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An investigation of
CERTAIN PROBLEMS IN THE USE OF
SAMPLE CORES IN ANALOG SIMULATION
OF MAGNETIC AMPLIFIER CIRCUITS

by

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A THESIS
SUBMITTED TO THE FACULTY
IN PARTIAL FULFILLMENT OF THE
REQUIREMENTS FOR THE DEGREE OF
MASTER OF SCIENCE IN ELECTRICAL ENGINEERING

Accepted (in revised form)
5-26-58
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Houston, Texas
May, 1958
FOREWORD

Although much of the work on this thesis was purely experimental, it was based on a general knowledge of the behavior and peculiarities of magnetic amplifier circuitry. Even with this knowledge, the extreme non-linearities involved caused considerable difficulty in the establishing of the best simulation process and demanded several revisions in the general methods being used. Even though many of the simulation methods and procedures were discarded as unsatisfactory for some reason, each provided a broader understanding of the behavior of magnetic amplifier circuits and components.

I am indebted to several people for valuable assistance in all phases of this study and would like to express my appreciation to them. Dr. P.E. Pfeiffer, the faculty advisor for the project, made several suggestions pertaining to computer technique and various theoretical aspects of the problem. W.R. Peters, technician for the Electrical Engineering Department, helped considerably with mechanical design on special equipment and with computer operational technique, in addition to keeping the necessary test equipment in working order.

I would also like to express my appreciation to the
McEvoy Company of Houston, Texas, for its generosity in financing a portion of the project, and to Magnetic Industries, Inc., for the use of the magnetic core materials and other components necessary in the various magnetic amplifier circuits investigated.
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SUMMARY

In order to establish the conversion factors between an actual circuit and its computer simulation, the simple half-wave magnetic amplifier circuit was used. The computer simulation of this circuit requires three operational amplifiers, with the associated scaling factors involving all of the variables of the actual circuit core material — area, mean length of path, and number of turns. In addition to establishing these scaling factors in the computer circuit, these variables affect the inductance-resistance ratio of the magnetic amplifier core and are used to determine the resistance to be added in the test core circuit so that the L/R of the test core matches that of the actual core being simulated.

Computer circuits were set up to simulate several actual circuits having cores with size and number of turns different from those of the test core of the computer circuit. Comparison of actual circuit data with simulated circuit data showed considerable difference, and this difference was found to be due to the difference in losses of the cores. It was necessary to scale the computer circuit data by a factor proportional to the ratio of the losses of the two cores. This factor was determined from a comparison of the B-H loops of the two cores, since the width of the B-H loop is proportional to the losses in the core. It should be noted,
in connection with this part of the study, that the geometry of the core is of some importance in the simulation process. Cores having different outside to inside diameter ratios have B-H curves which differ somewhat in slope.

Although the half-wave circuit serves to establish the necessary conversion factors in the simulation process, more complicated circuits should be studied to check the accuracy of the derived relationships. No data was taken on other circuits, due to time and computer limitations, but two more complicated circuits and their simulation are described, and recommendations for further studies are given in this report.
DEFINITION OF PROBLEM AND OBJECTIVES

I. STATEMENT OF PROBLEM

In the last few years, interest in magnetic amplifiers has increased considerably, due mainly to the great improvements in square-loop core materials and in metallic rectifiers. Also, magnetic amplifiers eliminate two serious disadvantages of vacuum tube circuitry — limited life and susceptibility to shock and vibration, and they are capable of being completely sealed, thus avoiding corrosion and problems with atmospheric and weather conditions. These advantages make the use of magnetic amplifiers quite attractive in a variety of applications.

The primary purpose of this thesis is the computer simulation of magnetic amplifier circuits, and, in particular, a study of the problems involved in the use of sample cores in the computer simulation. An exact solution of the circuit equations of a magnetic amplifier cannot be made analytically because of the non-linearities involved in these circuits. There have been several analytical studies made of magnetic amplifier circuits; however, the only attempt at a computer solution of magnetic amplifiers discovered was that made by a group of Westinghouse engineers. Because of the limited time available
for the study, no definite conclusions were reached concerning the ability to simulate magnetic amplifier circuits on the analog computer accurately.

In the use of sample cores for simulation of magnetic amplifier circuits, several problems are present in simulating circuits having cores with different size from that of the sample core. Of these problems, the one of most importance is correcting for difference in losses between the actual and sample cores.

II. OBJECTIVES

The main objectives of this study were as follows:

1. To devise a computer circuit for the simulation of a simple magnetic amplifier circuit.

2. To establish conversion factors to be used in conjunction with the computer circuit, so that the circuit parameters — number of turns, area of the core, and mean length of path of the core — can be varied easily.

3. To establish a method for taking into account differences in losses between actual and sample cores.

4. To simulate more complicated types of circuits in order to justify the methods evolved.
EXPERIMENTAL PROCEDURE AND RESULTS

I. THEORY OF COMPUTER CIRCUIT FOR SIMULATION OF HALF-WAVE MAG-AMP CIRCUIT

The computer circuit used throughout this study employs a sample core in the feedback circuit of an operational amplifier for simulation of the magnetic amplifier core material. The output voltage of this amplifier is proportional to \( N \frac{d\phi}{dt} \), where \( N \) is the number of turns on the sample core and \( \frac{d\phi}{dt} \) is the rate of change of flux created by a current applied to the input of the amplifier. This amplifier with the sample core as its feedback element provides the non-linearities of the square loop core material present in actual magnetic amplifier circuits.

The differential equations for the actual half-wave circuit shown in Fig. 1 are as follows:

For the gate circuit
\[
E_o = N_G \frac{d\phi}{dt} + i_L R_L + i_C R_D
\]
and for the control circuit
\[
E_c = N_C \frac{d\phi}{dt} + i_C R_C
\]

where:
- \( E_o \) = applied voltage
- \( N_G \) = number of turns of gate winding
- \( \frac{d\phi}{dt} \) = time rate of change of flux in core
\( i_L \) = current flowing in gate circuit
\( R_L \) = load resistance
\( R_D \) = diode resistance
\( E_C \) = control resistance
\( N_C \) = number of turns of control circuit
\( R_C \) = Resistance in control circuit
\( i_C \) = current flowing in control circuit

If \( R_C \) is of such a value that
\[
\frac{i_C R_C}{\frac{\partial \phi}{\partial t}} \gg N_C \frac{d\phi}{dt}
\]
then 2 becomes approximately
\[
E_C = i_C R_C . \quad [3]
\]
The corresponding computer equations are
\[
E_o = N_P \bar{I} + I_L R_L + I_L R_D \quad [4]
\]
and \( E_C = I_C R_C . \quad [5] \)

Rearrangement of 4 for ease of solution gives
\[
E_o - N_P \bar{I} = I_L R_L + I_L R_D \quad [6]
\]
The computer circuit to solve these equations is also shown in Fig. 1. \( R_1 \) in the computer circuit must be equal to \( R_L \) in the actual circuit so that the diode used in the computer circuit has the same characteristics as the one used in the actual circuit. It should be noted
that a resistance-capacitance network is added to amplifier No. 3. This is necessary to prevent oscillations in the circuit and is discussed further in Appendix I.

The power amplifier shown in Fig. 1 was necessary to provide enough current to saturate the sample core. \( R_d \) was used to adjust the L/R ratio of the sample core to match that of the actual core and is calculated in the following manner. The inductance of the core winding is directly proportional to the square of the number of turns, thus

\[
\frac{L_A}{L_T} = \left( \frac{N_A}{N_T} \right)^2
\]

where \( N_A = \) number of turns of actual gate winding

\( N_T = \) number of turns on test core

Then, from reference (2), if the L/R ratio in the test core is to be the same as that in the actual core, the resistances of the two windings must have the reciprocal of the above ratio

\[
\frac{L_A}{R_A} = \frac{L_T}{R_T}
\]

\[
R_T = R_A \frac{L_T}{L_A}
\]

and

\[
R_T = R_A \left( \frac{N_T}{N_A} \right)^2
\]
Then

\[ R_4 = R_T - R_o \]  \hspace{1cm} \text{[10]} 

Where

\[ R_o = \text{resistance of test core} \]

This, of course, assumes that the test core has a larger \( \frac{L}{R} \) than the actual core. In some of the circuits used, the actual core had the larger \( \frac{L}{R} \), and in these cases resistance was added in the actual circuit to get the correct value to match the test core.

For the computer circuit, the control current must be scaled to compare with that in the actual circuit being simulated. Referring to Fig. 1, the voltage across \( R_3 \), denoted \( E_c \), is used to calculate the current into the computer circuit \( I_c \). Since this current is working into the total number of turns on the computer test core, \( N_T \), and in the actual circuit the control current works into the control winding with turns \( N_C \), there must be a scale factor involving this turns ratio. Thus, if \( i_c \) represents the control current in the actual circuit, its computer simulation is represented by

\[ i_c = \frac{N_T}{N_C} \frac{E_c}{R_c} \]  \hspace{1cm} \text{[11]} 

The scaling factors \( \phi_2 \) and \( \phi_3 \) are used to adjust operational amplifier gains to account for circuits with cores having coils of different turns from the sample.
or test core. They are derived from a consideration of the voltage and current changes in the circuit when the number of turns is changed, and are given by,

\[ a_2 = a_3 = \frac{N_A}{N_T} \quad [12] \]

The other factor \( a_1 \) is used to adjust the input voltage to the computer circuit.

To simulate circuits using cores different in size from the test core, the factors \( a_2, a_3, R_4 \) and conversion from computer control voltage to actual control current must take into account the difference in areas and mean length of path between test and actual cores. The change in \( R_4 \) due to difference in size of actual and test cores was taken into account by considering that the inductance of the core winding is directly proportional to the area and inversely proportional to the mean length of path of the core.

Equation (9) becomes

\[ R_T = R_A \frac{A_T}{A_A} \frac{1_A}{1_T} \left( \frac{N_T}{N_A} \right)^2 \quad [13] \]

where

\[ A_T = \text{area of test core} \]
\[ A_A = \text{area of actual core} \]
\[ 1_T = \text{mean length of path of test core} \]
\[ 1_A = \text{mean length of path of actual core} \]

Similarly, since the magnetizing force is inversely
proportional to mean length of path, equation [11] becomes

\[ i_c = \frac{N_T}{N_A} \frac{l_A}{l_T} \frac{E_c}{R_c} \]  \[ 14 \]

The scaling factors \( a_2 \) and \( a_3 \) were calculated on the requirements of equal flux densities in the test core and actual core. The scaling factor \( a_3 \) was derived from the fact that the output current of amplifier No. 2, into \( R_2 \), represents the amount of magnetizing force \( (H) \) applied by the winding of the core. The magnetizing force \( H \) and current \( I \) are related by

\[ H = 0.4 \frac{NI}{I} \]  \[ 15 \]

Thus, in order to maintain equal magnetizing forces in the actual and test cores, the requirements for equal flux densities in the cores, the output current of amplifier No. 2 must be scaled by an additional factor of \( l_T/l_A \). Then \( a_3 \) becomes

\[ a_3 = \frac{N_A}{N_T} \frac{l_T}{l_A} \]  \[ 16 \]

The scaling factor \( a_2 \) was derived from the fact that the output voltage of amplifier No. 3 is proportional to

\[ N_T \frac{d\phi}{dt} = N_T \frac{d(B\phi A_T)}{dt} \]  \[ 17 \]

Thus, with equal flux densities in the actual and test cores, the output voltage of amplifier No. 3 must be scaled by an additional factor of \( A_A/A_T \).
Then

\[ a_2 = \frac{N_A}{N_T} \frac{A_A}{A_T} \]  \[18\]

The above theory assumes that the sample core has the same shape B-H loop as the cores in the actual circuits. This will be true in a great number of circuits where actual cores have the same ratio of outside to inside diameter as the sample core. This problem is discussed further in Section III.

II. COMPARISON OF COMPUTER AND ACTUAL CIRCUIT PERFORMANCE

The computer circuit, shown in Fig. 1, was set up to simulate a circuit using a core with three times the area, one-third the number of turns on the gate winding, and 1.61 times the mean length of path of the test core. The rectifier in the computer circuit is the same as that used in the actual circuit. The results of tests on this circuit are shown in Fig. 2. The solid curve shows the output characteristics of the actual circuit and the circled points are points taken on the computer circuit. There was considerable difference in the computer curve and actual curve. The computer curve has a slightly greater slope, a smaller minimum value of load current, and is displaced by approximately 1.9 ma control current to the left.

Pictures taken of the output voltage for the
OUTPUT CHARACTERISTICS
HALF-WAVE CIRCUIT

$E_o = 18 \text{v}$
$N_6 = 2000 \Omega$
$R_L = 500 \Omega$
$N_c = 500 \Omega$
$R_c = 20 \text{K} \Omega$
$N_L = 6000 \Omega$
$A_1 = 3 \text{A}$
$L_A = 1.61 \text{A}$

LOAD VOLTAGE - D.C. VOLTS

CONTROL CURRENT - 1934

ACTUAL CIRCUIT
COMPUTER CIRCUIT
COMPUTER CIRCUIT WITH CORRECTIONS FOR DIFFERENCE IN IQ OF ACTUAL AND TEST CORE
actual and computer circuit are shown in Fig. 3. Fig. 3b shows the output voltage of the computer and actual circuits with the same control current. The picture illustrates that the difference in output voltage for the same control current is due primarily to a difference in firing angle or point in each cycle when the core saturates. This difference was first attributed to experimental difficulties in the computer circuit. However, after further analysis, it was found that the discrepancy in actual and computer curves was due to differences in losses of the actual and test cores. These losses, primarily hysteresis and eddy currents, effectively widen the dynamic B-H loop. This results in a greater magnetizing force applied by the winding than is actually present in the core to cause flux change. Because of this effect, the actual core requires more control current than the test core for the same flux density. Also, the minimum value of load current will be greater for the actual core than the test core, since the minimum load current is proportional to the magnetizing current or losses in the core. Fig. 3a, a picture of the output voltage of the actual and computer circuit with control currents adjusted to give equal output voltage, illustrates this fact. The exciting current
FIGURE 3
Wave shapes in computer and actual circuits with different sized cores

\[ 3N_G = N_T \]
\[ A_A = 3A_T \]
\[ l_A = 1.61 l_T \]
\[ R_L = 500 \text{ ohms} \]
\[ E_o = 18v \]

FIGURE 4
Wave shapes in computer and actual circuits with different sized cores

\[ 12 N_G = N_T \]
\[ A_A = 3A_T \]
\[ l_A = 0.76 l_T \]
\[ R_L = 500 \text{ ohms} \]
\[ E_o = 5v \]
or current flowing before saturation is definitely larger for the actual circuit.

The coercive forces \( H_c \) of the actual and test cores were then measured, using a B-H loop displayed on an oscilloscope, to obtain the ratio of losses of the two cores. This ratio \( \frac{H_c_{\text{actual}}}{H_c_{\text{test}}} \) was found to be 1.45 for the cores used. Therefore, to account for differences in losses, the scaling factor \( a_1 \) must be reduced by 1.45. However, after saturation of the core, the hysteresis and eddy current losses are negligible since the flux remains constant. Therefore, the scaling factor should not be changed after saturation of the core. This was accomplished by a shunt, back biased diode placed across the input resistor to amplifier No. 2 as shown in Fig. 5. The resistor and battery in series with the diode were chosen such that the gain of the amplifier is increased by 1.45 for voltages greater than the voltage present during the exciting or magnetizing period.

The effect of the greater losses of the actual core on the control current was accounted for by adding a constant value of control current to each point on the computer curve. The amount of control current to be added was determined in the following manner:
Computer circuit for simulation of half-wave circuit, with corrections in gate current for losses in actual core different from those of test core.

Figure 5
Hc_A \over Hc_T = 1.45

\therefore \quad \frac{Ic_A}{Ic_T} = 1.45 \quad \text{(to obtain Hc)}

Hc is obtained at minimum output voltage (5) when

Ic_T = 4.2 \text{ ma (taken from circled points in Fig. 2.)}

Ic_A = 4.2 \times 1.45 = 6.1 \text{ ma, and the amount of control current to be added is}

6.1 - 4.2 = 1.9 \text{ ma.}

The dashed curve, shown in Fig. 2, is the resultant curve with corrections for difference in losses of the actual and test cores. This curve is in good agreement with the actual curve.

Figures 6 and 7 show the results of tests made on two other sizes of cores. Again, the solid curve is for the actual circuit, the circled points are for the computer circuit and the dashed curve is the computer curve corrected for differences in losses. These computer curves are also in fair agreement with the actual curves. However, the corrected computer and actual curves exhibit a marked difference at high values of output voltage, where other results have been in close agreement. The number of turns on the gate windings of the actual cores was not known and therefore was estimated from the voltage to cause saturation and the size of the core. An error in
OUTPUT CHARACTERISTICS
HALF-WAVE CIRCUIT.

$E_a = 5 \text{ V}$
$N_b = 500 \Omega$
$R_L = 500 \Omega$
$N_c = 1207 \Omega$
$R_C = 5 \text{ k}\Omega$

DIODE - 1N67

$N_T = 6000 \Omega$
$A_A = 5 A_T$
$I_A = 0.757 A_T$

ACTUAL CIRCUIT
COMPUTER CIRCUIT
COMPUTER CIRCUIT WITH CORRECTIONS FOR DIFFERENCE IN % OF ACTUAL AND TEST CIRCUIT

LOAD VOLTAGE - D.C. VOLTS

CONTROL CURRENT - mA
Fig. 7

Output Characteristics
Half-Wave Circuit

- $E_0 = 5 \text{ v}$
- $N_k = 1000 \text{ T}$
- $R_L = 500 \Omega$
- $N_l = 100 \text{ T}$
- $R_C = 5 \text{ k}\Omega$
- Diode - 1N67

- $N_T = 6000 \text{ T}$
- $A_+ = 4.5 A_1$
- $L_A = 0.757 \mu\text{T}$

Load Voltage - D-C Volts
Control Current - mA

- Actual Circuit
- Computer Circuit
- Computer Circuit with corrections for difference in $H_C$ of actual and test core
number of turns on the gate winding will affect not only the scaling factors $a_2$ and $a_3$ shown in Fig. 1, but also the resistance $R_4$ added in series with the test core. Since at high values of output voltage the reactance of the core is small compared to $R_4$, an error in $R_4$ will have a greater effect at high values of output voltages than at low to medium values. The difference in the actual and computer curves shown in Figures 6 and 7 at high values of output voltages, therefore, could be accounted for by error in estimation of the number of turns on the gate winding. Fig. 4 shows the uncorrected computer and actual circuit output voltages for the circuit whose characteristics are shown in Fig. 6. As in Fig. 3, Fig. 4 illustrates that the output voltage of the computer circuit, uncorrected for difference in losses of the actual and test cores, and the actual circuit differ primarily in firing angle.

III. ROLE OF B-H LOOP OF ACTUAL AND TEST CORES

The theory described in Section I assumes that the sample or test core has the same shape B-H loop as the actual core, i.e., equal slopes before and after saturation. This assumption was valid for the cores used in the circuits of Section II. However, there are many possible cores whose B-H loops have
different slopes from that of the sample core used. It has been shown (4) that the geometry of the core plays an important part in the slope of the B-H loop. The ratio of outside to inside diameter of the core is important in that it is a measure of the amount of deviation away from the mean length of magnetic path in the core. Since the saturation of the core depends on the ampere-turns per unit length of magnetic path, the core material at the extreme inside and outside of the toroid core saturates at different current levels from that calculated using the mean length of path. As the ratio of outside to inside diameter of the core is increased, there is a greater difference between these current saturation levels. This decreases the slope of the dynamic hysteresis loop, giving the effect of a poorer core material. Therefore, cores with the same ratio of outside to inside diameter have B-H loops with equal slopes. The width of the loop, however, will vary with the losses and thus the volume of the core. A sample core, then, can be used to simulate any circuit using cores having the same ratio of outside to inside diameter as the sample core.

Oscilloscope pictures of B-H loops of several cores are shown in Fig. 8. Fig. 8a shows B-H loops
FIGURE 8

a. B-H loops of two cores having nearly equal losses and ratios of OD/ID equal to 1.25 and 1.5.

b. B-H loops of two cores having nearly equal ratios of OD/ID and radically different losses.
of two cores with ratios of OD/ID equal to 1.25 and 1.5, and losses very nearly the same. The difference in slope of the B-H loop, due to the difference in ratio of OD/ID, can be seen from this picture. Fig. 3b shows the B-H loops of two cores with very nearly the same ratio of OD/ID but with losses radically different. It can be seen that, although there is considerable difference in the width of the loops, the slopes are very nearly the same.

IV. COMPUTER CIRCUITS FOR MORE COMPLEX MAGNETIC AMPLIFIER CIRCUITS

A. Full-wave center tap circuit with D-C output and suppressed even harmonics

The actual circuit is shown in Fig. 9 and is described by the following equations.

For the gate circuit,

\[ E_0 = N G_b \frac{d\phi_b}{dt} + i_b R_L + i_b R_D + i_a R_L \]  \hspace{1cm} [19]
\[ -E_0 = N G_a \frac{d\phi_a}{dt} + i_a R_L + i_a R_D + i_b R_L \]  \hspace{1cm} [20]

and \[ i_L = i_a + i_b \]  \hspace{1cm} [21]

For the control circuit,

\[ E_c = i_c R_c \]  \hspace{1cm} [22]

The corresponding computer equations are,

\[ E_0 = N_1 P_{\phi L} - X_2 R_L = X_1 R_L + X_1 R_D \]  \hspace{1cm} [23]
**Figure 9**
Center Tapped Full-Wave Circuit with D.C. Output

**Figure 10**
Full-Wave Circuit with A.C. Output, Free Even Harmonics
The computer circuit to solve these equations is shown in Fig. 11. If both cores in the actual circuit have the same size and number of turns and the two test cores have the same size and number of turns, then the scaling factors

\[ a_1 = a_3 \quad \text{and} \quad a_2 = a_4 \]

From equations [16] and [18]

\[ a_1 = \frac{N_A}{N_T} \frac{A_A}{A_T} \]

and \[ a_2 = \frac{N_A}{N_T} \frac{1_T}{1_A} \]

The conversion factor for control current, from equation, from equation [14] is

\[ i_c = \frac{N_T}{N_G} \frac{1_A}{1_T} \frac{1}{10K} E_c \]

B. Full-wave circuit with A-C output, free even harmonics

The actual circuit is shown in Fig. 10 and is described by the following equations

For the gate circuit,

\[ E_0 = NGA \frac{d\phi_A}{dt} + i_aR_D + i_aR_L + i_bR_L \]
FIGURE 11

Computer circuit for actual circuit shown in Figure 9.
The corresponding computer equations are as follows:

For the gate circuit,

\[ E_0 = N_{gb} \frac{d\phi}{dt} + i_b R_R + i_a R_L + i_a R_L \]  
\[ i_a + i_b = i_L \]  

and the control circuit,

\[ E_C = N_{gb} \frac{d\phi}{dt} + N_{gb} \frac{d\phi}{dt} + i_c R_L \]  

The computer circuit to solve these equations is shown in Fig. 12. If the two actual cores have the same size and number of turns, and the two test cores have the same size and number of turns, then the scaling factors \( a_1 = a_3, a_2 = a_4, \) and \( a_5 = a_6. \) From equations [16] and [18] the factors \( a_1, a_2, a_5 \) and \( a_7 \) are given by,

\[ a_1 = \frac{N_G}{N_T} \frac{AA}{AT} \]  
\[ a_2 = \frac{N_G}{N_T} \frac{LT}{LA} \]  
\[ a_5 = \frac{N_G}{N_T} \frac{AA}{AT} \]  
\[ a_7 = \frac{N_G}{N_T} \frac{LT}{LA} \]
Figure 12

Computer circuit for actual circuit shown in Figure 10
The conversion factor for control current is taken into account by a7 so that $E_c$ measured on the computer is the same as the actual circuit.

Although no data has been taken on the above computer circuits, due to time and computer limitations, they are included in this report as examples of how more complex magnetic amplifier circuits can be set up on a computer and simulated.

V. RECOMMENDATIONS AND SUGGESTIONS FOR FUTURE STUDIES

Although the primary objectives of this study were realized, much more work could be done along the same lines. From the standpoint of general magnetic amplifier circuitry, with the great variety of practical circuit arrangements, this study has barely scratched the surface. It has, however, provided the basic groundwork for computer studies of magnetic amplifiers and has established a starting point from which more fruitful studies could be made. For any future studies which may be undertaken, there are several improvements which should be made in the test equipment which was designed and built during the course of this study. The first and most important item is chopper stabilization of the operational amplifiers. In all of the computer simulations it was necessary to have the amplifiers
zeroed very accurately at all times. This was especially true for the amplifier in the core simulation circuit. Also, in connection with the core simulation circuit, a better test core should be obtained. It should be wound with as large a wire as possible so that a maximum L/R ratio is obtained, thus allowing this ratio to be varied simply by adding the necessary resistance in series with the winding. If possible, a better method of suppressing oscillation in the circuit should be devised, since the method being used does affect the simulation accuracy somewhat. Any future study should probably include the use of diode simulation circuitry in the computer, since in many cases the diode needed in an actual circuit would not be satisfactory for the computer circuit. This would also allow a study to be made of the manner in which the diode characteristics affect the behavior of a magnetic amplifier circuit.
CONCLUSIONS

Although all of the main objectives of this study were accomplished, it is felt that considerably more work could still be done within the scope of these objectives. This work is primarily in the area of simulation of more complex types of magnetic amplifier circuits. It appears that the use of an actual core in conjunction with an operational amplifier is the best method for simulating the core material. It should be noted that any future studies would benefit from the use of better power amplifiers and a better test core in the simulation of the core material. For studies of more complex circuits it would also be necessary to use a diode simulation circuit so that the operating characteristics of this component could be varied to match any diode which the actual circuit might require.

The methods developed for simulating circuits having different sized cores from that used in the computer circuit appears to be reasonably good; however, considerably more work needs to be done before any definite conclusions can be reached. It is felt that the method outlined in this report for taking into account losses of the actual core different from the test core will hold for other circuits as well. However, the data in this report does not provide enough evidence for all
circuits, to substantiate this feeling. It is this aspect of the study that needs more work. It is believed that the reasoning described in the development of the method is sound and would serve as a basis for more experimental work to establish exact procedures for converting computer results into predictions of more complicated actual circuit behavior.
APPENDIX I

It was pointed out in the procedure that it was necessary to devise some method for preventing oscillation in the computer circuit. The use of a condenser across the operational amplifier in the core simulation circuit stopped the oscillation, but it also had considerable effect on the circuit operation. In order to get data of sufficient accuracy, some other stabilization method was needed. The final stabilization network used is shown in Fig. 1. At the time, it was assumed that this network was simply reducing the high frequency response of the circuit enough to prevent oscillation, and since this stabilization problem was not directly related to the primary objectives of the thesis study, no effort was made to analyze the circuit completely. However, after the work pertaining to the main objectives of the thesis had been finished, it was felt that some attention should be given to this problem.

Fig. 13 shows a frequency response curve for the core simulation circuit with and without the stabilization network. For this data, the input was kept sufficiently low that core would not saturate. These curves indicate that the circuit response is primarily controlled by the self-resonance of the core, and the addition of the
Figure 13: Frequency Response of Core Simulation Circuit

- Without Stabilization Network
- With Stabilization Network

Gain vs. Frequency (CPS)
stabilization network had the effect of lowering the Q of the circuit. The resonance peak, without the stabilization network, occurs at about 3750 cps, and measurements on the core itself show that the self-resonant frequency is about 3800 cps. The instability problems can then be attributed to capacitive loading of the operational amplifier. Above the resonant frequency of the coil, the feedback is capacitive, so that at high frequencies the circuit acts as an integrator. Similar stability problems have been experienced with the integrators on the main computer. The stabilization network then, as assumed, provides a break in the high frequency response of the circuit so that the open loop gain is reduced below unity before a phase shift of 180° occurs, thus preventing oscillations.
REFERENCES


