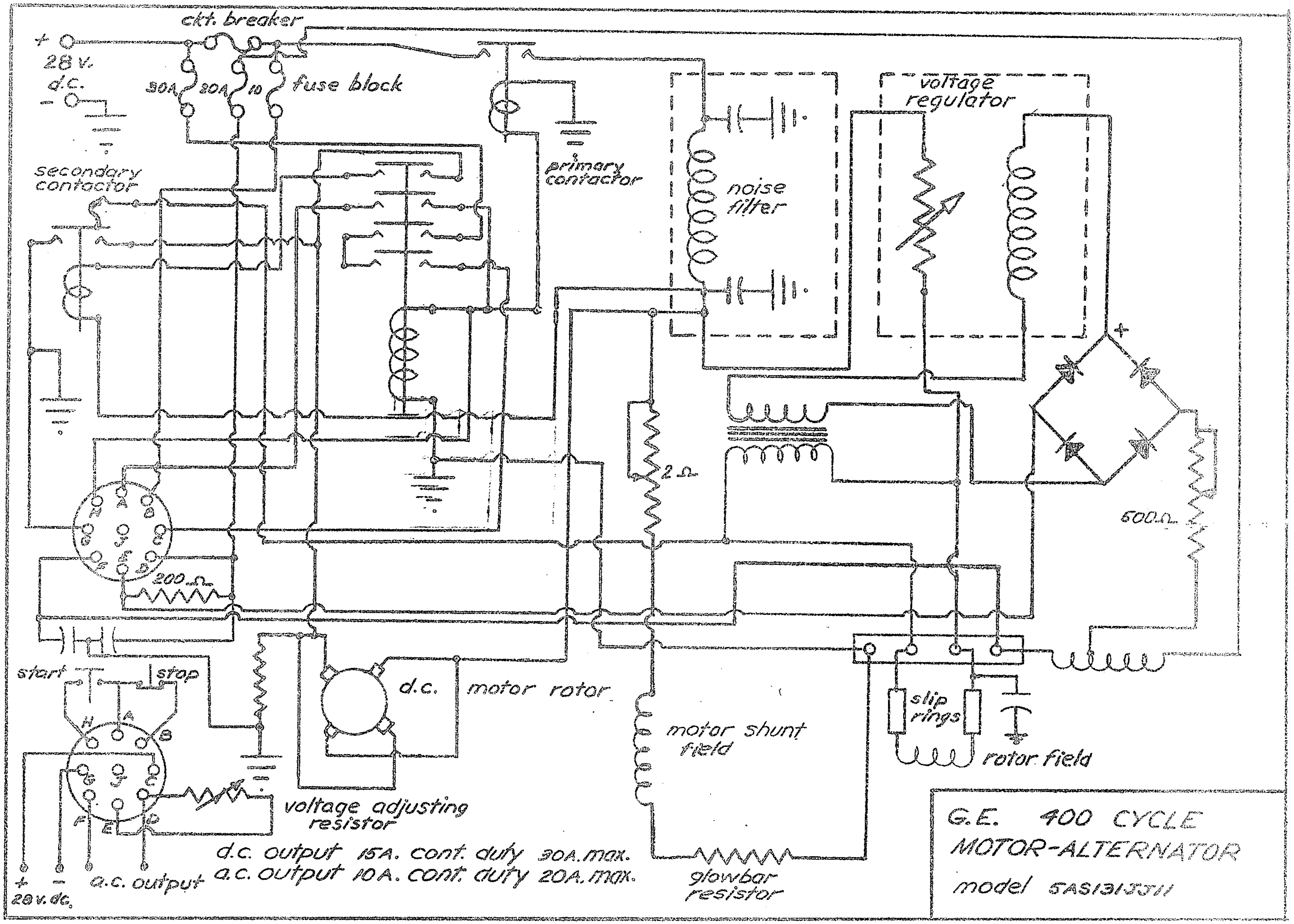
Control of the a.c. output voltage is obtained by use of a variable resistor across terminals D and E of the output plug. The exact value of this resistor is not given and would have to be determined experimentally if required. Its probable value is in the neighborhood of 200 ohms. If further voltage control is necessary, such as is required if a 120 v. output is desired at no load, this can be achieved by inserting resistance in series with the resistance of the voltage regulator (grey or blue leads).

Motor Speed Control

Motor speed can be controlled by controlling the current through the shunt field. This can be accomplished by raising or lowering the supply d.c. voltage. A 2 ohm variable resistor in series with the shunt field also provides some speed control.

Operation Notes

At small load values the voltage output from the alternator was much greater than the desired 120 volts until additional resistance was connected in series with the voltage regulator resistance. Also at small loads the 400 cycle output contained a very appreciable amount of approximately a 14th harmonic. This of course provided a very undesirable waveform. It was found that the effect of this harmonic decreased as load increased. To provide a suitable waveform for the experimental work the alternator output was filtered through an L-type filter consisting of a 4 mfd. shunt condenser followed by a 0·0175 henry series inductance. A 2·5 ampere dummy load was drawn by a resistance loading connected at the output of the filter. The power for the synchros was also drawn from the output of the filter. With this combination of filtering and loading the output voltage waveform to the synchros was relatively free of the undesired harmonic.



+ 28 v. d.c. −
ckt. breaker
30A
20A
10
fuse block
secondary contactor
primary contactor
noise filter
voltage regulator
+
2 Ω
600 Ω
200 Ω
A B C D E F G H J
start
stop
d.c. motor rotor
motor shunt field
slip rings
rotor field
voltage adjusting resistor
glowbar resistor
+ − 28 v. dc.
a.c. output
d.c. output 15A. cont. duty 30A. max.
a.c. output 10A. cont. duty 20A. max.
G.E. 400 CYCLE
MOTOR-ALTERNATOR
model 5AS131JJ11

fig. 50

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