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Author
Windsor, Alfred A.

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University of California

Ernest O. Lawrence Radiation Laboratory

Berkeley, California

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A new circuit for the simple series-connected saturable reactor corrects some fundamental limitations in stability, response time, and signal-to-noise ratio. The conventional saturable reactor has found wide application as a current transducer but is often thought of as a noisy device and vulnerable to stray magnetic fields. The new transformer maintains a continuous ampere-turn balance between primary and secondary and excels conventional circuits when stability and wide frequency response are important.

I. INTRODUCTION

In principle, direct current can be transformed if the transformer core doesn't become saturated. Unfortunately present cores have a limited saturation flux density and cannot couple very well between primary and secondary unless some means is found to keep the core in the high-permeability region of the B-H (hysteresis) loop.

The conventional transductor or series-connected saturable reactor maintains coupling between primary and secondary by cycling two transformers in and out of saturation to alternate the load current at the frequency of excitation. This method of current measurement has found wide acceptance and has been in use for nearly fifty years.[1, 2, 3] Many users of current shunts would be surprised by the accuracy and stability of the transductor having modern square-loop core material.

Second-order errors appear, however, when the transductor load current reverses. Both cores become saturated and remain saturated until the saturated reactances have been charged to the opposite polarity. This switching process uncouples the primary from the secondary, and the time required for coupling to be restored depends upon the magnitude of saturated reactance. The switching time represents an error; the error is 100% when the load current passes through zero. Momentary interruptions of the ampere-turn balance are always characteristic of the series-connected saturable reactor. The average output current is also affected because stray magnetic fields vary the saturated reactance and cause a negative magnetic-field coefficient (see Fig. 1).

II. THE NEW TRANSFORMER PRINCIPLE

To avoid saturation of a transformer core, there must be an ampere-turn balance between primary and secondary, and the net flux or volt-second area must be zero. Fig. 2 illustrates an experiment in which an ampere-turn balance is set up in the transformer T_2. D_1 is a half-wave rectifier and D_2 is a free-wheeling clamp on the secondary. E_s is increased until the secondary ampere-turns just balance the primary--a condition in which the positive area A_1 equals the negative area A_2 of E_{T2}. The secondary voltage of T_2 is $I_L\,(R_{T2} + R_L)$ as long as D_2 conducts. When E_s reverses, D_1 conducts and current shifts completely from D_2 to D_1 while E_s exceeds $I_L\,R_{T1}$. D_2 blocks at ϕ_1 and the voltage of T_2 reverses at ϕ_2. Area A_1 is formed from ϕ_2 to $(\pi - \phi_2)$. Area A_2 is formed from $(\pi - \phi_2)$ to $(2\pi + \phi_2)$.

Regardless of the care with which E_s is adjusted, the minor hysteresis loop of core 2 occupies a very uncertain flux level. When a square-loop core material is used, E_s must be precisely related to I_c or the core will saturate. The theoretical relationship between primary current and E_s can be found by equating A_1 and A_2 (Appendix I). When E_s is low A_1 is low, and the core saturates on every cycle from current in the primary. High E_s saturates the core during the interval of excess volt seconds of area A_1. If E_s were regulated or adjusted to every control current, the core would always remain in a high-permeability region, but this does not seem very practical.

The addition of transformer T_1 permits a range of control current to be balanced by a fixed supply voltage. T_1 is connected in series with T_2 as shown in Fig. 3. E_s is made high and T_2 couples the primary current to the load (as shown in Fig. 2) until the excess volt seconds begin to take core 2 into saturation. T_1 has fewer turns than T_2, thus demanding a higher secondary current for the same core excitation as core 2. As core 2 approaches saturation, its exciting current becomes high enough to bring core 1 into the high-permeability region. Figure 4 shows the voltage shifting from T_2 to T_1 at point C when transformer 2 reaches saturation. Figure 5 shows the hysteresis loop slightly displaced by the unequal secondary ampere-turns. Diode 2 remains blocked after E_s reverses at b (Fig. 4) and the ampere-turn balance is maintained by T_1 and T_2 in series until core 1 saturates at point f. At point f, current shifts back to diode 2 because E_s blocks diode 1 when core 1 saturates. After E_s reverses at point a, current shifts to diode 1 and blocks diode 2 when the loop current i_1 tends to exceed i_2.

Another way to understand the same sequence of operation is to consider T_1 first.[4] T_1 functions as a half-wave saturable reactor operated in the linear mode. Normally i_1 equals I_L, and diode D_2 is not conducting. During the time saturable reactor T_1 is saturated, i_1 is less than I_L, and the free wheeling diode D_2 provides an independent path for current I_L. Thus even though the ampere turn balance is lost with respect to T_1, the secondary of T_2 is in effect short circuited by the diode D_2 and the current I_L depends only upon the primary current I_c.

There are some distinct advantages in having the primary coupled to the load at all times. First, the noise amplitude is smaller and better defined. Ripple current due to E_s equals the coercive current corresponding to the width of the hysteresis loop.

Second, environmental changes in the B-H loops have a very small effect because the average coercive force almost adds to zero for an unsaturated square-loop core material.

III. DESIGN FACTORS

1. The linear relationship between primary and secondary currents will hold for only one polarity of primary current. To measure a reverse current D_1 and D_2 must be reversed.

2. If R_1 is omitted, the secondary current becomes excessive when the primary current is zero. R_1 prevents T_1 from going into the magnetic-amplifier mode at zero control current.

3. Core 1 must have enough cross-section area to absorb all of E_s plus a safety factor, while core 2 requires about half as much, since it cannot absorb more than $I_L(R_L + R_{T2})$ for a little more than half a cycle.

4. Fig. 5 shows that E_s max is limited by the volt-second capability of T_1 at low control currents. E_s min must be πI_L max $(R_{T2} + R_L)$ plus a saftey factor to accommodate for variations in E_s and to accommodate for the leakage reactance of T_1.

IV. A THIRD TRANSFORMER IS ADDED

Many applications require the ability to detect a wide range of frequency components and a wide range of amplitudes.[5,6] An alternating-current transformer may be added which serves the purpose of filtering out the coercive current ripple without narrowing the frequency response. If the first two cores are represented by a hypothetical direct current transformer with a coercive current generator at its output, the equivalent circuit of a low-noise unit can be represented by Fig. 6. An analysis of the circuit (Appendix II) shows the voltage across N_3 to be coercive ripple. Also i_x' is reduced by a factor of $(R_2/\omega_x L_3)$ and

$N_3/N_2 = (R_2 + R_3)/R_2$. Figure 7 shows a practical circuit which is used to measure large current pulses in the magnet of a particle accelerator. The transfer characteristic is about 3 db down at 100,000 cps, and peak direct current goes to 5000 amp. The noise in the output signal referred to the input current is less than 0.5 amp. The long-term stability, checked by a flux gate magnetometer, is better than 0.02%.

At first no attempt was made to use the wide-band direct-current transformer for absolute measurements, but subsequent checks have shown that it can be trimmed to measure with an error of less than 0.1% over a range of ten to one. Compensation for linearity, shown by Fig. 8, is achieved in the load circuit. A fixed bias derived from E_s plus an increased load resistance of about 0.2% was used to obtain the flatter curve. The inherent high-impedance secondary circuit makes the calibration independent of circuit constants except when the third transformer is combined through a resistance-mixing network. R_2 of Fig. 7 may contain a copper resistor to correct for variations in the resistance of T_3.

When the expected transient currents contain relatively few ampere seconds, the third transformer may be capacitively coupled. R_2 is replaced by a capacitor which offers low impedance to coercive currents and N_3 is made equal to N_1. Resistance-temperature effects of T_3 are then unimportant because of the constant-current characteristics of the T_1 - T_2 combination.

CONCLUSIONS

The circuits described have been useful in measuring and controlling large currents. Particle accelerators and auxilary equipment used in physics research often require precise measurements in adverse magnetic environments. The circuit shown in Fig. 3 can be a conventional transductor with minor modifications added, and has found the widest general use. "Three-core" circuits are more expensive but provide isolation, high output voltage, and wide frequency response at about the price of a low-voltage shunt. The wide-band transformers have made possible measurements that would not have been considered for the conventional transductor.

ACKNOWLEDGMENTS

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APPENDIX I

The areas are expressed by,

$$a_1 = 2\int_{\phi_2}^{\pi/2} E_s \sin\phi\, d\phi - (\pi - 2\phi_2)\, I_L\, (R_{T1} + R_{T2}\ R_L),$$

$$a_1 = 2\, E_s \cos\phi_2 - I_L\, (\pi - 2\phi_2)\, (R_{T1} + R_{T2} + R_L),$$

$$a_2 = (2\phi_2 + \pi)\, I_L\, (R_{T1} + R_{T2} + R_L) - 2\int_{\phi_1}^{\phi_2} E_s \sin\phi\, d\phi - (\pi + 2\phi_1)\, I_L\, R_{T1} \text{ and,}$$

$$a_2 = 2\, E_s \cos\phi_2 - 2\, E_s \cos\phi_1 + I_L\, (\pi + 2\phi_2)\, (R_{T1} + R_{T2} + R_L) - I_L\, R_{T1}\, (\pi + 2\phi_1).$$

Let $a_1 = a_2$, then

$$2\, E_s \cos\phi_1 = 2\pi\, I_L\, (R_{T1} + R_{T2} + R_L) - I_L\, R_{T1}\, (\pi + 2\phi_1),$$

$I_L = I_c/N_2$ condition for amp-turn balance,

and $E_s = I_c\, \pi/N_2 \cos\phi_1\, [R_{T2} + R_L + R_{T1}\, (1/2 - \phi_1/\pi)]$

If ϕ_1 is small,

$$E_{s\ (RMS)} = 2.22\, I_L\, (R_{T2} + R_L + R_{T1}/2).$$

APPENDIX II

Given,

i_1 = Primary current,

i_x = D-C trans. coercive ripple,

i_3 = Metering current,

and i_2 = D-C trans. current,

then

$$i_3 = \left(\frac{N_1}{N_3}\right) i_1 \left(\frac{j\omega L_3}{R_2 + R_3 + j\omega L_3}\right) + i_2 \left(\frac{R_2}{R_2 + R_3 + j\omega L_3}\right) + i_x \left(\frac{R_2}{R_2 + R_3 + j\omega_x L_3}\right).$$

Neglecting i_x and letting $i_1 N_1 = i_2 N_2$, then

$$i_3 N_3 = i_1 N_1 \left[\left(\frac{N_3}{N_2}\right) \frac{R_2}{R_2 + R_3 + j\omega L_3} + \frac{j\omega L_3}{R_2 + R_3 + j\omega L_3}\right].$$

Let $i_1 N_1 = i_3 N_3$ when $\omega = 0$ to prevent core 3 from saturating. Then

$$\frac{N_3}{N_2} = \frac{R_2 + R_3}{R_2}.$$

Substituting for $\frac{N_3}{N_2}$ shows $\frac{R_2 + R_3}{R_2 + R_3 + j\omega L_3} + \frac{j\omega L_3}{R_2 + R_3 + j\omega L_3} = 1$, and

$i_3 N_3 = i_1 N_1$ for all frequencies when $i_1 N_1 = i_2 N_2$ and when $\frac{N_3}{N_2} = \frac{R_2 + R_3}{R_2}$.

Also the voltage across L_3 is zero for all frequencies if R_3 is assumed to combine the internal resistance of L_3 with the load resistance because

$i_3 N_3 = i_2 N_2$, $\frac{i_2}{i_3} = \frac{N_3}{N_2} = \frac{R_3}{R_2} + 1$, $i_2 - i_3 = \frac{i_3 R_3}{R_2}$ and

the sum of currents into the junction of R_2 and R_3 must be zero. Then we find

$$\frac{E_{R_2}}{R_2} = i_2 - i_3 = \frac{i_3 R_3}{R_2}, \quad E_{L_3} = E_{R_2} - i_3 R_3 = 0,$$

and $E_{L_3} = 0$ for all frequencies.

It has been shown that an ideal direct-current transformer combined with an alternating-current transformer has a flat frequency response. But the purpose of L_3 is to reduce

i_x. We see from the first equation of Appendix II that the ripple component of current through L_3 is $i_x \left[1 \Big/ \left(\frac{N_3}{N_2} + j \frac{\omega_x L_3}{R_2} \right) \right]$.

If $\omega_x L_3/R_2$ is made large compared with N_3/N_2, then i_x is reduced by a factor of $R_2/\omega_x L_3$. In the solution to the first equation of Appendix II it is seen that the contribution of the alternating-current transformer is shown by $i_3 N_3 = i_1 N_1 \left[\frac{\omega^2 L_3^2}{\omega^2 L_3^2 + (R_2 + R_3)^2} \right]$, and the direct-current transformer by $i_3 N_3 = i_1 N_1 \left[\frac{(R_2 + R_3)^2}{\omega^2 L_3^2 + (R_2 + R_3)^2} \right]$.

Equating the alternating- and direct-current transformer expressions shows the cross-over frequency to be $\omega = (R_2 + R_3)/L_3$.

FIGURE LEGENDS

Fig. 1. A family of B-H loops for delta-max or orthonol core material shows reduced B_{max} when a transverse field is applied.

Fig. 2. Basic circuit and principle of operation.

Fig. 3. A practical circuit with B-H loops for low control current.

Fig. 4. Hysteresis loops showing how core 1 resets core 2 when N_2 is greater than N_1.

Fig. 5. Principle of operation--2000-amp saturable transformer.

Fig. 6. Equivalent circuit of wide-band direct-current transformer.

Fig. 7. Wide-band direct-current transformer.

Fig. 8. Graph of percent error vs primary current for wide band D-C transformer.

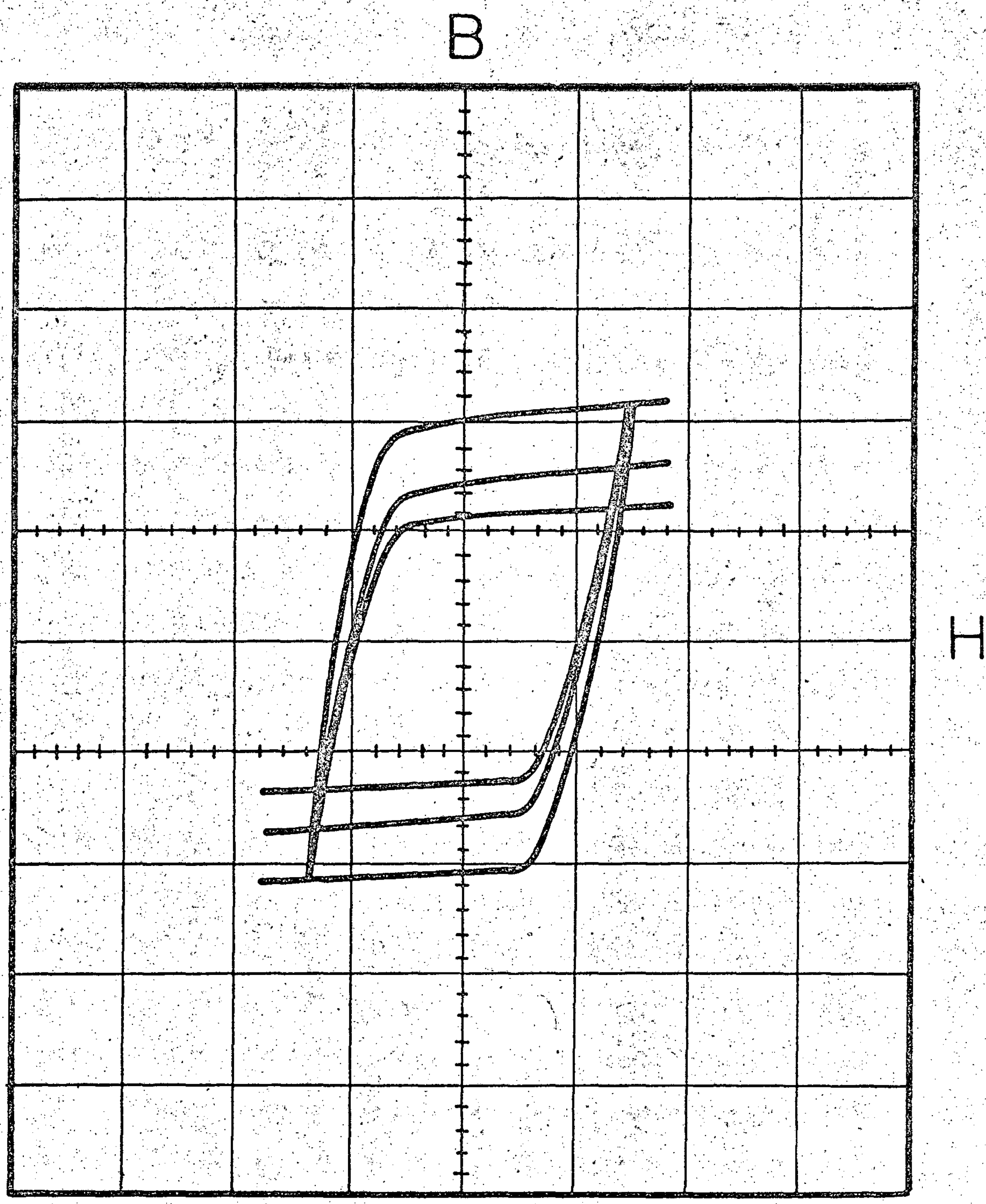

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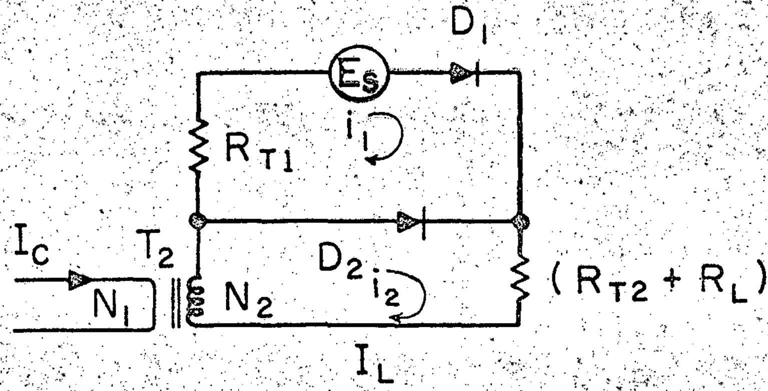
Fig. 1



$$I_C N_1 = I_L N_2$$

$$E_S \approx \pi I_L (R_{T2} + R_L)$$

when $a_1 = a_2$

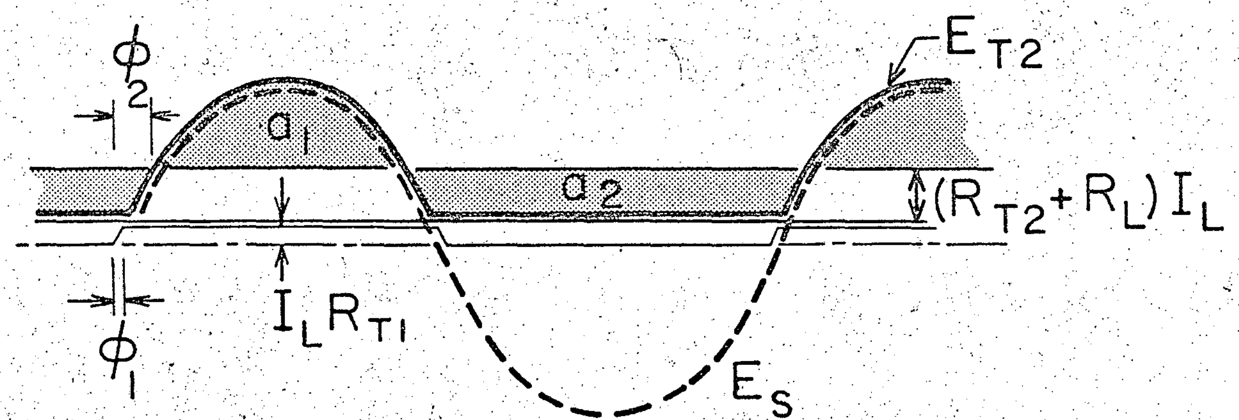


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Fig. 2

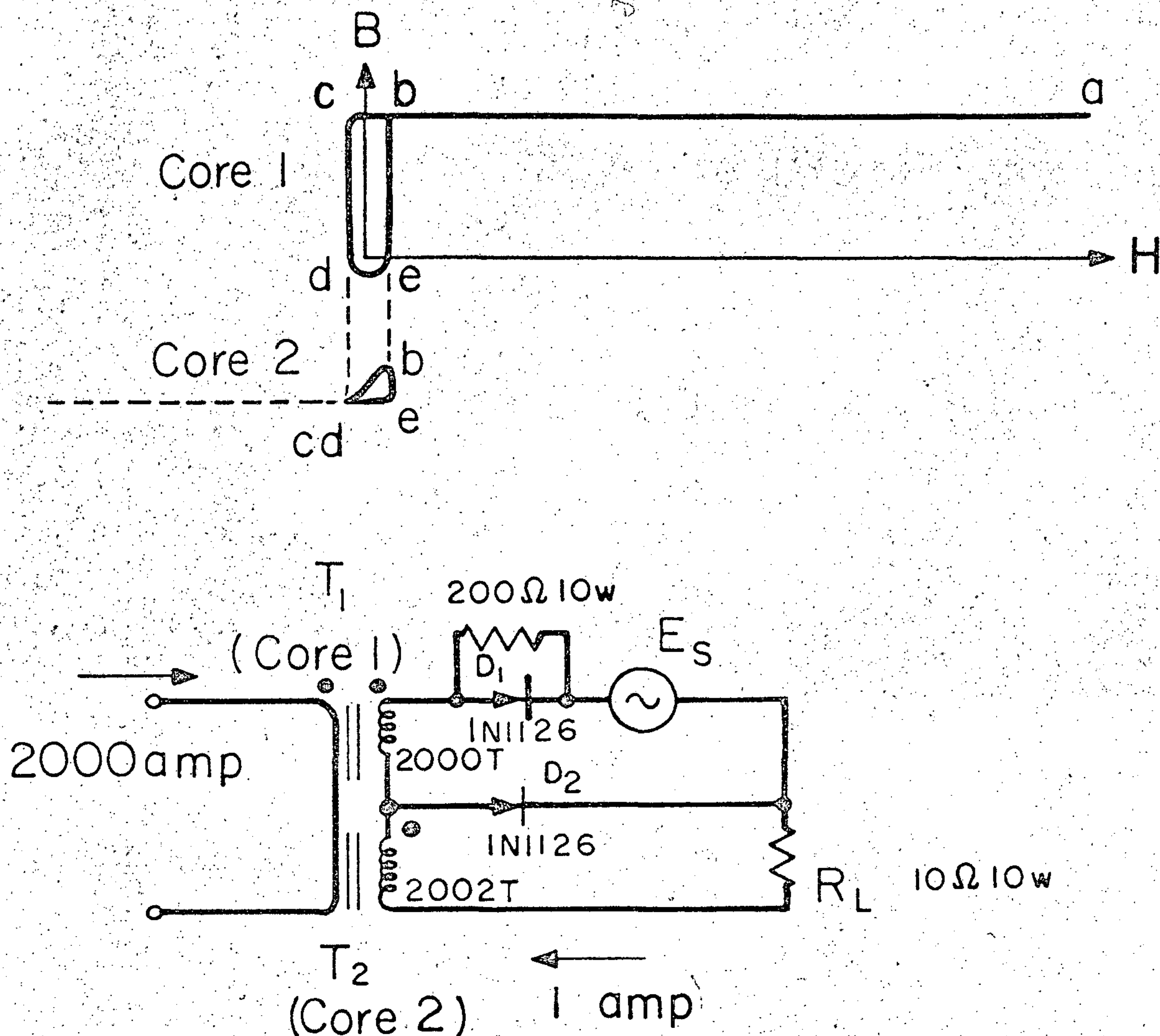

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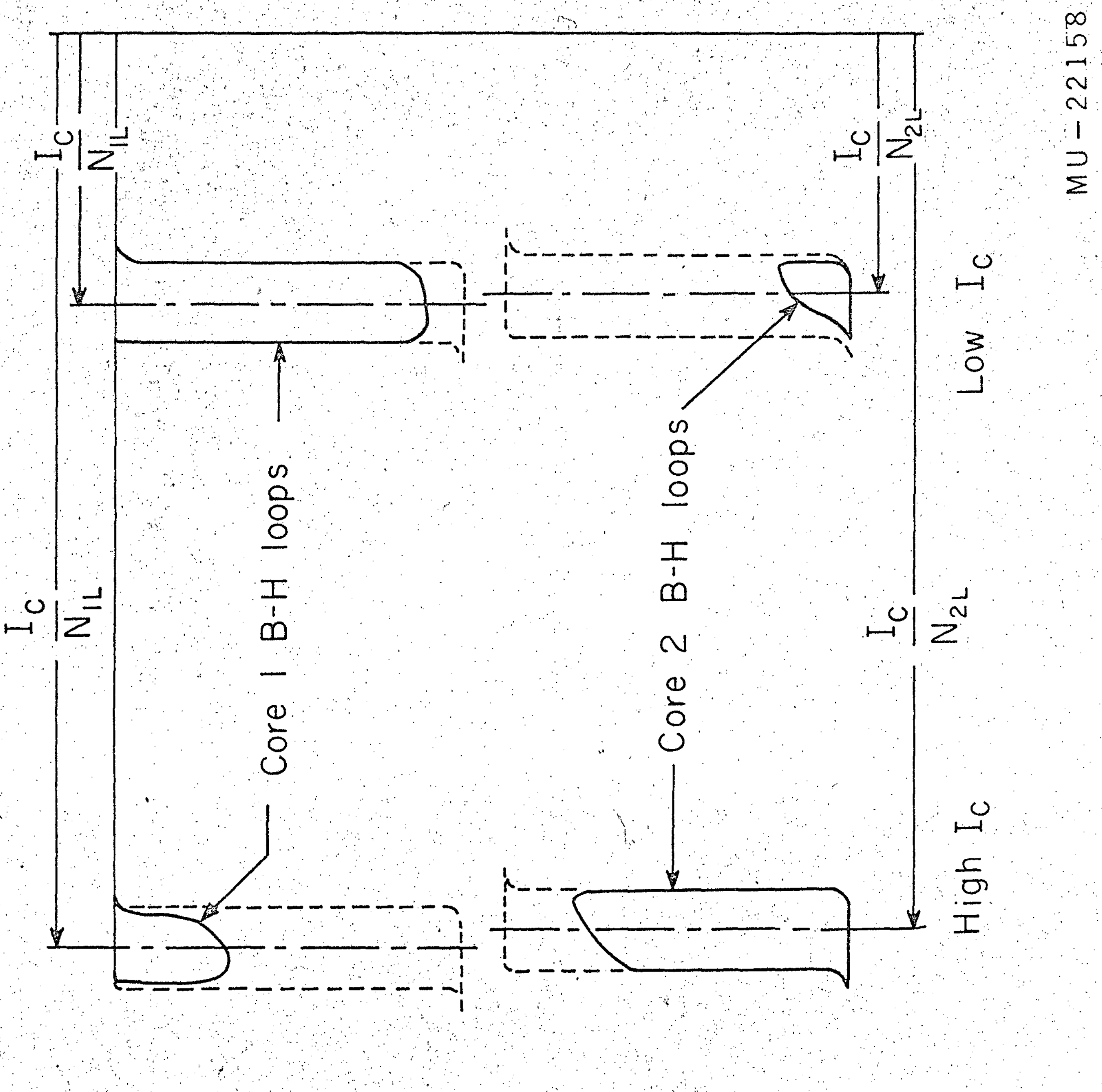
Fig. 3



I_c/N_{1L}
I_c/N_{1L}
Core 1 B-H loops
I_c/N_{2L}
I_c/N_{2L}
Core 2 B-H loops
Low I_c
High I_c
MU-22158

Fig. 4

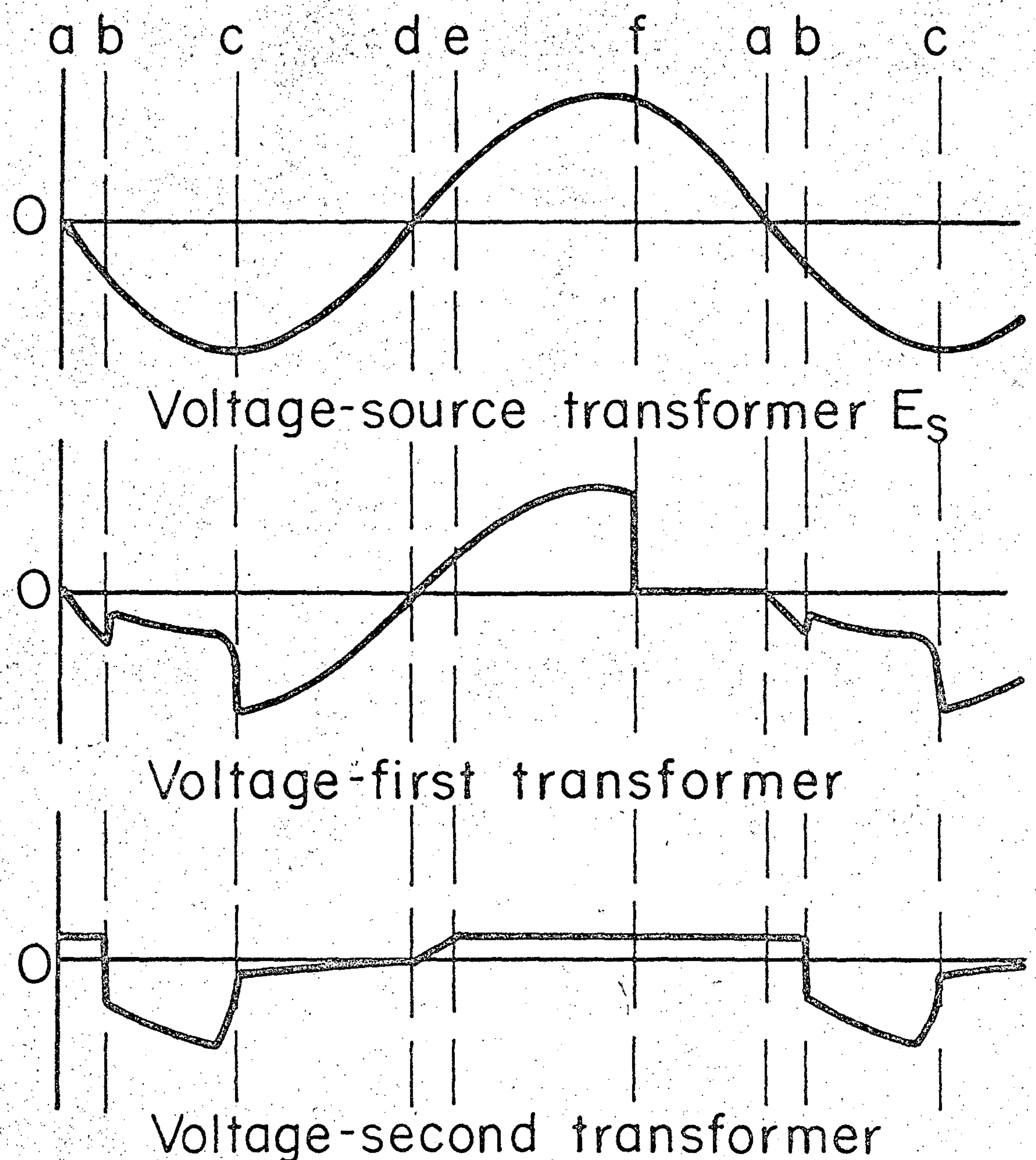


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Fig. 5

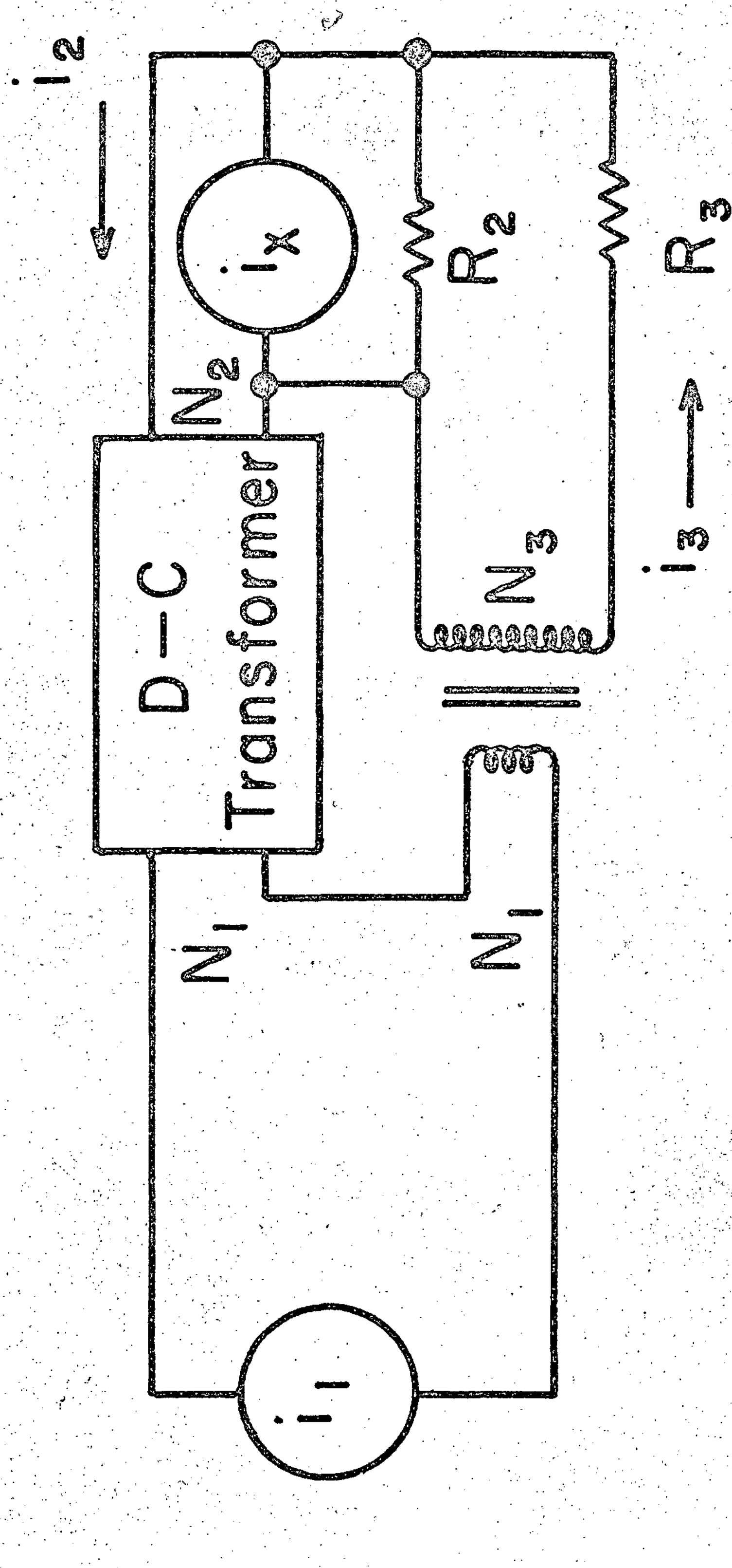

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Fig. 6

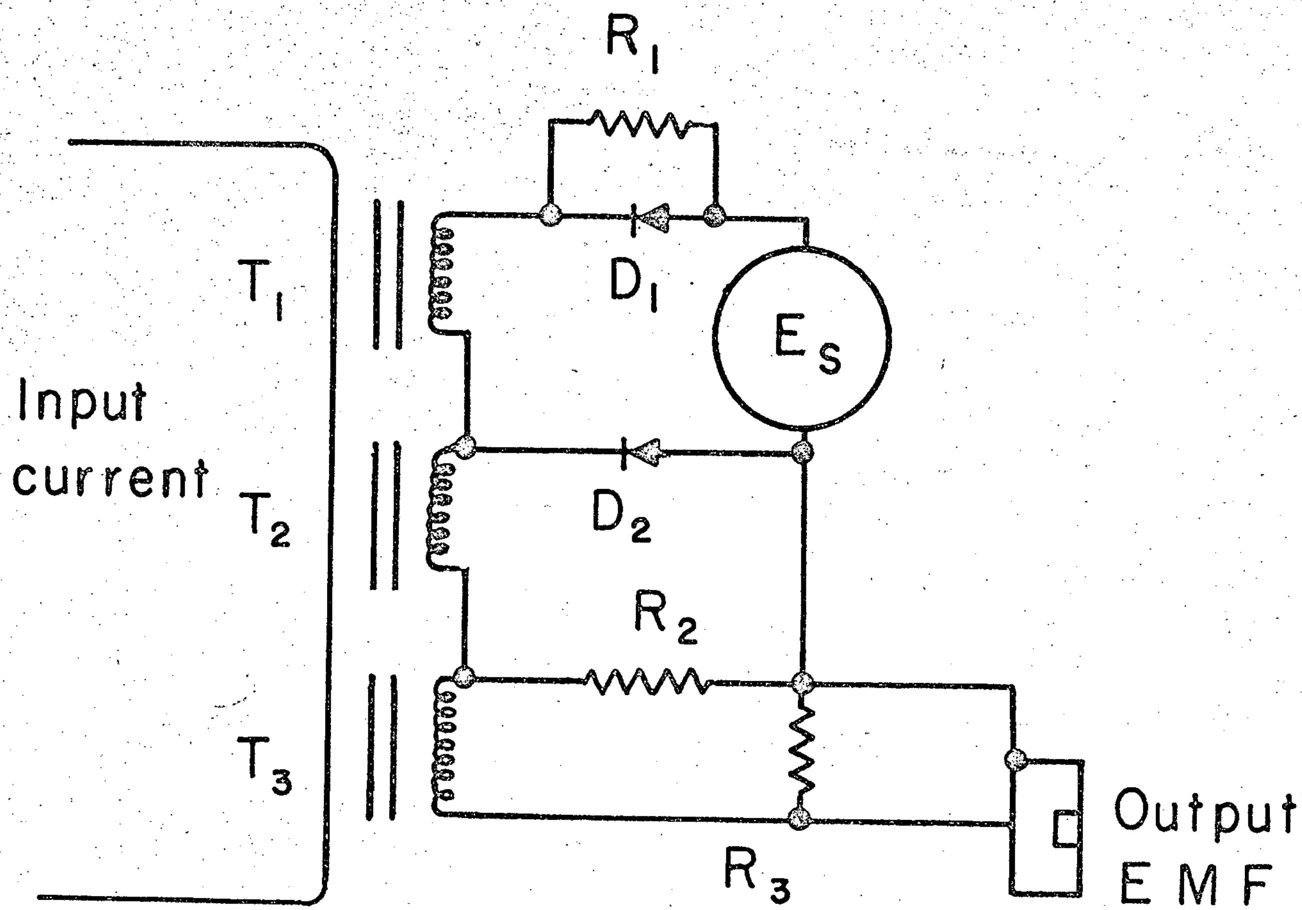

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Fig. 7

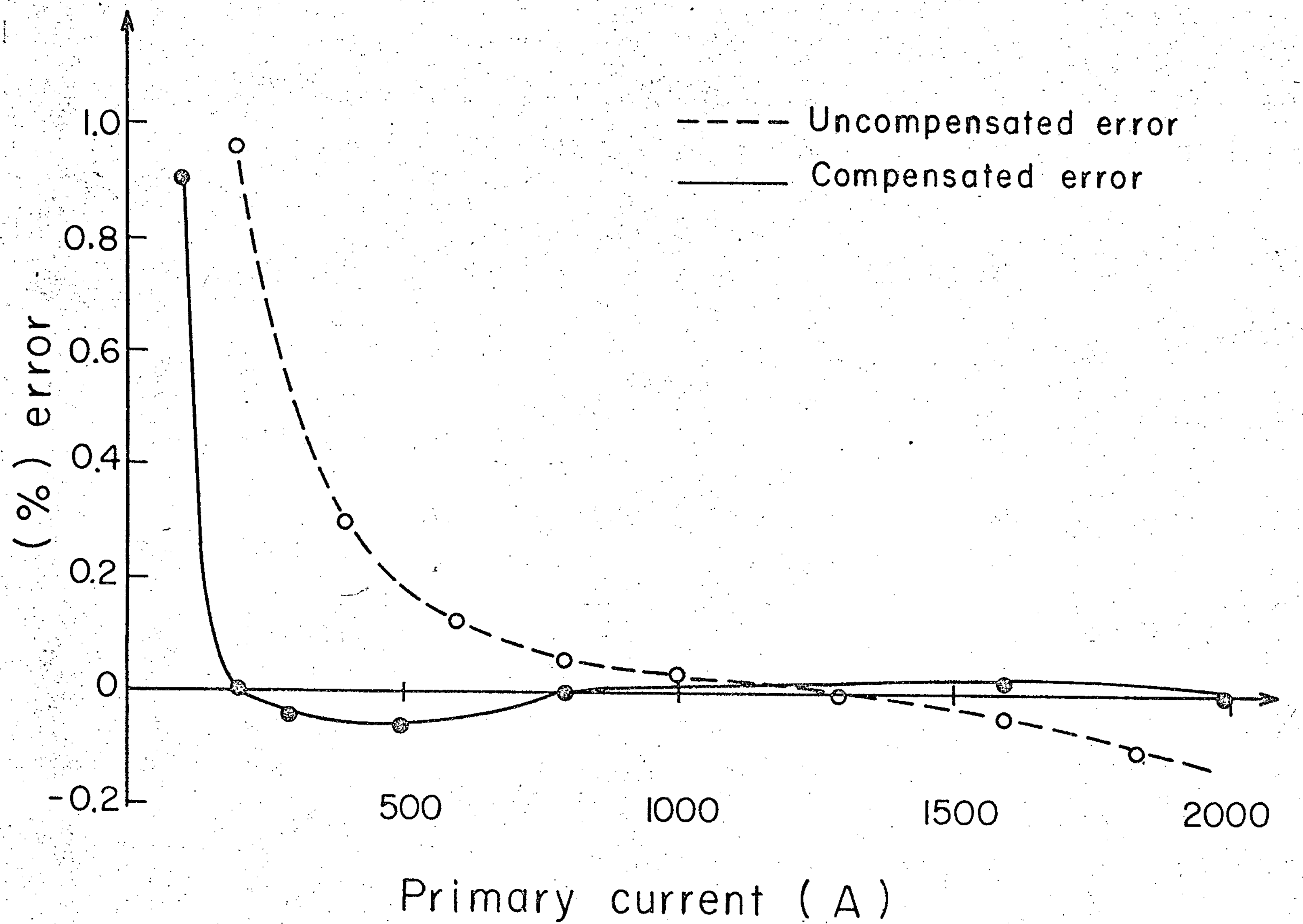

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Fig. 8