ELECTRONIC AUTOPILOT SIMULATOR

by

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A.B., Adelphi College, Garden City
(1943)

B.E.E., Polytechnic Institute of Brooklyn
(1946)

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Signature of Author

Department of Electrical Engineering, September 1947

Certified by

Thesis Supervisor

Chairman, Departmental Committee on Graduate Students
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Abstract

The design of the Electronic Autopilot Simulator for simulating the performance of autopilots with linear control systems is based upon the breakdown of actual autopilots into their frequency invariant and variant parts. The error-producing section of the simulator provides electrically the sensitivity constants of the autopilot controller and motion-transducers. The simulator controller section simulates the frequency response of the automatic control system of an actual autopilot. It is designed to simulate simple (no-lag), simple-lag, and quadratic-lag controllers over the range of simple-lag time constants from 0.02 to 2.00 seconds, and over the ranges of quadratic-lag parameters for natural frequencies from 0.5 to 30.0 cycles per second and damping factors from 0.15 to 1.50.

The Electronic Autopilot Simulator is designed to operate on 400-cycle suppressed-carrier input signals which represent the deviation of an aircraft away from its reference axes of motion. Its output for each axis of motion is a 400-cycle suppressed-carrier signal corresponding to the control-surface deflection which would be produced by the aircraft autopilot which is being simulated.
ELECTRONIC AUTOPILOT SIMULATOR

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1. INTRODUCTION

1.1 Statement of the Problem

The purpose of the Electronic Autopilot Simulator is to simulate electronically various simple linear control systems in order to produce output voltages proportional to the control surface deflections which would exist in an aircraft operating under simulated flight conditions. The development of such an autopilot simulator can be used to simplify the present techniques for the analysis and synthesis of autopilots, and can be used as part of a flight simulator to check the stability of various types of aircraft in controlled flight.

A brief discussion of simple linear autopilots will be presented first in order to acquaint the reader with the general types which can be simulated by the Electronic Autopilot Simulator. Section 2 will present a more detailed analysis of the operating characteristics of autopilots and the design procedures which were used to synthesize the simulator. By comparing calculated autopilot performance with the actual Electronic Autopilot Simulator performance, the accuracy of the simulator will be determined, followed by suggestions for improving the equipment and the simulation of autopilots.
1.2 Discussion of Autopilots

Most autopilots are based on the assumption that for small deviation angles in roll, pitch, and yaw over a short period of time, the problem of simultaneous control about an aircraft's three rotational axes can be reduced to three independent control problems in roll, pitch, and yaw. Such autopilots can be broken down into two major sections, the error-measuring section and the controller section, for each of the three airframe motions they control.

The error-measuring section is composed of absolute and relative motion transducers which produce output signals proportional to functions of the deviations of the aircraft from reference directions. Depending upon the complexity of the autopilot, the output signals may be proportional to the deviation from reference direction, proportional to the sum of the deviation and rate-of-change of deviation, or proportional to the sum of the deviation, rate-of-change of deviation, and acceleration of the deviation away from reference direction. Derivative error signals are used to increase the stability and the speed of response of the aircraft in flight.

The absolute motion deviations from straight level flight are detected in autopilots by free gyros (positional gyro) whose gimbals are coupled to the airframe. The signal

created by the motion of the airframe away from the
reference position which is represented by the position
of the gyro spin-axis is usually mechanical or electrical
in nature, and is produced by pneumatic or electrical pick-
offs from the gyro assembly.

The directional gyro rotates about an horizontal
axis, and provides azimuth deviation signals. The signifi-
cance of the azimuth signals depends upon the absolute
direction in which the gyro spin-axis is oriented. In most
commercial autopilots, the directional gyro axis is aligned
with the heading of the aircraft when the autopilot "takes
over", and produces signals proportional to the deviation
from the original heading of the aircraft. The vertical
gyro rotates with its spin-axis vertical, and provides
device signals in absolute roll and pitch.

The signal proportional to the rate-of-change of
development from the reference direction under flight conditions
is usually produced in an autopilot by a rate gyro. In
some cases, the rate-of-change of deviation signal is obtained
by an electrical differentiating circuit operating upon the
development signal obtained from the free gyro.

The acceleration of motion of the aircraft away
from the absolute reference direction is sometimes measured
by an angular accelerometer mounted on a gimbal frame which
is coupled to the airframe, but is usually derived in an autopilot by differentiating twice the original deviation signal.

The position of an aircraft relative to a specified flight path is usually measured by r-f signal detectors or radar-tracking equipment, depending upon the type of flight-path information provided. First and second derivative error signals are usually obtained by derivative networks operating upon the original error signals.

The automatic controller sections of an autopilot provide displacements of the control surfaces proportional to the sum of the error and derivative signals derived from the error measuring section. In any actual controller, however, time lag is present because of the mechanical or electrical components which make-up the controller. Many controllers can be represented by quadratic-lag transfer functions. With an automatic controller which is represented by a quadratic-lag transfer function, it is important for the natural frequency of the controller to be far above the natural-frequency oscillations of the aircraft in free flight.

1.3 History of the Problem

To the knowledge of the writer, no autopilot simulator on a 1:1 time basis has ever been designed as such.
A servomechanism simulator, however, for simulating a fire-control system has been designed and built by the Servomechanisms Laboratory of M.I.T. This fire-control simulator is based upon the principle outlined by Ragazinni, Randall, and Russell in a recent article\(^1\), that of utilizing stable high-gain d-c amplifiers along with R-C feedback networks to produce desired transfer functions. The limitations of this simulator at present are caused mainly by the limitations of the d-c amplifiers and the necessity for continuous adjustment of balancing networks to compensate for drift.

The Servomechanisms Laboratory at M.I.T. has also designed a filtering circuit to be used in a 400-cycle suppressed-carrier system for producing particular transfer functions. Since it is assumed that the flight simulator from which the autopilot simulator will receive signals operates on 400-cycle suppressed-carrier, this circuit is of particular interest as its principle can be applied to the simulation of a variable autopilot controller. Further details on this system cannot be presented here because of security classification.

2. BREAKDOWN OF ELECTRONIC AUTOPILOT SIMULATOR

2.1 General

From the previous general discussion of autopilots, the breakdown of the autopilot simulator for producing a voltage proportional to each control-surface deflection angle falls naturally into three channels, one for each set of control surfaces (i.e., for the ailerons which primarily determine roll, for the elevators which determine pitch, and for the rudder which primarily determines yaw).

Each channel is comprised of two sections, an error-producing section and a controller section. Since each channel can be identical provided sufficient ranges of sensitivity factors, follow-up ratios, and types of controllers are provided to cover the variations in response of the three types of control surfaces, the design, construction, and testing of only one channel will be discussed in this paper. A block diagram of a single channel of the Electronic Autopilot Simulator is shown in Figure 1.

2.2 Error-Producing Section

The purpose of the error-producing section is to simulate the combined error-controlling signal which
AUTOPILOT SIMULATOR CHANNEL

BLOCK DIAGRAM OF AUTOPILOT-SIMULATOR CHANNEL FOR SINGLE-AXIS INFORMATION
would result from the absolute and relative motion transducers in an actual autopilot. It is assumed that electrical voltages proportional to error signals in position, rate and acceleration of the aircraft about its reference axes will be provided as absolute and relative motion inputs to the error-measuring section. It is further assumed that the equipment (transducers) which are used in an actual autopilot to detect the motion of the aircraft away from reference introduce negligible time lags into the automatic control systems compared to the lags introduced by the controllers. Since relative motion problems such as the control of an aircraft homing on a fixed point and the control of an aircraft flying on a fixed point and the control of an aircraft flying on a beam can be considered by the autopilot simulator, both absolute and relative motion error signals are combined in this section to form the final controlling error signal to be fed into the controller section.

2.3 Controller Section

The purpose of the controller section is to produce a voltage proportional to the control-surface position which would result from the combined input controlling signal. The various responses of the autopilot controllers (position servomechanisms) are simulated in this section.
The combination of both the controller sensitivity and the transducer sensitivities in the steady state condition is called the "follow-up ratio", and is defined as the control-surface deflection per unit deflection of the aircraft about its corresponding axis of motion. Thus, the controller sensitivity is set into the simulator system at the input to the controller section as part of the error-producing section, and the controller section is a function of frequency alone.
3. ERROR-PRODUCING SECTION

To effect electrical signals comparable to those which would be produced by error-detecting instruments in an actual autopilot, the absolute and relative motion input error signals to the error-producing section are operated upon by the sensitivity constants of the error-detecting instruments in the autopilot to be simulated. The sensitivity constant is defined as the ratio of the output signal from the instrument to the input signal detected by the instrument. Since these constants are primarily determined by the geometry of the aircraft, the dimensions of the control surfaces, and the type of controller used, no definitive range-of-values for each of the six component sensitivity factors for each of the autopilot-simulator channels can be stated.

The error-producing section, provides, however, a continuously variable range of sensitivity factors for each error component from 0.1 to 10.0. This is accomplished by cascading a potentiometer, attenuator and amplifier for each error component as shown in block-diagram form in Figure 2. Standard D.A.C.L. plug-in components and associated panels are used.

1. Dynamic Analysis and Control Laboratory, M.I.T.
*Plug-in attenuators are coded RA-1, or RA-18 depending upon the amount of attenuation introduced.
The potentiometer, coded R-2, is a 10,000 ohm, wire wound, constant 5000-ohm output-impedance potentiometer, introducing a gain factor which can be varied continuously from one-tenth to unity. Manufacturing specifications for the R-2 potentiometers are indicated in Figure 3.

Since plug-in units are used, either a D.A.C.L. Ra-1 or Ra-10 attenuator can be plugged into the panel, depending upon whether the sensitivity factor to be introduced is between 1.0 and 10.0 or between 0.1 and 1.0 respectively. Specifications for the Ra-1 and Ra-10 attenuators are presented in Figure 4.

A standard D.A.C.L. Ar-6 amplifier (Figure 5) for providing a gain of ten, completes each of the six circuits used to simulate the sensitivity factors of the autopilot error-detecting instruments.

The summation of the simulated error-detected signals is accomplished in a D.A.C.L. Ra-6 summing circuit (Figure 6) whose output feeds into another Ar-6 amplifier. The attenuation of ten introduced by the summing circuit is cancelled by the gain of ten of the amplifier. Thus, the error-producing section fulfills its function of providing a combined error signal for positioning the aircraft control surfaces.
SPECIFICATION FOR DUAL POTENTIOMETER

Manufacturer: Fairchild Camera and Instrument Corp.
88-06 Van Wyck Blvd.
Jamaica 1, N. Y.

Attention: Mr. H. H. Everett

Type: Fairchild type 736, dual

Electrical Angle of Rotation: 310° (terminals 1-3 and 5-7)

Mechanical Angle of Rotation: unlimited

Length of Threaded Mounting Studs: See B - 10125 - 3

Length of Shaft: 0.2500 +0.0005 in. -0.0000 in.

Wiring Diagram:

Phasing: with slider 4 at tap 2, adjust slider 8 to be at tap 6

Front Unit: (terminals 1-2-3-4):

Resistance Between Terminals:

<table>
<thead>
<tr>
<th>Terminal</th>
<th>Resistance</th>
</tr>
</thead>
<tbody>
<tr>
<td>1-3</td>
<td>10,000 ohms ±1%</td>
</tr>
<tr>
<td>1-2</td>
<td>5,000 ohms ±1%</td>
</tr>
<tr>
<td>2-3</td>
<td>5,000 ohms ±1%</td>
</tr>
</tbody>
</table>

Resistance Function: linear ±0.33% resolution ±0.15%

Rotation: motion of slider from 3 toward 1 is produced by clockwise rotation, facing shaft end.

Figure 3
GAKS FILAMENTS RUN ON D.C.
1644 FILAMENTS RUN ON A.C.,
ILY WITH SHIELDED LEADS
RESISTORS @ WATT ALLEN-BRADLEY UNLESS NOTED
T-1 DAQL SPEC SPA 105-1

NOTES:

SCREEN SUPPLY +150 V

MALLO4 (SINGLE SIDED)
POSITION 2: 10K, DOUBLE SIDED
POSITION 3: 5000 OHMS (RESISTIVE)

G50069-1

RELEASED
V2.0

DYNAMICS ANALYSIS AND CONTROL LABORATORY OF THE
MASSACHUSETTS INSTITUTE OF TECHNOLOGY
DIVISION OF INDUSTRIAL COOPERATION PROJECT NS 0367

SCHEMATIC-REPEATER AMP.- TYPE A-16-

50069-1

RELEASED

MASSACHUSETTS INSTITUTE OF TECHNOLOGY

AN 3102-16-1 P

50069-1
D.A.C.L. $R_A$-10 and $R_A$-1 ATTENUATORS

FIGURE 4

D.A.C.L. $R_S$-6 SIX-CHANNEL SUMMING CIRCUIT

FIGURE 6
The sensitivity of the controller is introduced at the output of the summing-circuit amplifier by means of an additional potentiometer-attenuator-amplifier combination identical to the six described above. A block diagram of the complete error-measuring section is shown in Figure 2.

The physical layout of this section is comprised of four of the standard D.A.C.L. Ar-6 amplifier panels. Three of the panels are identical and each provides the necessary internal wiring for two potentiometer-attenuator-amplifier circuits. The fourth panel of the error-measuring section is also a standard type panel and has provision for one set of potentiometer-attenuator-amplifier plug-in units on one half of the panel and a summing circuit-amplifier combination on the other half of the panel.
4. CONTROLLER SECTION

4.1 Requirements

In order to keep the autopilot controller simulator as simple as possible and yet cover as wide a range as possible of autopilot controller characteristics, it has been assumed that the transfer functions of the autopilot servomechanisms to be simulated vary in complexity from a simple proportionality between each control surface deflection, $\delta$, and its controlling signal, $\Delta$, (deviation of the aircraft from reference direction) to a function involving a proportional factor modified by a simple quadratic response.

The most complicated type of autopilot-controller frequency response that has been considered for simulation is as follows:

$$\frac{\delta(s)}{\Delta(s)} = \frac{K}{1 + \frac{\zeta}{\pi f_o} s + \left(\frac{1}{2\pi f_o}\right)^2 s^2}$$

(1)

where $K$ is the sensitivity of the controller

$\zeta$ is a non-dimensional damping factor

$f_o$ is the natural frequency of the system

$s$ is the Laplacian operator.

The range of values of the above parameters as provided by the simulator controller section is dictated
by the range of values found in practical servomechanisms:

(a) $\xi$ can be varied from .15 to 1.5, from an extremely underdamped to an extremely overdamped system.

(b) $f_o$ can be varied from .5 cps. to 30.0 cps.

Equation (1) can be rewritten in terms of physical parameters of a system:

$$\frac{\delta(s)}{\Delta(s)} = \frac{K_0}{1 + \frac{f}{K_o} s + \frac{J}{K_o} s^2}$$

where $f$ is a coefficient of friction

$J$ is moment of inertia

and $K_o$ is elastance.

If the inertia can be assumed zero, equation (2) becomes

$$\frac{\delta(s)}{\Delta(s)} = \frac{K_0}{1 + \frac{f}{K_o} s} = \frac{K_0}{1 + T_o s}$$

an expression for a system with a simple lag whose time constant $T_o$ can vary from 2 seconds lag which occurs in some of the autopilots used on the large commercial airlines to 20 milliseconds lag found in the fastest servomechanisms available today.
If the system is further simplified by assuming that ideally, the time lag can be considered negligible, equation (3) assumes the form of the simplest type of function that can be simulated,

\[ \frac{\delta(s)}{\Delta(s)} = K_0 \]  

(4)

Actually, this case can never occur in a physical system.

Table 1 shows the range of types of autopilot controllers that are simulated by the Electronic Autopilot Simulator controller section. Table 2 indicates the range of values of the parameters that are available in the controller-section transfer functions. As indicated above, these parameter values are based on practical operating limitations of present-day control systems.
### Range of Types of Controllers Simulated

**Table 1**

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Minimum</th>
<th>Maximum</th>
</tr>
</thead>
<tbody>
<tr>
<td>Damping Ratio ( \xi )</td>
<td>.15</td>
<td>1.5</td>
</tr>
<tr>
<td>Natural Frequency ( f_o )</td>
<td>.5 cps.</td>
<td>30 cps.</td>
</tr>
<tr>
<td>Time Delay ( T_o )</td>
<td>20 milliseconds</td>
<td>2000 milliseconds</td>
</tr>
</tbody>
</table>

### Range of Values of Controller Parameters

**Table 2**
4.2 Theory of Design

Simple-Lag Controller - The electrical analogue of any simple-lag system is a single stage resistance-capacitance circuit such as shown in Figure 7 below.

![Simple R-C Network](image)

Simple R-C Network

Figure 7

whose transfer function

\[
\frac{e_{\text{out}}}{e_{\text{in}}} = \frac{1}{RCs + 1} = \frac{1}{T_0s + 1}
\]

(5)

where the Time Constant \( T_0 = RC \).

Since the above circuit operates as a simple time lag only under d-c conditions, it is necessary to provide means for detecting the 400-cycle suppressed-carrier input controlling signal to the controller section before applying
it to the R-C network, hereafter called the simple-lag function network of the system. The output from the function network must then be applied to a modulator circuit in order to obtain a 400-cycle suppressed-carrier signal modulated by the delayed error signal at the output of the controller section.

The schematic diagram of the simple-lag function network for providing ten discrete values of time lag $T_0$ from .02 seconds to 2.0 seconds is shown in Figure 8. A balanced circuit is used to minimize drift from the detector and aid in balancing the modulating input signal to the balanced modulator. The nondimensionalized frequency response of this network in magnitude and phase is plotted in Figures 9(a) and (b).

The phase-sensitive detector which is used as the input stage is a simple half-wave detector. The schematic diagram of the detector and its linearity characteristic with the function network acting as a 400-cycle filter on the output are shown in Figures 10 and 11(a) and (b).

The 400-cycle modulator presented in Figure 12 is a standard-type push-pull modulator producing suppressed carrier modulated signals across the output terminals. The linearity characteristic of the modulator is shown in Figure 13. The non-linearities of both the modulator and detector limit the operating range of the controller.
**Schematic**

**Condenser - C = 0.5 μF**

<table>
<thead>
<tr>
<th>T-Dial Position</th>
<th>Time Constant $t_0$ (Sec)</th>
<th>Resistors (Mega Ohms)</th>
</tr>
</thead>
<tbody>
<tr>
<td>OFF</td>
<td>-</td>
<td>R12 0.02</td>
</tr>
<tr>
<td>1</td>
<td>0.02</td>
<td>R23 0.03</td>
</tr>
<tr>
<td>2</td>
<td>0.05</td>
<td>R34 0.05</td>
</tr>
<tr>
<td>3</td>
<td>0.10</td>
<td>R45 0.10</td>
</tr>
<tr>
<td>4</td>
<td>0.20</td>
<td>R56 0.10</td>
</tr>
<tr>
<td>5</td>
<td>0.30</td>
<td>R67 0.20</td>
</tr>
<tr>
<td>6</td>
<td>0.50</td>
<td>R78 0.25</td>
</tr>
<tr>
<td>7</td>
<td>0.75</td>
<td>R89 0.25</td>
</tr>
<tr>
<td>8</td>
<td>1.00</td>
<td>R90(10) 0.50</td>
</tr>
<tr>
<td>9</td>
<td>1.50</td>
<td>R10(11) 0.50</td>
</tr>
<tr>
<td>10</td>
<td>2.00</td>
<td></td>
</tr>
</tbody>
</table>

**Component Values**

**Function Network for Simple-Lag Controller**

**Figure 8**
Figure 9(a)

Calculated frequency response (magnitude) of autopilot simple-lag control

Normalized frequency $\omega = 2\pi f T_0$

Electronic autopilot simulator controller section

Surf. defl. angle
Controller input
Calculated frequency response (phase) of autopilot simple-lag controller.
Schematic Diagram of 400-Cycle Phase Sensitive Detector
ELECTRONIC AUTOPILOT SIMULATOR

CONTROLLER SECTION - PHASE SENSITIVE DETECTOR

GAIN - 1.33
REFERENCE VOLTAGE - 110 VOLTS

PHASE-SENSITIVE DETECTOR - LINEARITY CHARACTERISTIC

FIGURE II (a)
**ELECTRONIC AUTOPilot SIMULATOR**

**Controller Section - Phase Sensitive Detector**

**Figure II (b)**

---

**Input Voltage**

- Line of Perfect Linearity
- **In Phase Input**
- **Reverse Phase Input**

**400-Cycle Phase Sensitive Detector - Linearity Characteristic**

---

*Note: The diagram illustrates the linearity characteristics of a 400-cycle phase sensitive detector with input voltage values ranging from 0 to 5.0 on the x-axis and output voltage values ranging from 0 to 100 on the y-axis.*
SCHEMATIC DIAGRAM - 400 Hz MODULATOR

SLW 8-4-47
ELECTRONIC AUTOPilot SIMULATOR
.Controller Section - 400µ Modulator

Input Modulator Output
(d.c.) (µ) (400µ)

Gain - 0.625

400-Cycle Modulator - Linearity Characteristic

Figure 13 (a)
400-CYCLE MODULATOR- LINEARITY CHARACTERISTIC

Figure 13(b)

2/19/47
Quadratic Lag Controller - Assuming an ideal phase-sensitive detector and ideal modulator, the transfer function of the feedback circuit shown in Figure 14 below can be expressed as

\[
\Delta = \frac{KF(s)}{1 + KF(s)} = \frac{K}{\frac{1}{F(s)} + K}
\]

(6)

where \( K \) = open loop gain.

BLOCK DIAGRAM OF SIMPLIFIED QUADRATIC-LAG CONTROLLER SECTION

Figure 14

If the network used to fulfill the function \( F(s) \) is composed of two cascaded simple resistance-capacitance circuits (Figure 15), the overall closed loop
characteristic of the combination becomes the expression for a quadratic-lag controller.

The transfer characteristic of the function network $F(s)$, assuming zero input impedance, infinite load impedance and zero initial charge on the condensers, can be determined by solving the loop equations of the network in terms of the Laplacian operator $s$.

The loop equations of the network are:

$$e_1(s) = (R_1 + \frac{1}{C_1s})i_1(s) - (\frac{1}{C_1s})i_2(s) \quad (7)$$

$$0 = \left(-\frac{1}{C_1s}\right)i_1(s) + \left(R_2 + \frac{1}{C_1s} + \frac{1}{C_2s}\right)i_2(s) \quad (8)$$

Solving for the current $i_2$,

$$i_2(s) = \frac{\frac{e_1(s)}{C_1s}}{\left(R_1 + \frac{1}{C_1s}\right)\left(R_2 + \frac{1}{C_1s} + \frac{1}{C_2s}\right) - \left(\frac{1}{C_1s}\right)^2}$$

(9)

and since

$$e_2(s) = \frac{i_2(s)}{C_2s}$$

then

$$F(s) = \frac{\frac{e_2(s)}{e_1(s)}}{\frac{1}{R_1C_1R_2C_2s^2 + (R_1C_1 + R_2C_2 + R_1C_2)s + 1}} = \frac{1}{R_1C_1R_2C_2s^2 + (R_1C_1 + R_2C_2 + R_1C_2)s + 1}$$

(10)

By substituting the expression for $F(s)$ of equation (10) in equation (6), the closed-loop transfer function of the system becomes

$$\delta = \frac{K}{R_1C_1R_2C_2s^2 + (R_1C_1 + R_2C_2 + R_1C_2)s + (1 + K)}$$

(11)

A separation of the frequency variant and invariant terms yields

$$\delta = \frac{\left(\frac{K}{1+K}\right)}{R_1C_1R_2C_2s^2 + \frac{R_1C_1 + R_2C_2 + R_1C_2}{1+K}s + 1}$$

(12)
If the combined gain of the open loop system \( K \) is made very large with respect to unity, then equation (12) can be rewritten as

\[
\frac{\Delta}{\bar{\Delta}} = \frac{1}{R_1 C_1 R_2 C_2 s^2 + \frac{R_1 C_1 + R_2 C_2 + R_1 C_2}{K} + 1}
\]  

which can be compared to the desired transfer function of a quadratic-lag controller as stated in section 4.1,

\[
\frac{\Delta}{\bar{\Delta}} = \frac{1}{\left(\frac{1}{2\pi f_0}\right)^2 s^2 + \frac{\tau}{\pi f_0} s + 1}
\]  

The desired natural frequency obtained by equating the second-order terms in the denominators of equations (13) and (1) is

\[ f_0 = \frac{1}{2\pi} \sqrt{\frac{K}{R_1 C_1 R_2 C_2}} \]

The damping factor, determined by equating the first-order terms in equations (13) and (1) and substituting for "f_0" becomes
It is of interest to note that the natural frequency $f_o$ is dependent only upon the product of the resistances and capacitances of the function network, whereas the damping factor $\zeta$ of the quadratic-lag function is dependent only upon the ratios of the resistance and capacitance in one R-C combination to the resistance and capacitance of the second R-C combination. This fact simplified the switching problem of multiple combinations of natural frequencies and damping factors as described below.

Theoretically there are an infinite number of combinations of resistances and capacitances which can be used to obtain the required ranges of values (Table 2) for $f_o$ and $\zeta$. Actually, however, the practical limitations of leakage resistance in high-capacitance condensers and unstable high-value resistors limited the choice of components considerably.

Since it is most desirable from the point of view of the switching problem to have each parameter dependent upon only one type of element, it was decided
to let the product of the capacitances \((C_1 C_2)\) be maintained constant, and thus have \(f_0\) a function only of the product of the resistances \((R_1 R_2)\), and to let the ratio of the resistances \(\left(\frac{R_1}{R_2}\right)\) be maintained constant, and let \(t\) thus be a function only of the ratio of the capacitances \(\frac{C_1}{C_2}\).

If we let
\[
C_1 C_2 = K_1 \tag{16}
\]
and
\[
\frac{R_1}{R_2} = K_2 \tag{17}
\]
where \(K_1\) and \(K_2\) are constants, then these expressions substituted in equations (14) and (15) yield

\[
f_0 = \left(\frac{1}{2\pi} \frac{K}{K_1}\right) \sqrt{\frac{1}{R_1 R_2}} \tag{18}
\]

and
\[
t = \left(\frac{1}{2 \frac{K}{K_2}}\right) \left[\frac{K_2 \left(\frac{C_1}{C_2} + 1\right) + 1}{\frac{C_1}{C_2}}\right] \tag{19}
\]

The solution of equations (18) and (19) in terms of \(R_1 R_2\) and \(\frac{C_1}{C_2}\), as indicated below, provided a means for determining the range of elements required to simulate the ranges of \(f_0\)'s and \(t\)'s specified.
The expression for the product of the resistances

\[ R_1 R_2 = \left( \frac{1}{2\pi f_0} \right)^2 \frac{K}{K_1} \]  \hspace{1cm} (20)

is plotted in Figure 16 as a function of \( f_0 \) for three different values of \( K_1 \), assuming an open-loop system gain \( K \) of 100.

Figure 17 shows the plot of the ratio of resistances as a function of the ratio of the capacitances, the open-loop system gain \( K \), and the damping factor \( \zeta \).

These plots in addition to Figure 18 were used not only to determine the optimum values of \( K_1 \) and \( K_2 \) (\( C_1 C_2 \) and \( \frac{R_1}{R_2} \)) but also the optimum ranges of element values which could be used to simulate the required ranges of quadratic-lag parameters. Both \( K_1 \) and \( K_2 \) were chosen as unity in order to limit the range of \( R_1 R_2 \) to

\[ .0028 \leq R_1 R_2 \leq 10 \text{ Megohms} \]

as determined by Figure 16, and to limit the range of \( \frac{C_1}{C_2} \) as determined by Figure 18 to

\[ .0043 \leq \frac{C_1}{C_2} \leq 1.0. \]
ELECTRONIC AUTOPILOT SIMULATOR

CONTROLLER SECTION - FUNCTION NETWORK

ELECTRONIC AUTOPILOT SIMULATOR
QUADRATIC LAG CONTROLLER

\[ f_0 \text{ vs. } R_1, R_2 \]

\[ f_0 = \frac{1}{2\pi R_1 R_2 \sqrt{VR_1 R_2}} \]

\[ K = 100 \]

\[ C_1 C_2 = 1 \mu F^2 \]

\[ C_1 C_2 = 9 \mu F^2 \]

\[ C_1 C_2 = 25 \mu F^2 \]
ELECTRONIC AUTOPILOT SIMULATOR SECTION

QUADRATIC-LAG CONTROLLER SIMULATOR

RATIO OF RESISTANCES $R_1 / R_2$ IN FUNCTION NETWORK

RATIO OF CAPACITANCES $C_1 / C_2$ IN FUNCTION NETWORK

DAMPING RATIO TO BE SIMULATED ($\xi$), AND

OPEN-LOOP GAIN ($K$) OF CONTROLLER SIMULATOR

$$\frac{K}{K_2} = \frac{\sqrt{\frac{R_1}{R_2}}}{\sqrt{\frac{C_1}{C_2}}} = \frac{1 + \xi^2}{\xi^2 - 1}$$

FUNCTION NETWORK FOR QUADRATIC-LAG CONTROLLER

FIGURE 17
QUADRATIC LAG CONTROLLER
DAMPING RATIO $\xi$ VS. RATIO OF CAPACITANCES $C_1/C_2$

$$\frac{C_1}{C_2} = \left[ \frac{\xi}{\sqrt{1 - \frac{2}{K}} - 1} \right]^2$$

For the condition that $\frac{C_1}{C_2} = 1$

$K = 100$
These element ranges provide a range of natural frequencies \((f_0)\) from .5 to 30.0 cycles per second, and a range of damping factors \((\zeta)\) from .15 to 1.5.

Since \(K_1 = C_1C_2 = 1\), then \(C_1\) is the reciprocal of \(C_2\). Also, since \(K_2 = \frac{R_1}{R_2} = 1\), then \(R_1\) and \(R_2\) are equal and are referred to hereafter as \(R\).

Figure 19 shows the switching arrangement and schematic diagram of the function network for the quadratic-lag controller. Table 3 presents the parameter and element values associated with each switch position. As in the simple-lag function network, a circuit balanced to ground is used.

The phase-sensitive detector which converts the 400-cycle error-modulated suppressed-carrier input signal to the modulating signal, is the same simple half-wave detector (Figure 10) used in the simple-lag controller section. The R-C function network which follows the detector also acts as a 400-cycle filter on its output.

The same simple-lag controller standard-type balanced modulator is used for reconverting the low frequency error signal back to a modulated suppressed-carrier signal (Figure 12). Both d-c and a-c balance potentiometers are incorporated into the circuit. The non-linear effects of noise at low voltages and saturation at high voltages of the detector and modulator are the
SCHEMATIC DIAGRAM OF FUNCTION NETWORK FOR QUADRATIC-LAG CONTROLLER
limiting factors on the operating range of the entire system.

In order to obtain an open-loop gain of one-hundred, two D.A.C.L. Ar-8 amplifiers are used along with a potentiometer on the output of the modulator for adjusting the gain.

Negative feedback in the quadratic-lag controller section is obtained by using the final stage Ar-8 amplifier switched to its balanced output position. The positive output to ground is taken as the output from the controller section, while the negative output to ground is fed back to a summing circuit at the input. The other input to the summing circuit is the output signal \( \delta \) from the error-producing section. Actually the summing circuit which introduces an attenuation of ten, is also used, with the feedback input grounded, as the input attenuator for the simple-lag controller.

If Equation (10), the expression for the open-loop frequency-variant response, is rewritten with the Laplacian operator "s" replaced by its imaginary component "j2\( \pi f \)" with values of \( K_1 \) and \( K_2 \) substituted for \( \frac{R_1}{R_2} \) and \( C_1 C_2 \), respectively, and with the terms rearranged, then the equation becomes
ELECTRONIC AUTOPILOT SIMULATOR

Function Network for Quadratic-Lag Controller

Component Values

<table>
<thead>
<tr>
<th>Dial Position</th>
<th>C-1 (µf)</th>
<th>C-2 (µf)</th>
<th>C-3 (µf)</th>
<th>C-4 (µf)</th>
<th>C-5 (µf)</th>
<th>C-6 (µf)</th>
<th>C-7 (µf)</th>
<th>C-8 (µf)</th>
<th>C-9 (µf)</th>
<th>C-10 (µf)</th>
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<tbody>
<tr>
<td>OFF</td>
<td>.15</td>
<td>.30</td>
<td>.45</td>
<td>.60</td>
<td>.75</td>
<td>.90</td>
<td>1.05</td>
<td>1.20</td>
<td>1.35</td>
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</table>

**TABLE 3**

<table>
<thead>
<tr>
<th>f₀-Dial Position</th>
<th>f₀</th>
<th>R</th>
<th>Resistors (Megohms)</th>
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</thead>
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<tr>
<td>OFF</td>
<td></td>
<td></td>
<td>R-12 1.58</td>
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<tr>
<td>1</td>
<td>.5</td>
<td></td>
<td>R-23 0.786</td>
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<tr>
<td>2</td>
<td>1.0</td>
<td></td>
<td>R-34 0.265</td>
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<tr>
<td>3</td>
<td>2.0</td>
<td></td>
<td>R-45 0.205</td>
</tr>
<tr>
<td>4</td>
<td>3.0</td>
<td></td>
<td>R-56 0.096</td>
</tr>
<tr>
<td>5</td>
<td>5.0</td>
<td></td>
<td>R-67 0.070</td>
</tr>
<tr>
<td>6</td>
<td>7.0</td>
<td></td>
<td>R-78 0.051</td>
</tr>
<tr>
<td>7</td>
<td>10.0</td>
<td>.158</td>
<td>R-89 0.0164</td>
</tr>
<tr>
<td>8</td>
<td>15.0</td>
<td></td>
<td>R-9(10) 0.0278</td>
</tr>
<tr>
<td>9</td>
<td>20.0</td>
<td></td>
<td>R-10(11) 0.0528</td>
</tr>
<tr>
<td>10</td>
<td>30.0</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>
where \( f = \text{frequency} \). Normalizing the frequency of this expression with respect to \( R \) by letting

\[
u = \frac{2\pi f R}{\text{letting}}
\]
yields the new simplified equation below in terms of the variable \( u \),

\[
F(ju) = \frac{1}{(1 - u^2) + ju (C_1 + 2C_2)}
\]

Figures 20(a) and (b) show plots of the open-loop calculated frequency response of the quadratic-lag controller as expressed in Equation (22) in both magnitude and phase. The curves were calculated for the values of \( C \) corresponding to the ten damping factors \( \zeta \) which are simulated, and with the frequency scale calibrated in terms of the resistances associated with the ten natural frequencies simulated by the autopilot simulator.

The closed-loop quadratic-lag transfer function (Equation (13)) normalized in a similar fashion becomes

\[
\Delta (ju) = \frac{1}{1 - \frac{u^2}{K} + j \frac{u}{K} (C_1 + 2C_2)}
\]
Figure 20(a) shows the calculated frequency response (magnitude) of a quadratic-lag function network. The normalized frequency \( \omega = 2\pi f \) is plotted against the input and output values, with natural frequency \( f_0 \) (1/s) given alongside. The diagram illustrates the relationship between frequency and gain for different values of normalized frequency, demonstrating the network's frequency response characteristics.
ELECTRONIC AUTOPILOT SIMULATOR
CONTROLLER SECTION- FUNCTION NETWORK

NORMALIZED FREQUENCY \( \omega = 2\pi f \)

CALCULATED FREQUENCY RESPONSE (PHASE) OF QUADRATIC-LAG FUNCTION NETWORK
and is plotted in magnitude and phase in Figures 21(a) and (b). These curves indicate the desired response of the controller section for the simulation of autopilots with quadratic-lag controllers. In a later section these response curves will be compared with the actual response curves obtained from the autopilot-simulator.

**Simple Controller** - The simple (no-lag) controller is simulated by the Electronic Autopilot Simulator by placing a short circuit between the input and output terminals of the controller section. The entire controller section including the switching which is used for selecting the particular type of controller to be simulated is shown in block diagram form in Figure 22.

4.3 **Physical Layout of Controller Section**

Physically, the controller section comprises four standard-size panels and can be mounted on a 70"x19" rack along with the four error-producing section panels to form one complete channel of the Electronic Autopilot Simulator.

The first Ar-8 amplifier in the controller section (Amplifier 1) is mounted on Panel 1. Panel 2 comprises the summing circuit, detector, and modulator as shown in schematic form in Figure 23. A Mallory three-position
CALCULATED FREQUENCY RESPONSE (MAGNITUDE) OF AUTOPilot QUADRATIC-LAG CONTROLLER

FIGURE 21(a)

NORMALIZED FREQUENCY $u = 2\pi f R$

SURF. DEFL. ANGLE
CONTROLLER INPUT

$E = 0.15$

$E = 0.30$

$E = 0.45$

$E = 0.60$

$E = 0.75$

$E = 0.90$

$E = 1.05$

$E = 1.20$

$E = 1.35$

$E = 1.50$

ELECTRONIC AUTOPILOT SIMULATOR
CONTROLLER SECTION
FIGURE 2(A)

CALCULATED FREQUENCY RESPONSE (PHASE) OF AUTOPILOT QUADRATIC-LAG CONTROLLER
NOTE: ALL SWITCHES ARE GANGED TOGETHER

BLOCK DIAGRAM OF CONTROLLER SECTION
multiple-circuit switch for selecting the controller type, and the summing circuit, are actually mounted in the center of the chassis and shielded to isolate the detector and modulator circuits and thus prevent signals by-passing the function network which is placed on Panel 3.

Diagrams of the front-panel layouts of Panels 2 and 3 are shown in Figure 24. Panel 4 comprises the final amplifier stage (Amplifier 2) and associated transformer input.

4.4 Correlation of Experimental and Theoretical Response

General - The representative experimental response curves shown in Figures 25 to 28 prove qualitatively the theory of the design of the Electronic Autopilot Simulator controller section. The correlation of these curves with the exact desired response, however, on the basis of the calculated circuit parameters is limited, but not beyond understanding when all circuit factors are considered.

The assumption of zero input impedance to the function network in the present state of the circuit has been proved unjustified; a great deal of 60-cycle noise which is being picked-up by unshielded leads in the function network has made high-frequency low-voltage response inaccurate and difficult to obtain; the relative narrow
ELECTRONIC AUTOPilot SIMULATOR - CONTROLLER SECTION

FUNCTION NETWORK
SIMPLE LAG
+0
-0

QUADRATIC LAG

CONTROLLER TYPE

SIMPLE LAG
QUADRATIC LAG

Note: Panels 1 and 4 - Standard D.A.C.L. A1-B Amplifiers

Figure 24
operating ranges of the detector and modulator (Figures 12 and 13) have proved inadequate for closed-loop quadratic-lag operation since it is difficult to maintain the error signal above the noise level with the open-loop gain setting held constant at the value of 100, the basis on which all the controller parameters have been calculated. For the reasons stated above, the Electronic Autopilot Simulator in its present state is only a rough approximation of actual autopilot response. As the simulator of autopilots with three distinct types of controllers (simple, simple-lag, and quadratic-lag), however, the present simulator may be of value in simulating a particular type of response.

Quadratic-Lag Response - Analyzing more closely the quadratic-lag magnitude response curves shown in Figures 25(a) to 25(h), and reviewing the procedure used in obtaining them, it appears that the decrease in gain-setting of the modulator output potentiometer to obtain a relatively undistorted modulated output, has been the major factor in causing a consistent change from the desired and calculated peak of a magnitude of 3.2 at the normalized frequency of 10 to a peak of approximately the magnitude of 2.3 at a normalized frequency of 8.8.

Both the damping factor $\zeta$ and natural frequency $f_0$ are functions of open-loop gain $K$ by the relations
$f_0 = 0.5 \text{ cps}$

$\xi = 0.15$

**Comparison of Calculated and Measured Frequency Response (Magnitude) Characteristic for Quadratic-Lag Controller**
\( f_o = 1.0 \text{ cps} \)
\( \alpha = .15 \)

Comparison of Calculated and Measured Frequency Response (Magnitude) Characteristic for Quadratic-Lag Controller

**Figure 2.5 b**
ELECTRONIC AUTOPilot SIMULATOR
CONTROLLER SECTION

$\omega_0 = 2.0 \text{ cps}$
$\zeta = 0.15$

Comparison of Calculated and Measured Frequency Response (Magnitude) Characteristic for Quadratic-Lag Controller

Figure 25C
Comparison of calculated and measured frequency response (magnitude) characteristic for quadratic-lag controller.

$\omega_0 = 3.0$ cps
$\xi = 0.15$

Figure 25d
Comparison of calculated and measured frequency response (magnitude) characteristic for quadratic-lag controller

\[ f_0 = 5.0 \text{ cps} \]
\[ \beta = 0.15 \]
$f_0 = 7.0 \text{ cps}$

$
\xi = 0.15
$

Comparison of Calculated and Measured Frequency Response (Magnitude) Characteristic for Quadratic-Lag Controller

Figure 25\text{f}
$f_0 = 10.0$ cps
$\zeta = .15$

Comparison of calculated and measured frequency response (magnitude) characteristic for quadratic-lag controller.

Figure 25 g
COMPARISON OF CALCULATED AND MEASURED FREQUENCY RESPONSE (MAGNITUDE) CHARACTERISTIC FOR QUADRATIC-LAG CONTROLLER

$\omega_0 = 15.0$ cps
$\zeta = 0.15$

FIGURE 25h
Electronic Autopilot Simulator
Controller Section - Function Network

Measured Frequency Response of Quadratic-Lag Function Network
\[ f_0 = 5.0 \text{ cps} \]
\[ \xi = 0.15 \]

Comparison of calculated and measured frequency response (phase) characteristic for quadratic-lag controller.
presented below:

\[ \zeta = \frac{1}{K} \left[ \frac{C_1}{C_2} + \frac{1}{2} \frac{C_1}{\sqrt{C_2}} \right] \]

\[ f_o = \sqrt{K} \left[ \frac{1}{2\pi R^2} \right] \]

If \( K \) were decreased by a factor of two, \( \zeta \) would increase by a factor of \( \sqrt{2} \) and \( f_o \) would decrease by a factor of \( \sqrt{2} \) or the corresponding \( R \) would decrease by a factor of \( \frac{\sqrt{2}}{4} \).

Assuming that the decrease in gain was by this amount, the change in the calculated curve of \( \zeta = .15 \) as shown in Figures 25(a) to (h) would be a decrease in the peak from 3.2 to 2.3, and a shift in the occurrence of the peak from the normalized frequency \( u = 2\pi R \) of ten to \( \frac{10}{\sqrt{2}} \) or 8.4. Apparently this decrease and shift in the peaks of the experimental curves from that of the calculated curve has occurred. In other words rather than simulating a damping factor \( \zeta \) of .15 in the first switch position on the function network panel, a \( \zeta \) of approximately .23 is being simulated, and a corresponding decrease in each of the natural frequency dial positions tested is also occurring.
The tapering off of high-frequency response, approaching the limit of about .3 rather than zero, is due to the large quantity of 60-cycle pick-up in the system. With both the input and feedback short-circuited, the system can be balanced only to a minimum of one-tenth of a volt. The open-loop characteristics of a single switch setting, which may be compared to the corresponding ideal curves in Figures 20(a) and (b), are shown in Figure 26.

Within the accuracy of the method used for determining phase, by beating the input suppressed-carrier signal 90 degrees out of phase with the output on an oscilloscope and interpreting the pattern produced, the phase characteristic of the open-loop response appears to check closely the desired response. The large amount of 60-cycle pick-up present when the closed-loop characteristics were measured and the difficulty of accurately determining the phase made it very difficult to determine the closed-loop phase characteristics. Both the measured and theoretical frequency-phase response for the closed-loop quadratic-lag controller are shown in Figure 27. The 90-degree point on the plot, which could be determined more accurately than any other point, appears to check the theory derived from the magnitude response curves of the natural frequency having

been reduced by a factor of $\sqrt{2}$ by a decrease in open-loop gain.

**Simple-Lag Response** - Figures 28(a) to (c) show representative magnitude response measurements for the function-network switch set to "Simple Lag", along with calculated response curves for various time-constants.

It was found that the assumption of zero input-impedance to the function network for the controller switch on Panel 2 set to the simple-lag position (Figure 23) was completely unjustified, since in this position the function network was fed from the detector directly from the 50,000-ohm source-impedance summing circuit. The U.T.C.-A-19 two-to-one input transformer and the approximate two-to-one transfer impedance of the half-wave detector itself, produced an effective input-impedance to the function network of 400,000 ohms. This input impedance to the function network completely invalidated the specified time constants as set by the R-C constants of the function network.

In order to avoid this difficulty, the switching on Panel 2 (Figure 23) was set to "Quadratic Lag" for the measurement of simple-lag response. This change introduced the low-impedance output from the Ar-8 amplifier (set to a gain of ten) as the input to the phase-sensitive detector for the simulation of simple as well as
FIGURE 28 (a)

SWITCH POSITION \( T_0 = 0.02 \) SEC

- CALCULATED RESPONSE
- MEASURED RESPONSE

(\( T_0 = 0.02 \) PLUS \( T_0 = 0.05 \)

INPUT IMPEDANCE EFFECT)

FREQUENCY (CYCLES/SEC)

COMPARISON OF CALCULATED AND MEASURED FREQUENCY RESPONSE (MAGNITUDE)

FOR SIMPLE-LAG CONTROLLER
SWITCH POSITION $T_0 = 0.05$ SEC

- **CALCULATED RESPONSE**
- **MEASURED RESPONSE**

$T_0 = 0.05$

$T_0 = 0.08$

$T_0 = 0.05$ PLUS INPUT IMPEDANCE EFFECT

**Comparison of Calculated and Measured Frequency Response (Magnitude)**

FOR SIMPLE-LAG CONTROLLER
SWITCH POSITION $T_o = 0.10$ SEC

CALCULATED RESPONSE

MEASURED RESPONSE

$T_o = 0.13$

$T_o = 0.10$

(Frequency (cycles/sec))

Comparison of calculated and measured frequency response (magnitude) for simple-lag controller
quadratic-lag controllers. Thus the input impedance to the function network was reduced to approximately 30,000 ohms, taking into account the leakage resistance of the input transformer to the detector.

If this 30,000 ohm resistance is considered in series with the resistance of the function network for the formation of each R-C constant, then each calibrated time constant of the simple-lag function network (Figure 8) must be increased by .03 seconds to include the input-impedance effect. The response characteristics for three time-constant switch positions, as shown in Figures 28(a) to (c), appear to check this analysis.

Simple (no-lag) Response - Since a shielded cable is used to short-circuit the input and output terminals of the controller section for simple-controller response, the output from the controller section is a perfect reproduction of its input for the simulation of response of a simple no-lag controller.
5. CONCLUSIONS AND SUGGESTIONS FOR FUTURE WORK

Reviewing the correlation of experimental and theoretical response characteristics for the Electronic Autopilot Simulator controller section, it appears that the theory of the design of the autopilot simulator is basically sound, but that there remains a great deal of work to be done on its individual components to make the simulator function with the desired precision and accuracy.

The modification of circuit parameters to include the effects thus far considered negligible, the redesign of the detector and modulator to extend the simulator linear operating range, the shielding of all unshielded leads and the introduction of 60-cycle trap circuits to eliminate 60-cycle pickup, and the design of a 400-cycle band-pass filter to eliminate harmonics but not introduce phase shift up to 30 cycles, are a few suggested methods of improving the simulator performance. With these improvements, the Electronic Autopilot Simulator should be capable of simulating to a high degree of accuracy the performance of any simple linear autopilot.