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Spacecraft Transformer and Inductor Design

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Colonel W. T. McLyman

August 15, 1977

National Aeronautics and
Space Administration

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PREFACE

The work described in this report was performed by the Control and Energy Conversion Division of the Jet Propulsion Laboratory.

ACKNOWLEDGMENT

The author is grateful to Dr. G. W. Wester, S. Nagano, E. L. Sheldon and Mary Fran Buehler for their assistance and suggestions in preparation of this report.

ABSTRACT

The conversion process in spacecraft power electronics requires the use of magnetic components which frequently are the heaviest and bulkiest items in the conversion circuit. They also have a significant effect upon the performance, weight, cost, and efficiency of the power system.

This handbook contains eight chapters, which pertain to magnetic material selection, transformer and inductor design tradeoffs, transformer design, iron core dc inductor design, toroidal powder core inductor design, window utilization factors, regulation, and temperature rise. Relationships are given which simplify and standardize the design of transformers and the analysis of the circuits in which they are used.

The interactions of the various design parameters are also presented in simplified form so that tradeoffs and optimizations may easily be made.

CONTENTS

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<u>Tables</u>

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LIST OF SYMBOLS

α	regulation, %
A_c	effective iron area, cm^2
A_p	area product, $W_a \times A_c$, cm^4
A_t	surface area of a transformer, cm^2
A_w	wire area, cm^2
$A_{w(B)}$	bare wire area, cm^2
AWG	American Wire Gauge
B_{ac}	alternating current flux density, teslas
B_{dc}	direct current flux density, teslas
B_m	flux density, teslas
B_s	flux density to saturate
cir-mil	area of a circle whose diameter = 0.001 inches
D	lamination tongue width, cm
E	voltage
Eng	energy, watt seconds
η	efficiency
f	frequency, Hz
F	fringing flux factor
G	window height, cm
H	magnetizing force ampturns/cm
H_s	magnetizing force to saturate
I	current, amps
I_o	load current, amps
I_p	primary current, amps

LIST OF SYMBOLS (contd)

I_s	secondary current, amps
J	current density, amps/cm^2
J_p	primary current density, amps/cm^2
J_s	secondary current density, amps/cm^2
K	constant
K_e	electrical coefficient
K_g	geometry coefficient
K_i	gap loss coefficient
K_j	current density coefficient
K_p	area product coefficient
K_s	surface area coefficient
K_u	window utilization factor
K_v	volume coefficient
K_w	weight coefficient
L	inductance, henry
l_g	gap, cm
l_m	magnetic path, cm
ℓ	linear dimension, cm
m	meter
MLT	mean length turn, cm
μ_Δ	effective permeability
μ_m	core material permeability
μ_o	absolute permeability
μ_r	relative permeability

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LIST OF SYMBOLS (contd)

N	turns
P	power, watts
ϕ	flux webers
P_{cu}	copper loss, watts
P_{fe}	core loss, watts
P_{in}	input power, watts
P_o	output power, watts
Ψ	heat flux density, watts/cm^2
P_p	primary loss, watts
P_s	secondary loss, watts
P_Σ	total loss (core and copper), watts
P_t	apparent power, watts
R	resistance, ohms
ρ	resistivity
R_E	equivalent core-loss (shunt) resistance, ohms
R_{cu}	copper resistance, ohms
R_o	load resistance, ohms
R_p	primary resistance, ohms
R_s	secondary resistance, ohms
R_t	total resistance, ohms
S_1	conductor area/wire area
S_2	wound area/usable window
S_3	usable window area/window area

LIST OF SYMBOLS (contd)

S_4	usable window area/usable window area + insulation area
T	flux density, teslas
V_o	load voltage, volts
Vol	volume, cm^3
W_a	window area, cm^2
W_t	weight, grams
ζ	zeta resistance correction factor for temperature

CHAPTER I

MAGNETIC MATERIALS SELECTION FOR STATIC INVERTER AND CONVERTER TRANSFORMERS

A. INTRODUCTION

Transformers used in static inverters, converters and transformer-rectifier (T-R) supplies intended for spacecraft power applications are usually of square loop tape toroidal design. The design of reliable, efficient, and lightweight devices for this use has been seriously hampered by the lack of engineering data describing the behavior of both the commonly used and the more exotic core materials with higher frequency square wave excitation.

A program has been carried out at JPL to develop this data from measurements of the dynamic B-H loop characteristics of the different tape core materials presently available from various industry sources. Cores were procured in both toroidal and "C" forms and were tested in both upgapped (uncut) and gapped (cut) configurations. The following describes the results of this investigation.

B. TYPICAL OPERATION

Transformers used for inverters, converters, and T-R supplies operate from the spacecraft power bus, which could be dc or ac. In some power applications, a commonly used circuit is a driven transistor switch arrangement such as that shown in Fig. 1-1.

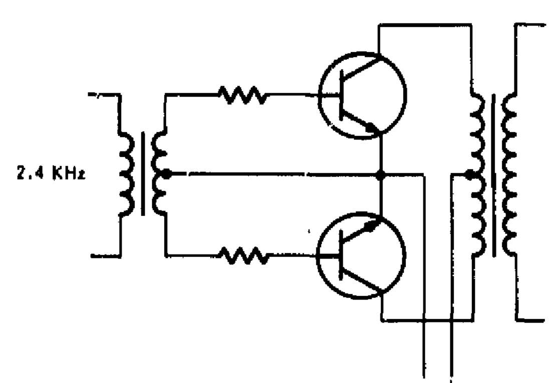

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Fig. 1-1. Typical driven transistor inverter

One important consideration affecting the design of suitable transformers is that care must be taken to ensure that operation involves balanced drive to the transformer primary. In the absence of balanced drive, a net dc current will flow in the transformer primary, which causes the core to saturate easily during alternate half-cycles. A saturated core cannot support the applied voltage, and, because of lowered transformer impedance, the current flowing in a switching transistor is limited mainly by its beta. The resulting high current, in conjunction with the transformer leakage inductance, results in a high voltage spike during the switching sequence that could be destructive to the transistors. To provide balanced drive, it is necessary to exactly match the transistors for V_{CE} (SAT) and beta, and this is not always sufficiently effective. Also, exact matching of the transistors is a major problem in the practical sense.

C. MATERIAL CHARACTERISTICS

Many available core materials approximate the ideal square loop characteristic illustrated by the B-H curve shown in Fig. 1-2.

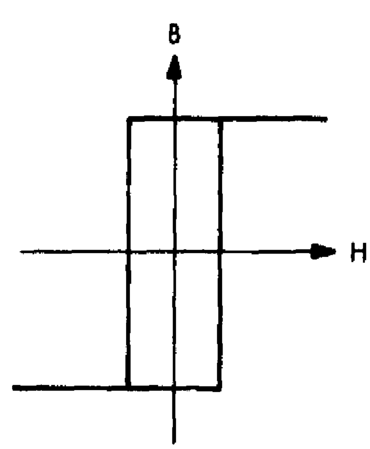

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Fig. 1-2. Ideal square B-H loop

Representative dc B-H loops for commonly available core materials are shown in Fig. 1-3. Other characteristics are tabulated in Table 1-1.

Many articles have been written about inverter and converter transformer design. Usually, the author's recommendation represents a compromise among material characteristics such as those tabulated in

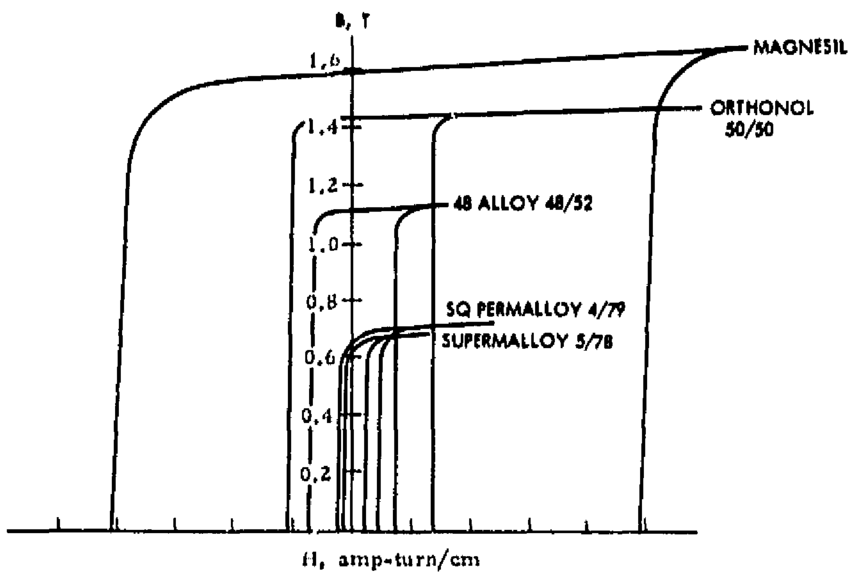


Fig. 1-3. The typical dc B-H loops of magnetic materials

Table 1-1. Magnetic core material characteristics

Trade names	Composition	Saturated flux density,[a] (tesla)	DC coercive force, amp-turn/cm	Squareness ratio	Material density, g/cm³	Loss factor at 3 kHz and 0.5 T, W/kg
Magnesil Silectron Microsil Supersil	3% Si 97% Fe	1.5-1.8	0.5-0.75	0.85-1.0	7.63	33.1
Deltamax Orthonol 49 Sq. Mu	50% Ni 50% Fe	1.4-1.6	0.125-0.25	0.94-1.0	8.24	17.66
Allegheny 4750 48 Alloy Carpenter 49	48% Ni 52% Fe	1.15-1.4	0.062-0.187	0.80-0.92	8.19	11.03
4-79 Permalloy Sq. Permalloy 80 Sq. Mu 79	79% Ni 17% Fe 4% Mo	0.66-0.82	0.025-0.05	0.80-1.0	8.73	5.51
Supermalloy	78% Ni 17% Fe 5% Mo	0.65-0.82	0.0037-0.01	0.40-0.70	8.76	3.75

[1] 1 T = 10^4 Gauss

[2] 1 g/cm³ = 0.036 lb/in.³

Table 1-1 and displayed in Fig. 1-3. These data are typical of commercially available core materials that are suitable for the particular application.

As can be seen, the material that provides the highest flux density (silicon) would result in smallest component size, and this would influence the choice, if size were the most important consideration. The type 78 material (see the 78% curve in Fig. 1-3) has the lowest flux density. This results in the largest size transformer, but, on the other hand, this material has the lowest coercive force and the lowest core loss of any core material available.

Usually, inverter transformer design is aimed at the smallest size, with the highest efficiency, and adequate performance under the widest range of environmental conditions. Unfortunately, the core material that can produce the smallest size has the lowest efficiency. The highest efficiency materials result in the largest size. Thus the transformer designer must make tradeoffs between allowable transformer size and the minimum efficiency that can be tolerated. The choice of core material will then be based upon achieving the best characteristic on the most critical or important design parameter, and acceptable compromises on the other parameters.

Based upon analysis of a number of designs, most engineers select size rather than efficiency as the most important criteria and select an intermediate loss factor core material for their transformers. Consequently, square loop 50-50 nickel-iron has become the most popular material.

D. CORE SATURATION DEFINITION

To standardize the definition of saturation several unique points on the B-H loop are defined as shown in Fig. 1-4.

The straight line through $(H_o, 0)$ and (H_s, B_s) may be written as:

$$B = \left(\frac{dB}{dH}\right)(H - H_o) \tag{1-1}$$

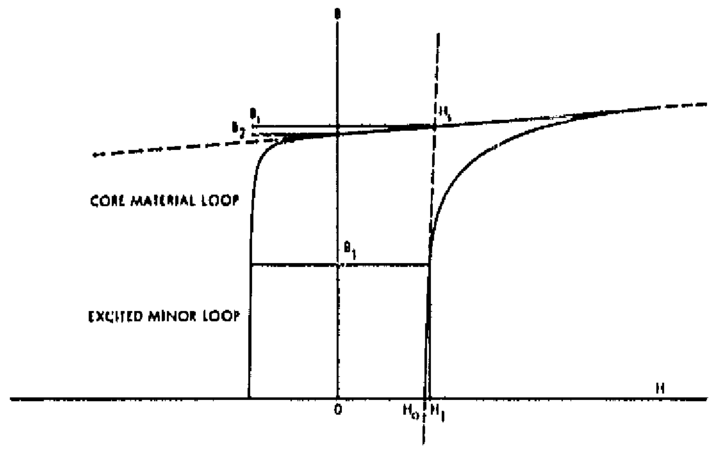


Fig. 1-4. Defining the B-H loop

The line through (0, B_s) and (H_s, B_s) has essentially zero slope and may be written as:

$$B = B_2 \approx B_s \tag{1-2}$$

Equations (1) and (2) together defined "saturation" conditions as follows:

$$B_s = \left(\frac{dB}{dH}\right)(H_s - H_o) \tag{1-3}$$

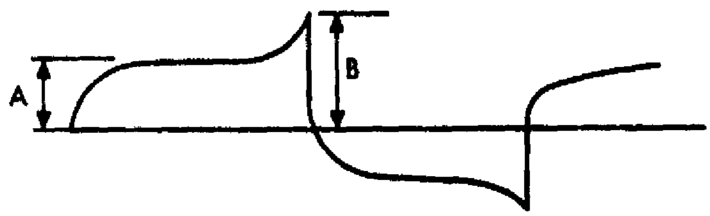
Solving Eq. (1-3) for H_s,

$$H_s = H_o + \frac{B_s}{\mu_o} \tag{1-4}$$

where

$$\mu_o = \frac{dB}{dH}$$

by definition.



SATURATION OCCURS WHEN B = 2A

Fig. 1-5. Excitation current

Saturation occurs by aefinition is when the peak exciting current is twice the average exciting current as shown in Fig. 1-5. Analytically this means that:

$$H_{pk} = 2H_s \qquad (1\text{-}5)$$

Solving Eq. (1-1) for H_1, we obtain

$$H_1 = H_o + \frac{B_1}{\mu_o} \qquad (1\text{-}6)$$

To obtain the presaturation dc margin (ΔH), Eq. (1-4) is subtracted from Eq. (1-3):

$$\Delta H = H_s - H_1 = \frac{B_s - B_1}{\mu_o} \qquad (1\text{-}7)$$

The actual unbalanced dc current must be limited to

$$I_{dc} \leqq \frac{\Delta H l_m}{N} \text{ (amperes)} \qquad (1\text{-}8)$$

where

N = TURNS

l_m = mean magnetic length

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Combining Eqs. (1-7) and (1-8) gives

$$I_{dc} \leqq \frac{(B_s - B_1) l_m}{\mu_o N} \qquad (1-9)$$

As mentioned earlier, in an effort to prevent core saturation, the switching transistors are matched for beta and V_{CE}(SAT) characteristics. The effect of core saturation using an uncut or ungapped core is shown in Fig. 1-6, which illustrates the effect on the B-H loop when traversed with a dc bias. Figure 1-7 shows typical B-H loops of 50-50 nickel-iron excited from an ac source with progressively reduced excitation; the vertical scale is 0.4 T/cm. It can be noted that the minor loop remains at one extreme position within the B-H major loop after reduction of excitation. The unfortunate effect of this random minor loop positioning is that when conduction again begins in the transformer winding after shutdown, the flux swing could begin from the extreme, and not from the normal zero axis. The effect of this is to drive the core into saturation with the production of spikes that can destroy transistors.

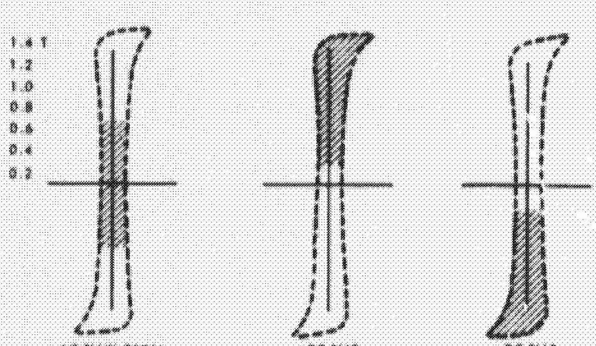

Fig. 1-6. B-H loop with dc bias

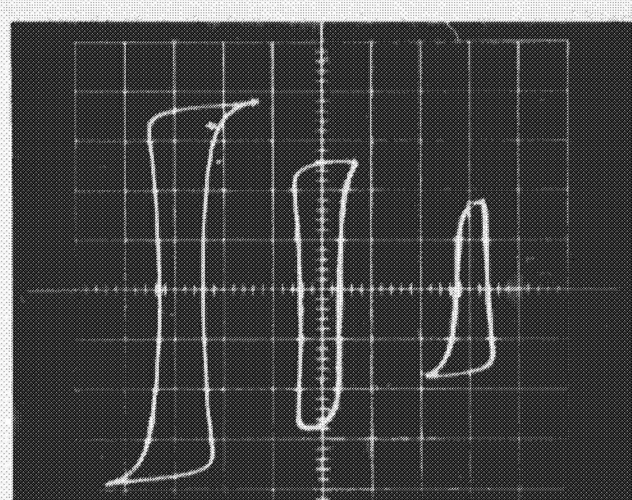

Fig. 1-7. Typical square loop material with ac excitation

E. THE TEST SETUP

A test fixture, schematically indicated in Fig. 1-8, was built to effect comparison of dynamic B-H loop characteristics of various core materials. Cores were fabricated from various core materials in the basic core configuration designated No. 52029 for toroidal cores manufactured by Magnetics, Inc. The materials used were those most likely to be of interest to designers of inverter or converter transformers. Test conditions are listed in Table 1-2.

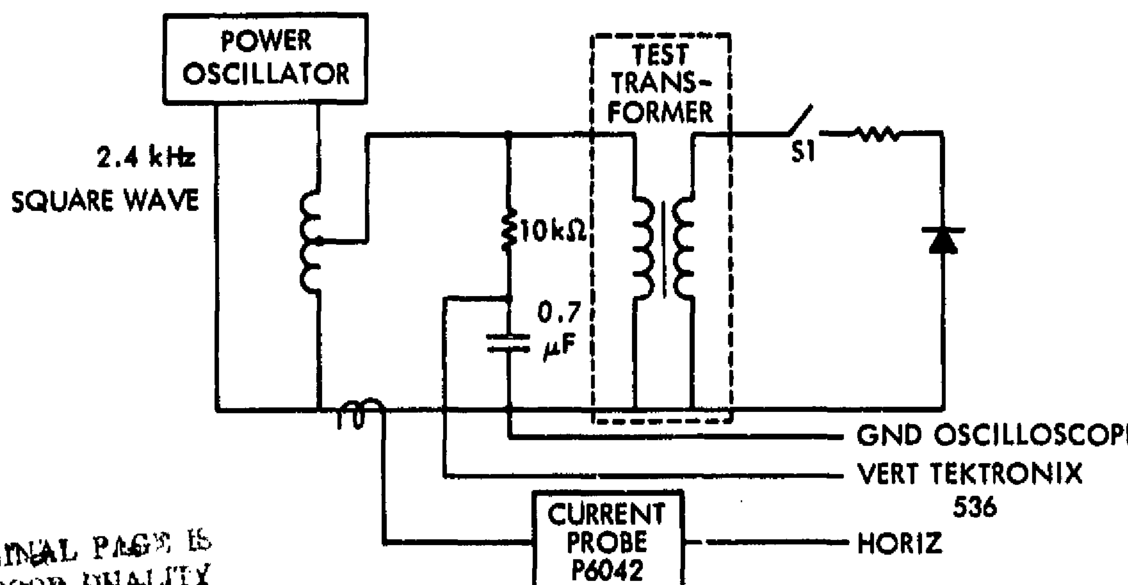

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Fig. 1-8. Dynamic B-H loop test fixture

Table 1-2. Materials and test conditions

Core type	Material	B_m, T	N_T	Frequency, kHz	l_m, cm
52029 (2A)	Orthonol	1.45	54	2.4	9.47
52029 (2D)	Sq. Permalloy	0.75	54	2.4	9.47
52029 (2F)	Supermalloy	0.75	54	2.4	9.47
52029 (2H)	48-Alloy	1.15	54	2.4	9.47
52029 (2H)	Magnesil	1.6	54	2.4	9.47

Winding data was derived from the following:

$$N = \frac{V \cdot 10^4}{4.0 \cdot B_m \cdot f \cdot A_c} \quad (1\text{-}10)$$

The test transformer represented in Fig. 1-9 consists of 54-turn primary and secondary windings, with square wave excitation on the primary. Normally switch S1 is open. With switch S1 closed, the secondary current is rectified by the diode to produce a dc bias in the secondary winding.

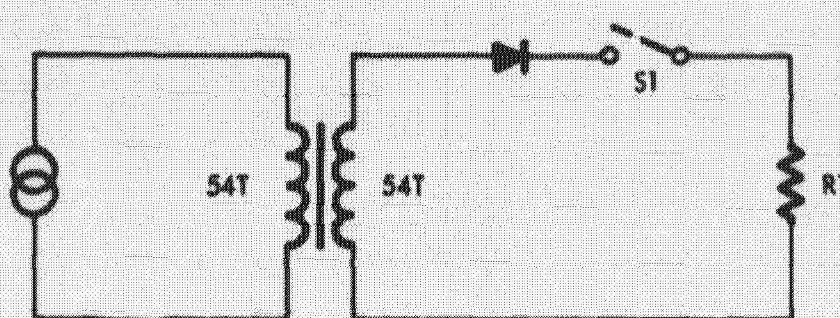
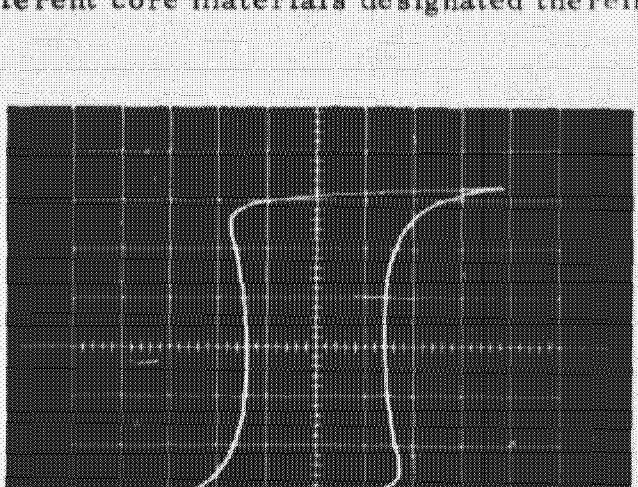


Fig. 1-9. Implementing dc unbalance

Cores were fabricated from each of the materials by winding a ribbon of the same thickness on a mandrel of a given diameter. Ribbon termination was effected by welding in the conventional manner. The cores were vacuum impregnated, baked, and finished as usual.

Figures 1-10 through 1-14 show the dynamic B-H loops obtained for the different core materials designated therein.



VERT = 0.5 T/cm
HORIZ = 100 mA/cm

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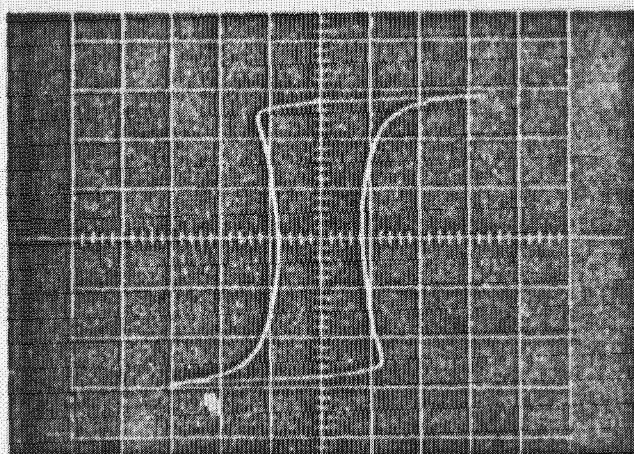
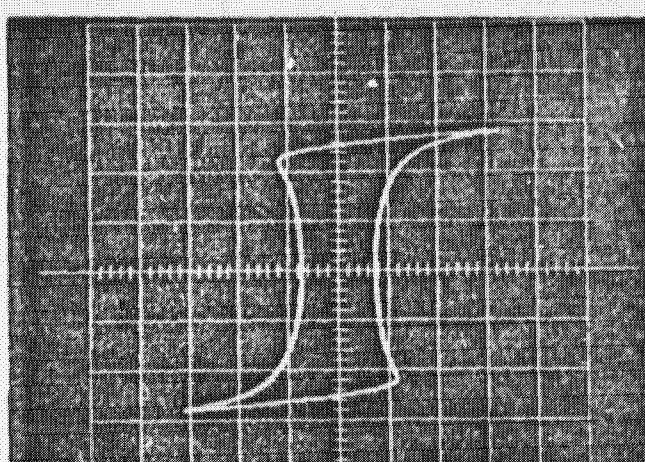
Fig. 1-10. Magnesil (K) B-H loop



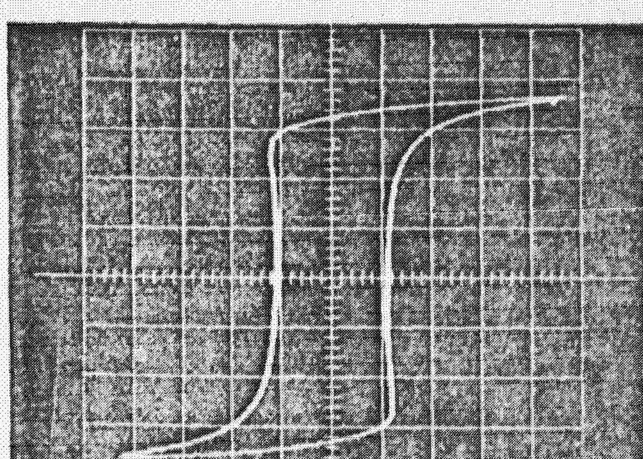
VERT = 0.5 T/cm
HORIZ = 50 mA/cm

Fig. 1-11. Orthonol (A) B-H loop



VERT = 0.5 T/cm
HORIZ = 50 mA/cm

Fig. 1-12. 48 Alloy (H) B-H loop



VERT = 0.2 T/cm
HORIZ = 10 mA/cm

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Fig. 1-13. Sq. Permalloy (P) B-H loop

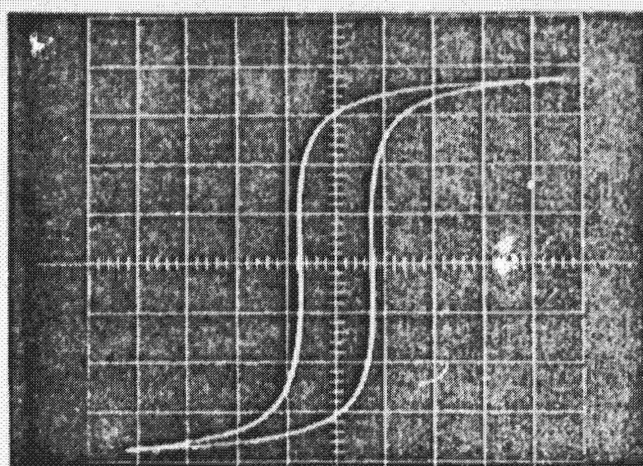

Fig. 1-14. Supermalloy (F) B-H loop

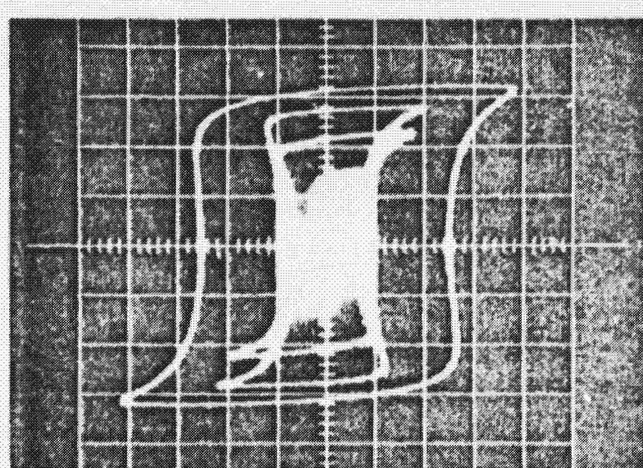

Fig. 1-15. Composite 52029 (2K), (A), (H), (P), and (F) B-H loops

Figure 1-15 shows a composite of all the B-H loops. In each of these, switch S1 was in the open position so that there was no dc bias applied to the core and windings.

The photographs designated Figures 1-16 through 1-20 show the dynamic B-H loop patterns obtained for the designated core materials when the test conditions included a sequence in which switch S1 was open, then closed, and then opened. It is apparent from this data that with a small amount of dc bias, the minor dynamic B-H loop can traverse the major B-H loop from saturation to saturation. In Figs. 1-16 to 1-20, note that after the dc bias had been removed, the minor B-H loops remained shifted to one side or the other. Because of the ac coupling of the integrator to

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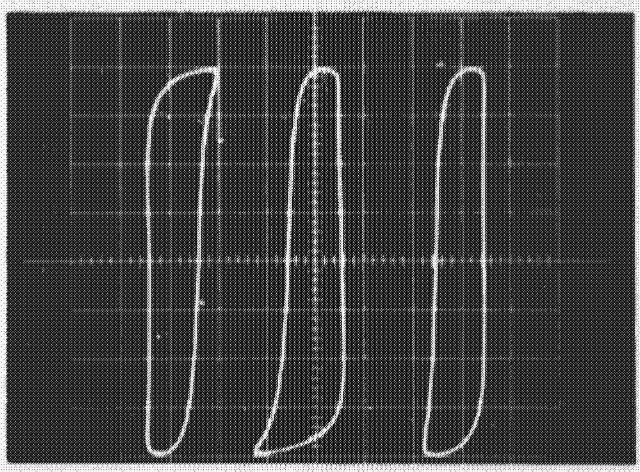
the oscilloscope, the photographs do not present a complete picture of what really happens during the flux swing.



VERT = 0.3 T/cm
HORIZ = 200 mA/cm

Fig. 1-16. Magnesil (K) B-H loop with and without dc bias

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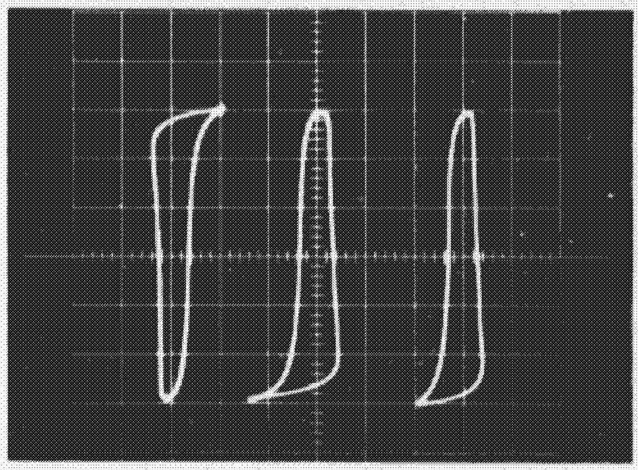

VERT = 0.2 T/cm
HORIZ = 100 mA/cm

Fig. 1-17. Orthonol (A) B-H loop with and without dc bias

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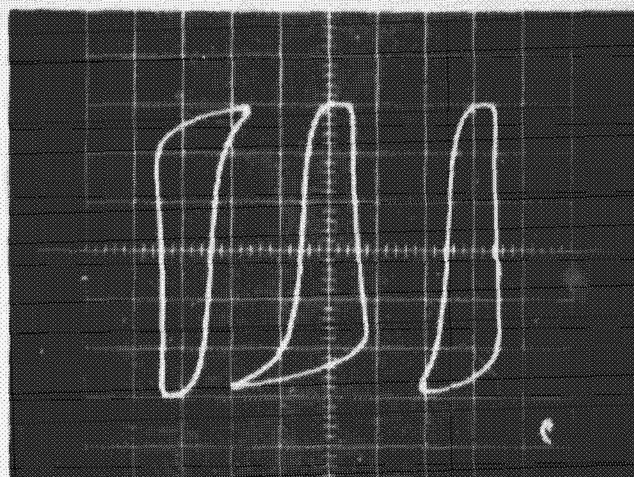
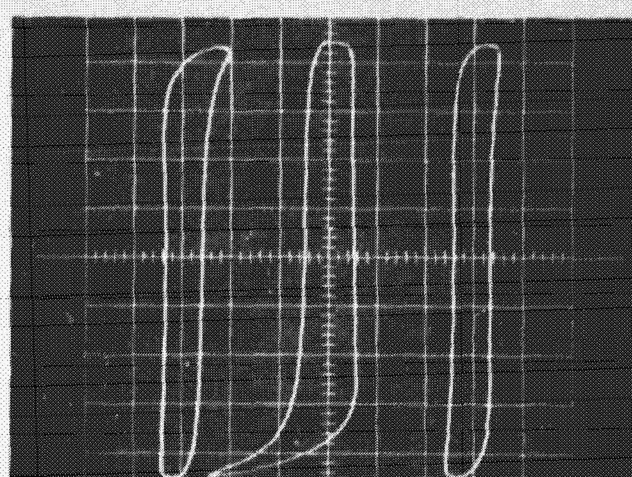


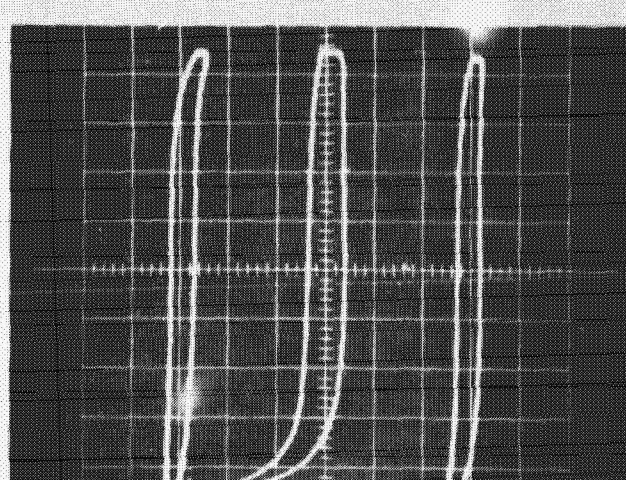
VERT = 0.2 T/cm
HORIZ = 50 mA/cm

Fig. 1-18. 48 Alloy (H) B-H loop with and without dc bias



VERT = 0.1 T/cm
HORIZ = 20 mA/cm

Fig. 1-19. Sq. Permalloy (P) B-H loop with and without dc bias



VERT = 0.1 T/cm
HORIZ = 20 mA/cm

Fig. 1-20. Supermalloy (F) B-H loop with and without dc bias

F. CORE SATURATION THEORY

The domain theory of the nature of magnetism is based on the assumption that all magnetic materials consist of individual molecular magnets. These minute magnets are capable of movement within the material. When a magnetic material is in its unmagnetized state, the individual magnetic particles are arranged at random, and effectively neutralize each other. An example of this is shown in Fig. 1-21, where the tiny magnetic particles are arranged in a disorganized manner. The north poles are represented by the darkened ends of the magnetic particles. When a material is magnetized, the individual particles are aligned or oriented in a definite direction (Fig. 1-22).

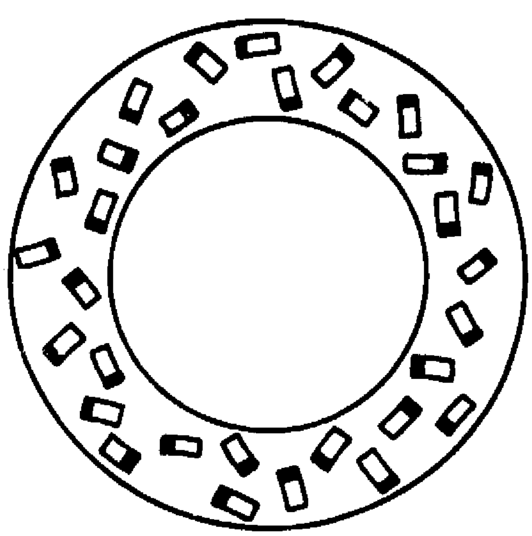


Fig. 1-21. Unmagnetized material

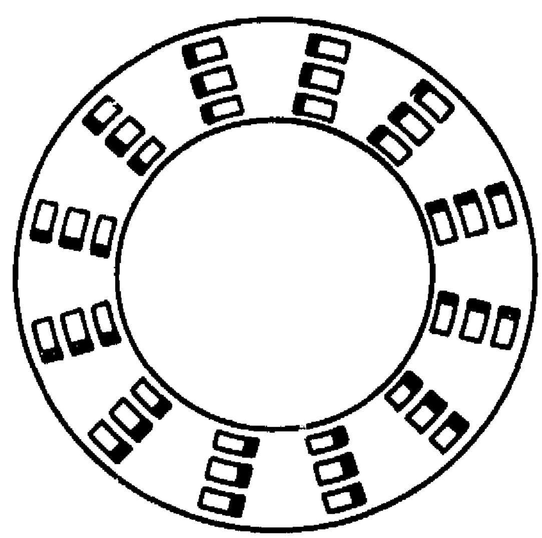


Fig. 1-22. Magnetized material

The degree of magnetization of a material depends on the degree of alignment of the particles. The external magnetizing force can continue up to the point of saturation, that is, the point at which essentially all of the domains are lined up in the same direction.

In a typical toroid core, the effective air gap is less than 10^{-6} cm. Such a gap is negligible in comparison to the ratio of mean length to permeability. If the toroid were subjected to a strong magnetic field (enough to saturate), essentially all of the domains would line up in the same direction.

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If suddenly the field were removed at B_m, the domains would remain lined up and be magnetized along that axis. The amount of flux density that remains is called residual flux or B_r. The result of this effect was shown earlier in Figs. 1-16 to 1-20.

G. AIR GAP

An air gap introduced into the core has a powerful demagnetizing effect, resulting in "shearing over" of the hysteresis loop and a considerable decrease in permeability of high-permeability materials. The dc excitation follows the same pattern. However, the core bias is considerably less affected by the introduction of a small air gap than the magnetization characteristics. The magnitude of the air gap effect also depends on the length of the mean magnetic path and on the characteristics of the uncut core. For the same air gap, the decrease in permeability will be less with a greater magnetic flux path but more pronounced in a low coercive force, high-permeability core.

H. EFFECT OF GAPPING

Figure 1-23 shows a comparison of a typical toroid core B-H loop without and with a gap. The gap increases the effective length of the magnetic path. When voltage E is impressed across primary winding N_1 of a transformer, the resulting current i_m will be small because of the highly inductive circuit shown in Fig. 1-24. For a particular size core, maximum inductance occurs when the air gap is minimum.

When S1 is closed, an unbalanced dc current flows in the N_2 turns and the core is subjected to a dc magnetizing force, resulting in a flux density that may be expressed as

$$B_{dc} = \frac{0.4\pi N I_{dc} \times 10^{-4}}{l_g + \frac{l_m}{\mu_r}} \quad \text{[teslas]} \qquad (1\text{-}11)$$

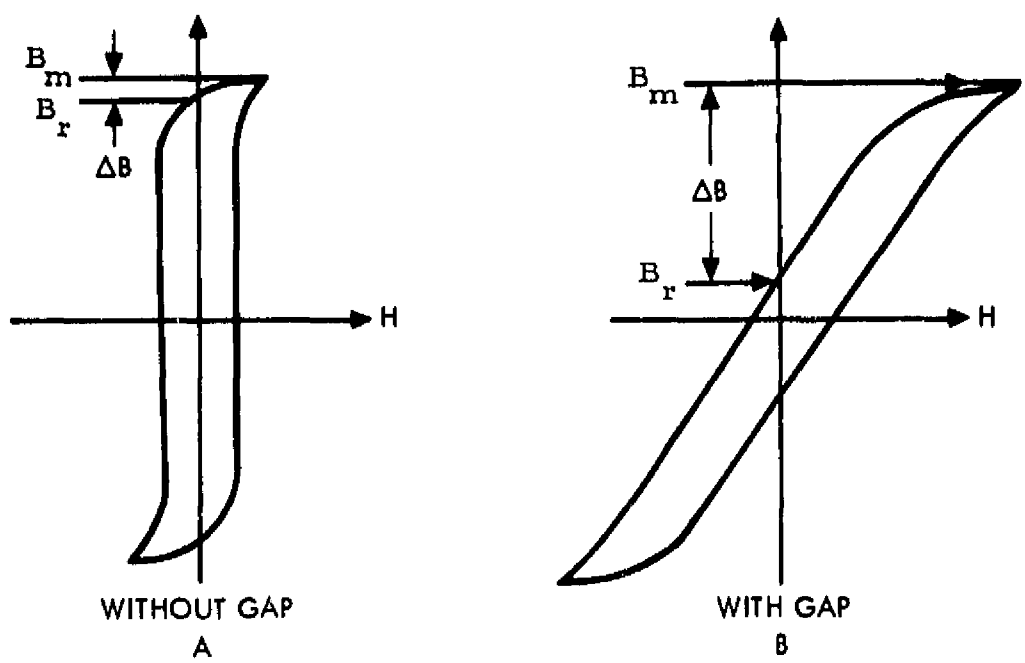

Fig. 1-23. Air gap increases the effective length of the magnetic path

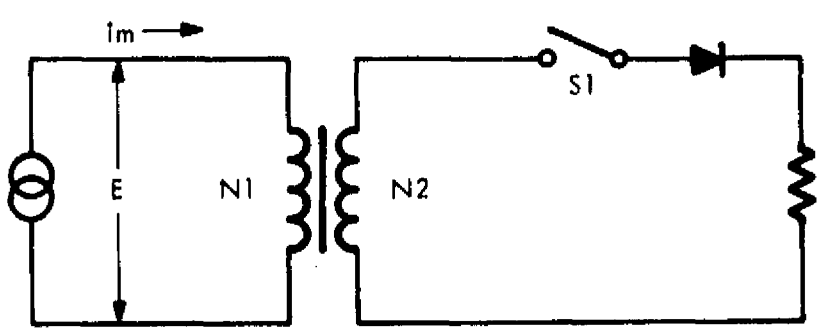

Fig. 1-24. Implementing dc unbalance

In converter and inverter design, this is augmented by the ac flux swing, which is:

$$B_{ac} = \frac{E \cdot 10^4}{K \cdot f \cdot A_c \cdot N} \quad \text{[teslas]} \quad (1\text{-}12)$$

If the sum of B_{dc} and B_{ac} shifts operation above the maximum operating flux density of the core material, the incremental permeability (μac) is reduced. This lowers the impedance and increases the flow of magnetizing current i_m. This can be remedied by introducing an air gap into the core assembly, which effects a decrease in dc magnetization in the core. However, the amount of air gap that can be incorporated has a practical limitation since the air gap lowers impedance, which results in increased magnetizing current (i_m) which is inductive. The resultant voltage spikes produced by such currents apply a high stress to the switching transistors, and may cause failure. This can be minimized by tight control of lapping and etching of the gap to keep the gap to a minimum.

From Fig. 1-23, it can be seen that the B-H curves depict maximum flux density B_m and residual flux B_r for ungapped and gapped cores, and that the useful flux swing is designated ΔB, which is the difference between them. It will be noted in Fig. 1-23a that B_r approaches B_m, but that in Fig. 1-23b there is a much greater ΔB between them. In either case, when excitation voltage is removed at the peak of the excursion of the B-H loop, flux falls to the B_r point. It is apparent that introducing an air gap then reduces B_r to a lower level, and increases the useful flux density. Thus insertion of an air gap in the core eliminates, or reduces markedly, the voltage spikes produced by the leakage inductance due to the transformer saturation.

Two types of core configurations were investigated in the ungapped and gapped states. Figure 1-25 shows the type of toroidal core that was cut

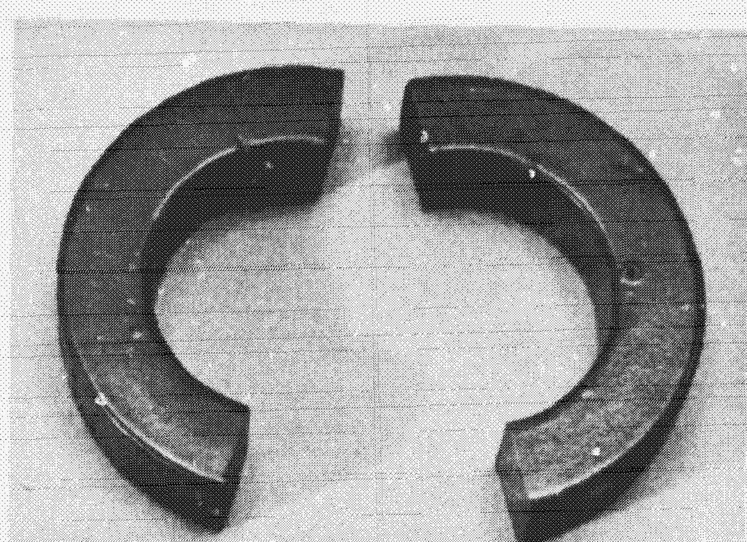

Fig. 1-25. Typical cut toroid

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and Fig. 1-26 shows the type of C core that was cut. Toroidal cores as conventionally fabricated are virtually gapless. To increase the gap, the cores were physically cut in half and the cut edges were lapped, acid etched to remove cut debris, and banded to form the cores. A minimum air gap on the order of less than 25 μm was established.

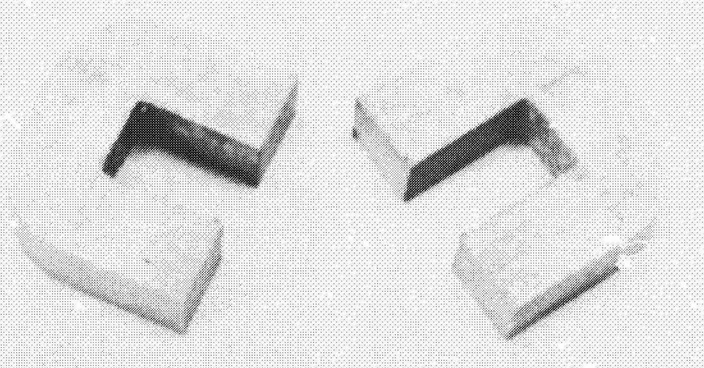

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Fig. 1-26. Typical cut "C" core

As will be noted from Figs. 1-27 to 1-31, which show the B-H loops of the uncut and cut cores, the results obtained indicated that the effect of gapping was the same for both the C-cores and the toroidal cores subjected to testing. It will be noted however, that gapping of the toroidal cores produced a lowered squareness characteristic for the B-H loop as shown in Table 1-3; this data was obtained from Figs. 1-27 to 1-31. Also, from Figs. 1-27 to 1-31, ΔH was extracted as shown in Fig. 1-32 and tabulated in Table 1-4.

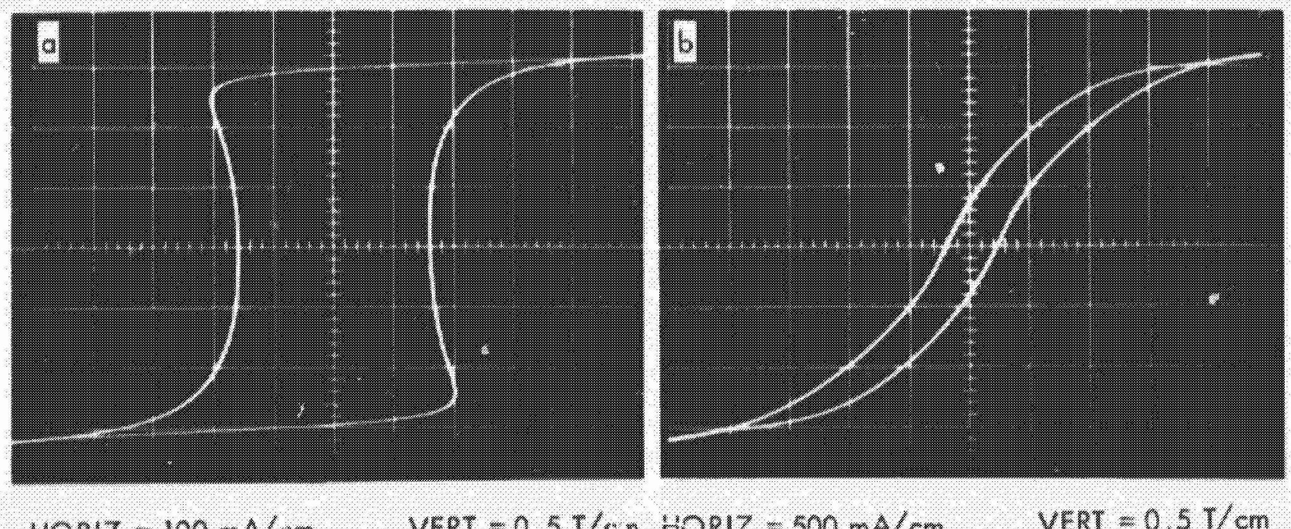

Fig. 1-27. Magnesil 52029 (2K) B-H loop, (a) uncut and (b) cut

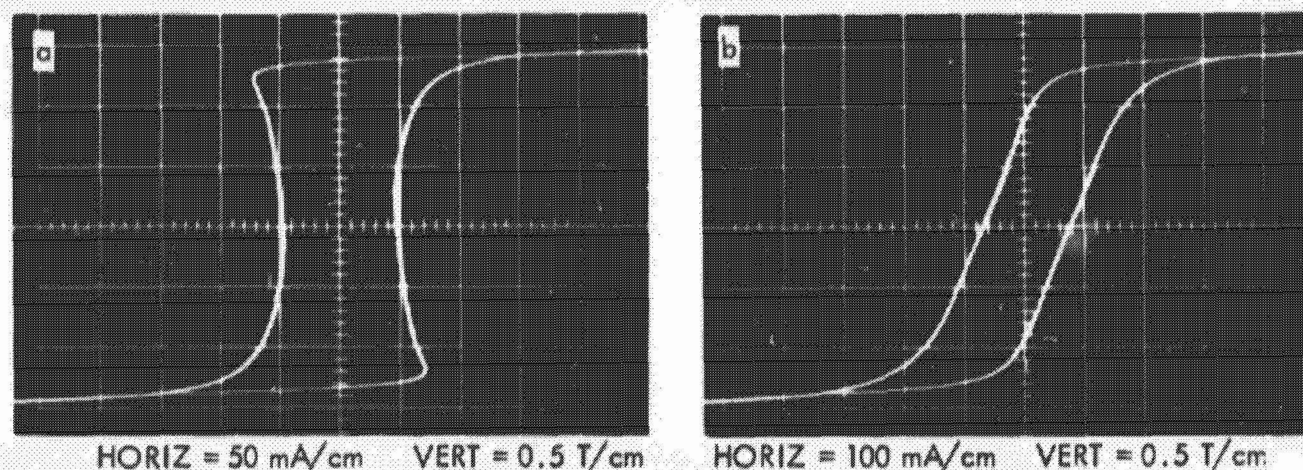

Fig. 1-28. Orthonol 52029 (2A) B-H loop, (a) uncut and (b) cut

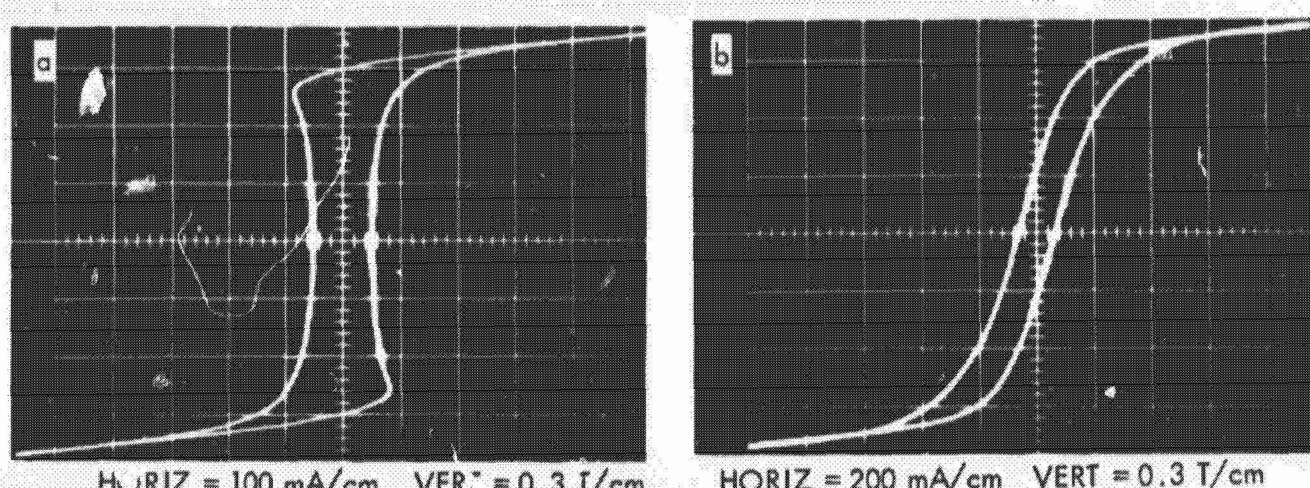

Fig. 1-29. 48 Alloy 52029 (2H) B-H loop, (a) uncut and (b) cut

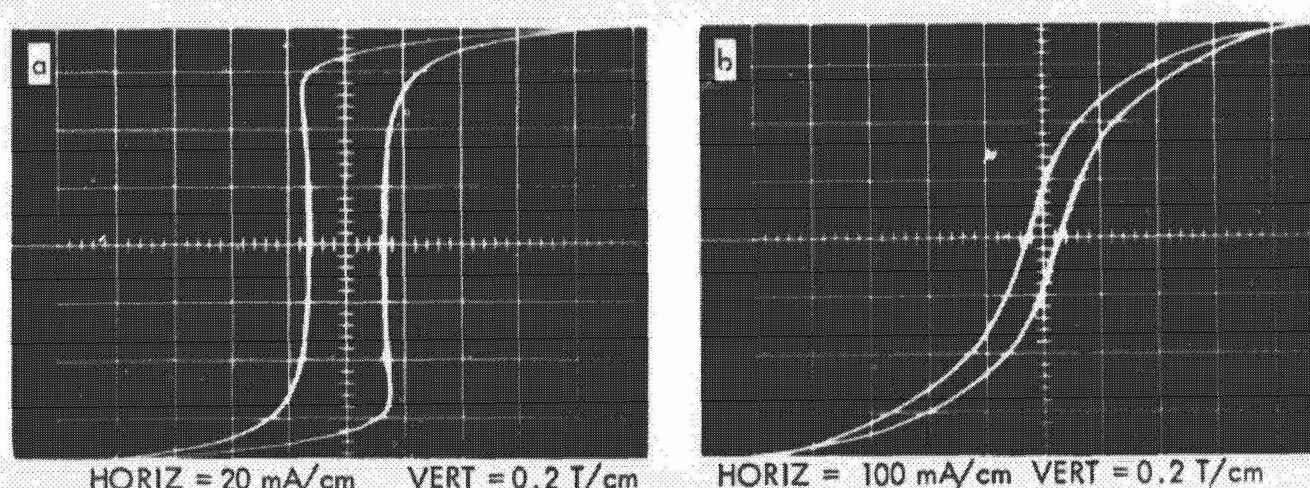

Fig. 1-30. Sq. Permalloy 52029 (2D) B-H loop, (a) uncut and (b) cut

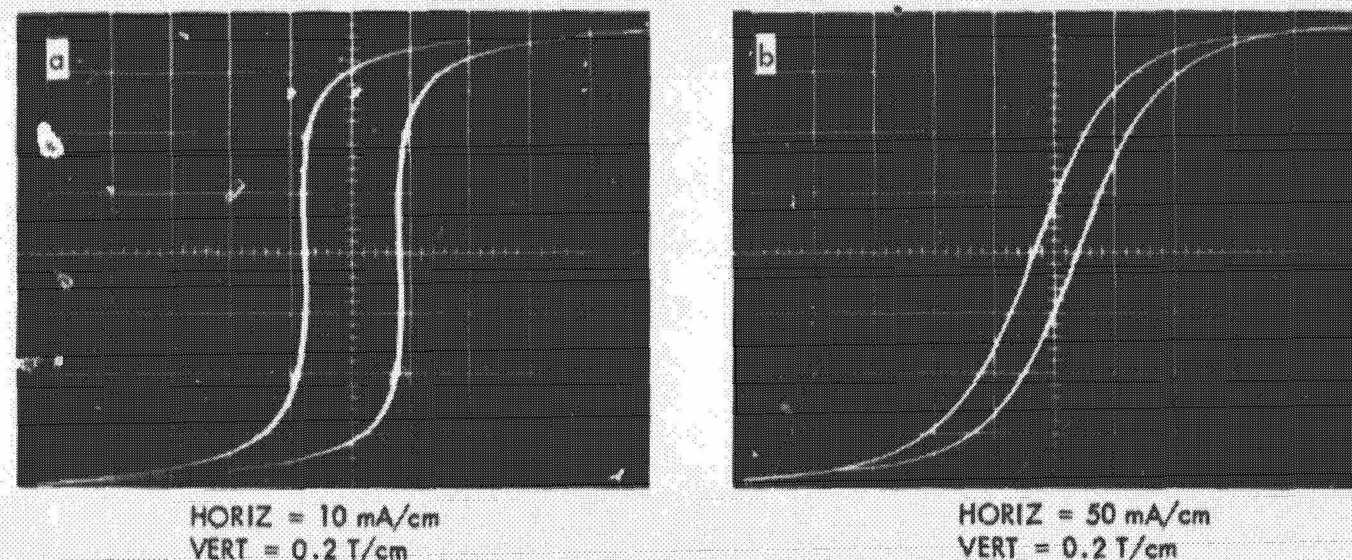

Fig. 1-31. Supermalloy 52029 (2F) B-H loop, (a) uncut and (b) cut

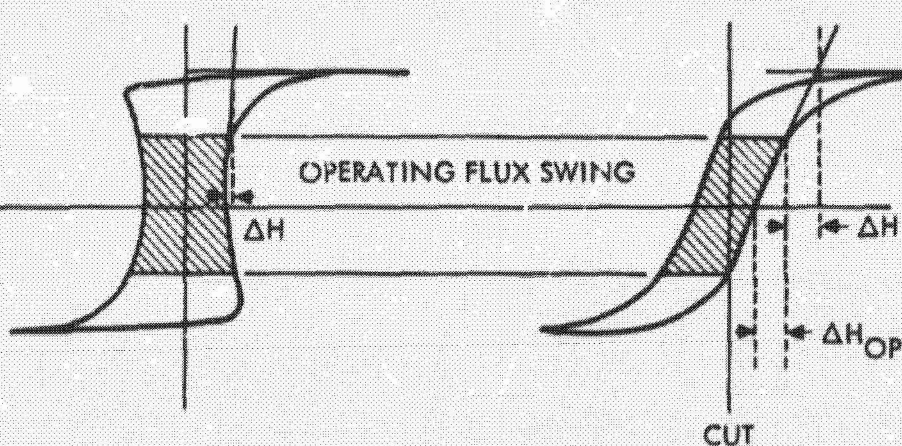

Fig. 1-32. Defining ΔH and ΔH_{OP}

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Table 1-3. Comparing B_r/B_m on uncut and cut cores

Code	Material	Uncut B_r/B_m	Cut B_r/B_m
(A)	Orthonol	0.96	0.62
(D)	Mo-Permalloy	0.86	0.21
(K)	Magnesil	0.93	0.22
(F)	Supermalloy	0.81	0.24
(H)	48 Alloy	0.83	0.30

Table 1-4. Comparing ΔH-ΔH_{OP} on uncut and cut cores

Material	B_m, (tesla)	B_{ac}, (tesla)	B_{dc}, (tesla)	amp-turn/cm			
				Uncut		Cut	
				ΔH_{OP}	ΔH	ΔH_{OP}	ΔH
Orthonal	1.44	1.15	0.288	0.0125	0.0	0.895	0.178
48 Alloy	1.12	0.89	0.224	0.0250	0.0	1.60	0.350
Sq. Permalloy	0.73	0.58	0.146	0.01	0.005	0.983	0.178
Supermalloy	0.68	0.58	0.136	0.0175	0.005	0.491	0.224
Magnesil	1.54	1.23	0.31	0.075	0.025	7.15	1.78

A direct comparison of cut and uncut cores was made electrically by means of two different test circuits. The magnetic material used in this branch of the test was Orthonol. The operating frequency was 2.4 kHz, and the flux density was 0.6 T. The first test circuit, shown in Fig. 1-33, was a driven inverter operating into a 30 W load, with the transistors operating into and out of saturation. Drive was applied continuously. S1 controls the supply voltage to Q1 and Q2.

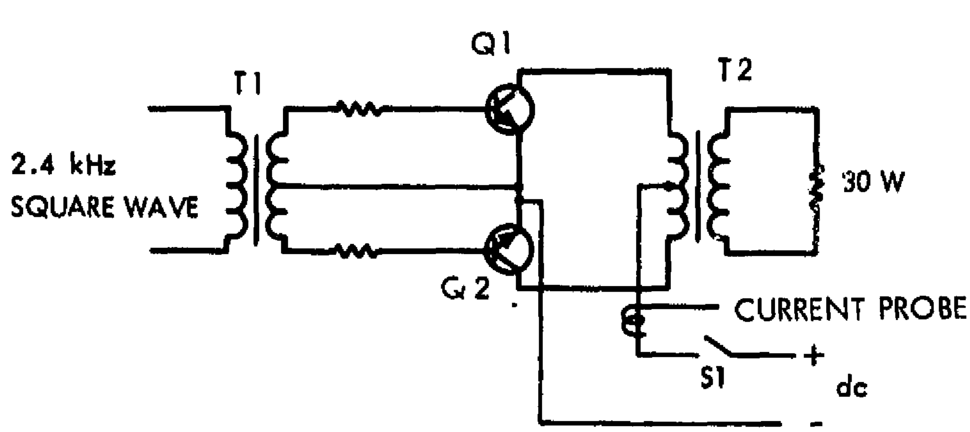

Fig. 1-33. Inverter inrush current measurement

With switch S1 closed, transistor Q1 was turned on and allowed to saturate. This applied $E-V_C(SAT)$ across the transformer winding. Switch S1 was then opened. The flux in transformer T2 then dropped to the residual flux density B_r. Switch S1 was closed again. This was done several times in succession to catch the flux in an additive direction. Figures 1-34 and 1-35 show the inrush current measured at the center tap of T2.

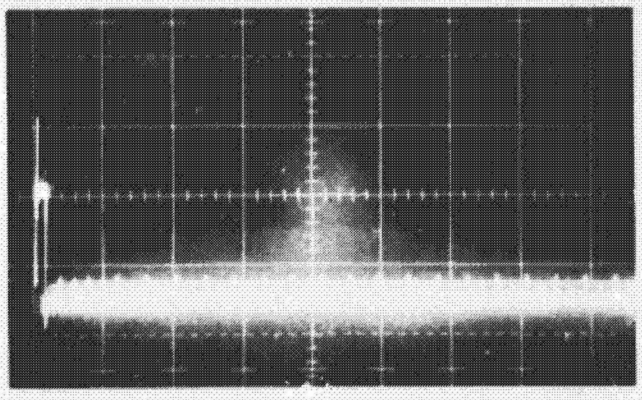

Fig. 1-34. Typical inrush of an uncut core in a driven inverter

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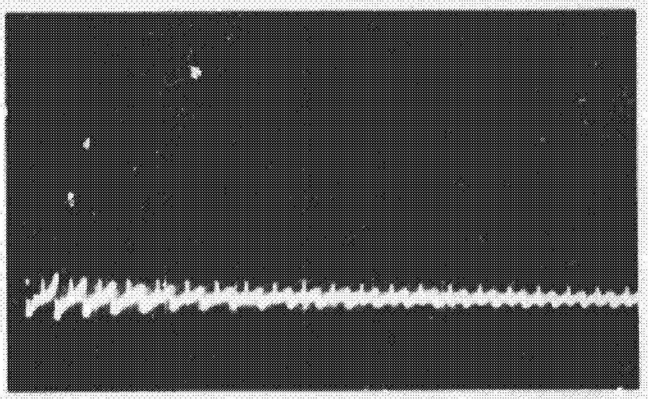

Fig. 1-35. Typical inrush current of a cut core in a driven inverter

It will be noted in Fig. 1-34 that the uncut core saturated and that inrush current was limited only by circuit resistance and transistor beta. It can be noted in Fig. 1-35 that saturation did not occur in the case of the cut core. The high inrush current and transistor stress was thus virtually eliminated.

The second test circuit arrangement is shown in Fig. 1-36. The purpose of this test was to excite a transformer and measure the inrush current using a current probe. A square wave power oscillator was used to excite transformer T2. Switch S1 was opened and closed several times to catch the flux in an additive direction. Figures 1-37 and 1-38 shcw inrush current for a cut and uncut core respectively.

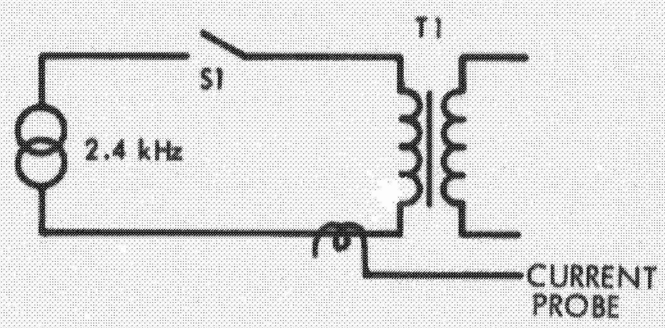

Fig. 1-36. T-R supply current measurement

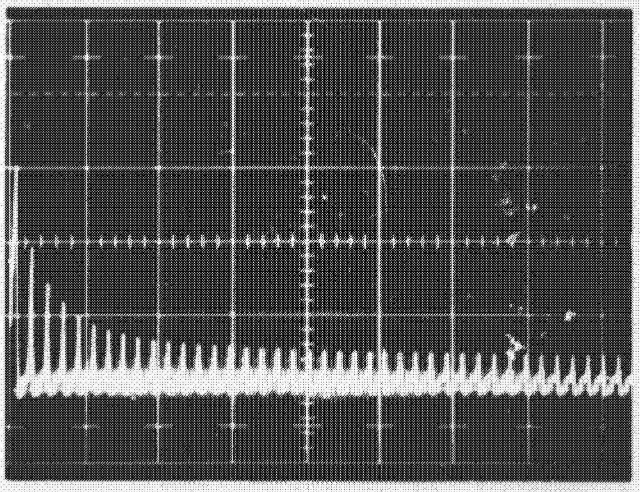

Fig. 1-37. Typical inrush current of an uncut core operating from an ac source

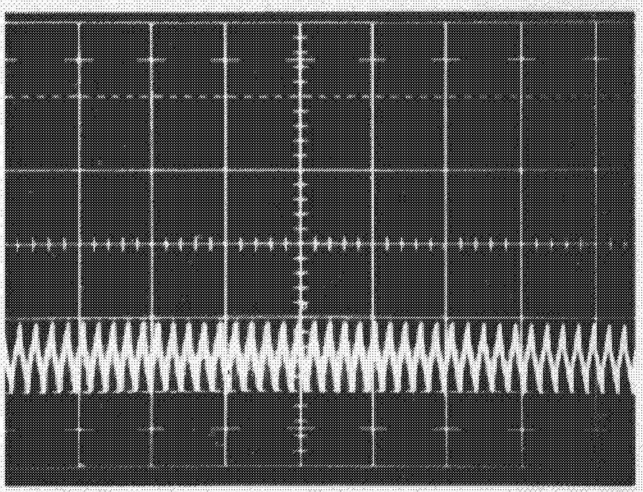

Fig. 1-38. Typical inrush current of a cut core in a T-R

A small amount of air gap, less than 25 μm, has a powerful effect on the demagnetizing force and this gap has little effect on core loss. This small amount of air gap decreases the residual magnetism by "shearing over" the hysteresis loop. This eliminated the problem of the core tending to remain saturated.

A typical example showing the merit of the cut core was in the checkout of a Mariner spacecraft. During the checkout of a prototype science package, a large (8 A, 200 μs) turn-on transient was observed. The normal running current was 0.06 A, and was fused with a parallel-redundant 1/8-A fuse as required by the Mariner Mars 1971 design philosophy. With this 8-A inrush current, the 1/8-A fuses were easily blown. This did not happen on every turn-on, but only when the core would "latch up" in the wrong direction for turn-on. Upon inspection, the transformer turned out to be a 50-50 Ni-Fe toroid. The design was changed from a toroidal core to a cut-core with a 25-μm air gap. The new design was completely successful in eliminating the 8-A turn-on transient.

I. A NEW CORE CONFIGURATION

A new configuration has been developed for transformers which combines the protective feature of a gapped core with the much lower magnetizing current requirement of an uncut core. The uncut core functions under normal operating conditions, and the cut core takes over during abnormal conditions to prevent high switching transients and their potentially destructive effect on the transistors.

This configuration is a composite of cut and uncut cores assembled together in concentric relationship, with the uncut core nested within the cut core. The uncut core has high permeability and thus requires a very small magnetizing current. On the other hand, the cut core has a low permeability and thus requires a much higher magnetization current.

The uncut core is designed to operate at a flux density which is sufficient for normal operation of the converter. The uncut core may saturate under the abnormal conditions previously described. The cut core then takes over and supports the applied voltage so that excessive current does not flow. In a sense it acts like a ballast resistor in some circuits to limit current flow to a safe level.

The photographs designated Figures 1-39 and 1-40 show the magnetization curves for a composite core of the same material, at two different flux densities. The much lower B_r characteristics of the composite as compared to the uncut core is readily apparent.

The desired features of the composite core can be obtained more economically by utilizing different materials for the cut and uncut portions of the core. It was found that when the design required high nickel (4/79) the cut portion could be low nickel (50/50) and because low nickel has twice the flux density as high nickel the core was made 66 percent high nickel and 33 percent low nickel.

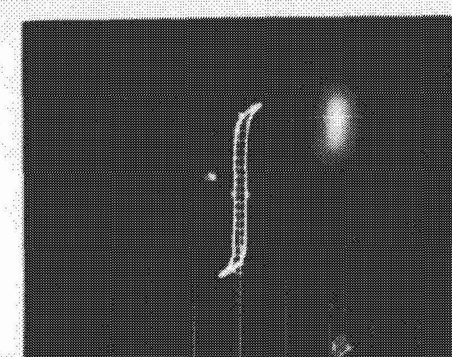

Fig. 1-39. The uncut core excited at 0.2 T/cm

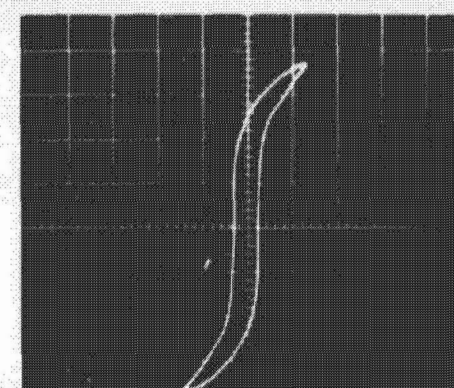

Fig. 1-40. Both cores cut and uncut excited at 0.2 T/cm

The photograph designated Figure 1-41 shows a cut core at the right and an uncut core at the left. Both have been impregnated to bond the ribbon layers together. The photograph designated Figure 1-42 shows in the lower portion, a cut core assembled by banding together with a smaller uncut core. The O. D. of the latter has been trimmed to fit within the I. D. of the cut core by peeling a wrap or two of the ribbon steel. The upper view shows an assembly of the nested cores.

In order to provide uniformity of characteristics for the gapped cores, a gap dimension of 50 μm is recommended so that variations produced by thermal cycling will not affect this gap greatly. This is now obtained by inserting a sheet of paper or film material between the core ends during banding. Then the composite core is placed in the aluminum box and sealed.

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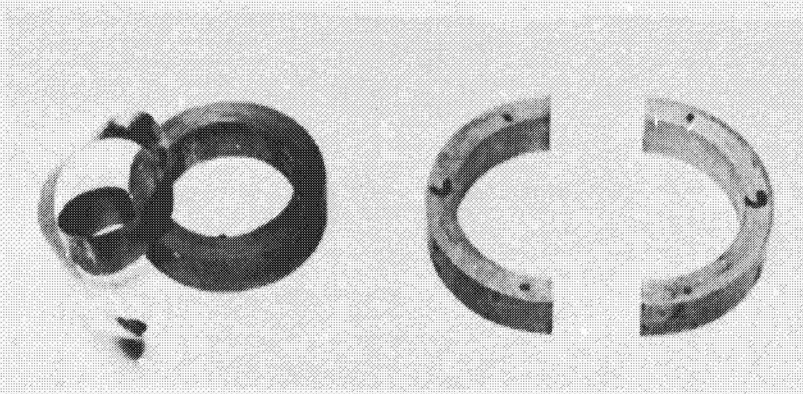


Fig. 1-41. Cores before assembly

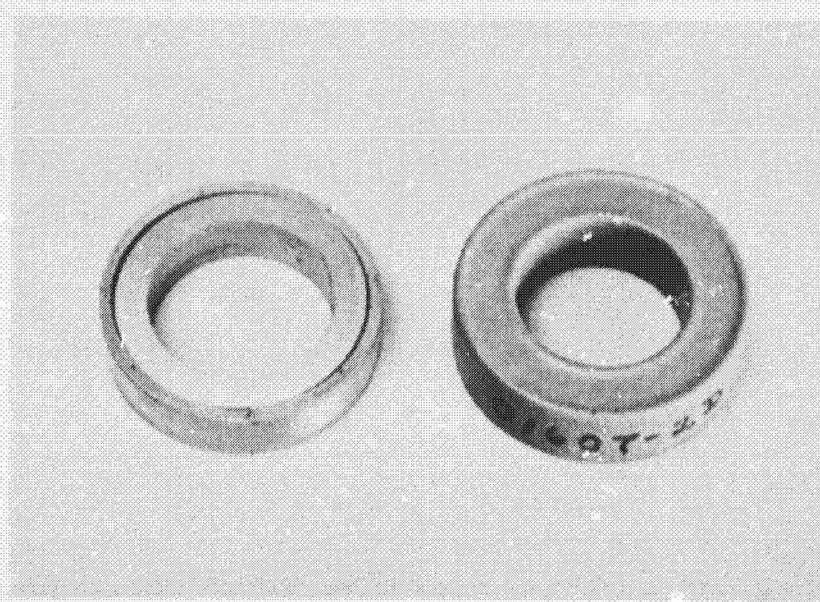


Fig. 1-42. Cores after assembly

This same protective feature can be accomplished in transformers with laminationed cores. When laminations are stacked by interleaving them one by one, the result will be minimum air gap as shown in Figure 1-43 by the squareness of the B-H loop. Shearing over of the B-H loop or decreasing the residual flux is shown in the next Figure 1-44 and is accomplished by butt stacking half of lamination in the core cross section which introduces a small amount of air gap.

Table 1-5 compiles a list of composite cores manufactured by Magnetics Inc., alongside their standard dimensional equivalent cores. Also included in Table 1-5 is the cores' area product A_p, which is described in Chapter 2.

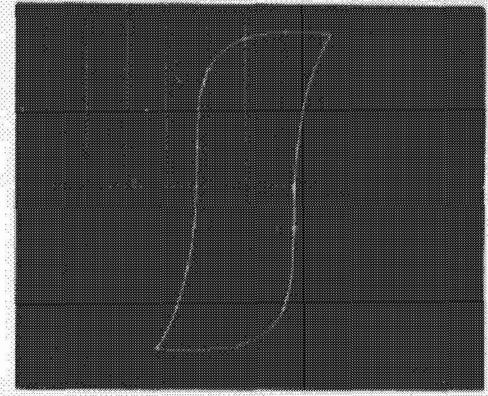
Table 1-5. Composite cores

Composite	Standard	A_p, cm^4
01605-2D	52000	0.0728
01754-2D	52002	0.144
01755-2D	52076	0.285
01609-2D	52061	0.389
01756-2D	52106	0.439
01606-2D	52094	0.603
01757-2D	52029	1.090
01758-2D	52032	1.455
01607-2D	52026	2.180
01759-2D	52038	2.910
01608-2D	52035	4.676
01623-2D	52425	5.255
01624-2D	52169	7.13

A_c = 66% Square Permalloy 4/79.

A_c = 33% Orthonol 50/50.

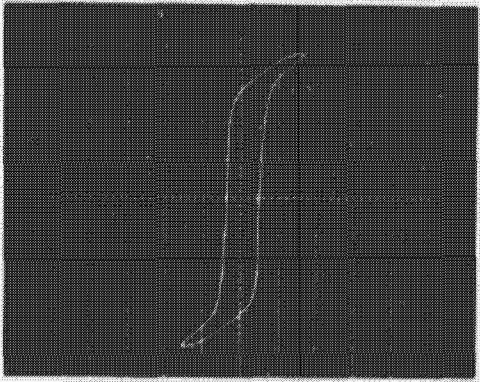
lg = 2 mil Kaption.



0.1 T/cm

10 mA/cm

Fig. 1-43. Stack 1 x 1



0.1 T/cm

50 mA/cm

Fig. 1-44. Stack one half 1 x 1 and one half butt stack

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J. SUMMARY

Low-loss tape-wound toroidal core materials that have a very square hysteresis characteristic (B-H loop) have been used extensively in the design of spacecraft transformers. Due to the squareness of the B-H loops of these materials, transformers designed with them tend to saturate quite easily. As a result, large voltage and current spikes, which cause undue stress on the electronic circuitry, can occur. Saturation occurs when there is any unbalance in the ac drive to the transformer, or when any dc excitation exists. Also, due to the square characteristic, a high residual flux state (B_r) may remain when excitation is removed. Reapplication of excitation in the same direction may cause deep saturation and an extremely large current spike, limited only by source impedance and transformer winding resistance, can result. This can produce catastrophic failure.

By introducing a small (less than 25-μm) air gap into the core, the problems described above can be avoided and, at the same time, the low-loss properties of the materials retained. The air gap has the effect of "shearing over" the B-H loop of the material such that the residual flux state is low and the margin between operating flux density and saturation flux density is high. The air gap thus has a powerful demagnetizing effect upon the square loop materials. Properly designed transformers using "cut" toroid or "C-core" square loop materials will not saturate upon turn-on and can tolerate a certain amount of unbalanced drive or dc excitation.

It should be emphasized, however, that because of the nature of the material and the small size of the gap, extreme care and control must be taken in performing the gapping operation, otherwise the desired shearing effect will not be achieved and the low-loss properties will be lost. The cores must be very carefully cut, lapped, and etched to provide smooth, residue-free surfaces. Reassembly must be performed with equal care.

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CHAPTER II

TRANSFORMER DESIGN TRADEOFFS

A. INTRODUCTION

Manufacturers have for years assigned numeric codes to their cores; these codes represent the power-handling ability. This method assigns to each core a number which is the product of its window area (W_a) and core cross section area (A_c) and is called "Area Product," A_p.

These numbers are used by core suppliers to summarize dimensional and electrical properties in their catalogs. They are available for laminations, C-cores, pot cores, powder cores, and toroidal tape-wound cores.

The author has developed additional relationships between the A_p numbers and current density J for a given regulation and temperature rise. The area product A_p is a dimension to the fourth power ℓ^4, whereas volume is a dimension to the third power ℓ^3 and surface area A_t is a dimension to the second power ℓ^2. Straight-line relationships have been developed for A_p and volume, A_p and surface area A_t, and A_p and weight.

These relationships can now be used as new tools to simplify and standardize the process of transformer design. They make it possible to design transformers of lighter weight and smaller volume or to optimize efficiency without going through a cut and try design procedure. While developed specifically for aerospace applications, the information has wider utility and can be used for the design of non-aerospace transformers as well.

Because of its significance, the area product A_p is treated extensively. A great deal of other information is also presented for the convenience of the designer. Much of the material is in graphical or tabular form to assist the designer in making the tradeoffs best suited for his particular application in a minimum amount of time.

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B. THE AREA PRODUCT A_p AND ITS RELATIONSHIPS

The A_p of a core is the product of the available window area W_a of the core in square centimeters (cm^2) multiplied by the effective cross-sectional area A_c in square centimeters (cm^2) which may be stated as

$$A_p = W_a A_c \qquad [cm^4]$$

Figures 2-1 - 2-5 show in outline form five transformer core types that are typical of those shown in the catalogs of suppliers.

There is a unique relationship between the area product A_p characteristic number for transformer cores and several other important parameters which must be considered in transformer design.

Table 2-1 was developed using the least-squares curve fit from the data obtained in Tables 2-2 through 2-7. The area product A_p relationships with volume, surface area, current density, and weight for pot cores, powder cores, laminations, C-cores, and tape-wound cores will be presented in detail in the following paragraphs.

Table 2-1. Core configuration constants

Core	Losses	K_j (25°C)	K_j (50°C)	(x)	K_s	K_w	K_v
Pot core	$P_{cu}=P_{fe}$	433	632	-0.17	33.8	48.0	14.5
Powder core	$P_{cu} \gg P_{fe}$	403	590	-0.12	32.5	58.8	13.1
Lamination	$P_{cu}=P_{fe}$	366	534	-0.12	41.3	68.2	19.7
C-core	$P_{cu}=P_{fe}$	323	468	-0.14	39.2	66.6	17.9
Single-coil	$P_{cu} \gg P_{fe}$	395	569	-0.14	44.5	76.6	25.6
Tape-wound core	$P_{cu}=P_{fe}$	250	365	-0.13	50.9	82.3	25.0

$J = K_j A_p^{(x)}$ $\qquad A_t = K_s A_p^{0.50}$

$W_t = K_w A_p^{0.75}$ $\qquad Vol = K_v A_p^{0.75}$

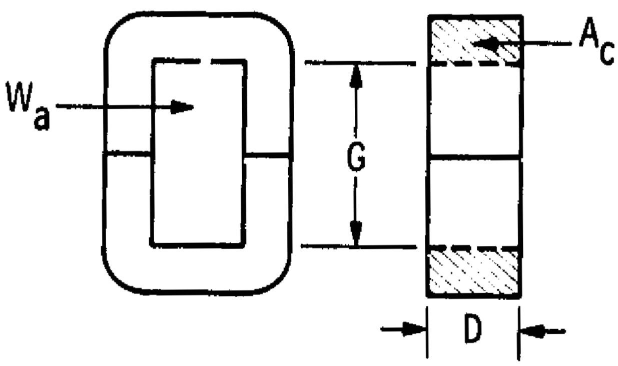

Fig. 2-1. C-core

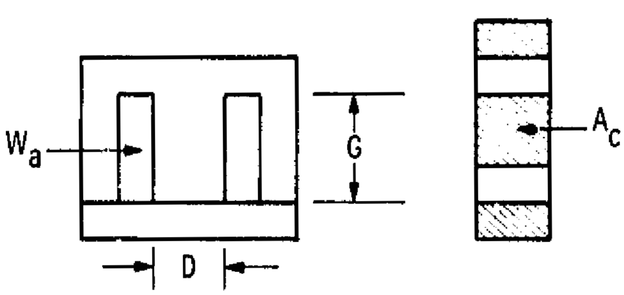

Fig. 2-2. EI lamination

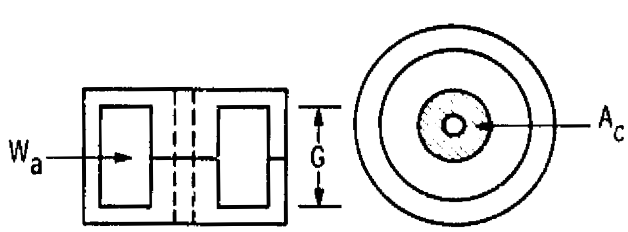

Fig. 2-3. Pot core

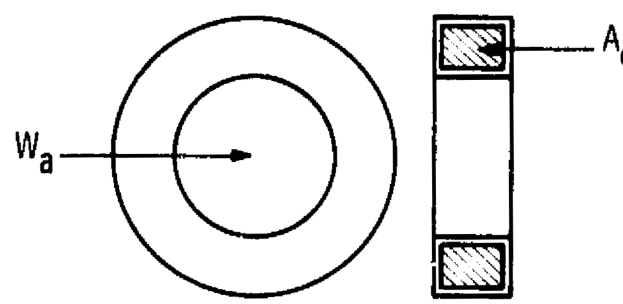

Fig. 2-4. Tape-wound toroidal core

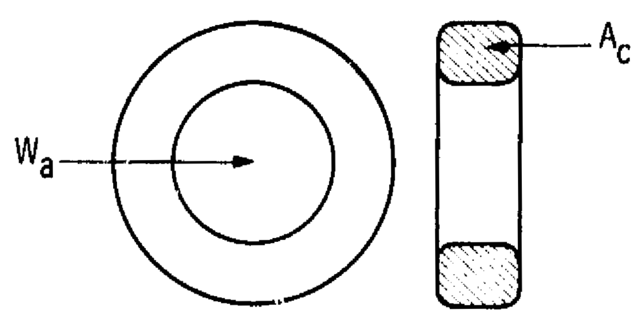

Fig. 2-5. Powder core

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Definitions for Table 2-2

Information given is listed by column as:

1. Manufacturer part number
2. Surface area calculated from Figure 2-22
3. Area product effective iron area times window area
4. Mean length turn
5. Total number of turns and wire size using a window utilization factor $K_u = 0.40$
6. Resistance of the wire at 50°C
7. Watts loss is based on Figure 7-2 for a ΔT of 25°C with a room ambient of 25°C surface dissipation times the transformer surface area, total loss is P_{cu}
8. Current calculated from column 6 and 7
9. Current density calculated from column 5 and 8
10. Resistance of the wire at 75°C
11. Watts loss is based on Figure 7-2 for a ΔT of 50°C with a room ambient of 25°C surface dissipation times the transformer surface area, total loss is P_{cu}
12. Current calculated from column 10 and 11
13. Current density calculated from column 5 and 12
14. Effective core weight for silicon plus copper weight in grams
15. Transformer volume calculated from Figure 2-6
16. C[illegible]ffective cross-section

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Table 2-2. Powder core characteristics

	1	2	3	4	5		6	7	8	9	10	11	12	13	14		15	16
	Core	A_t cm^2	A_p cm^4	MLT cm	N	AWG	Ω @ 50°C	P_Σ	$I = \sqrt{\frac{W}{\Omega}}$	ΔT 25°C $J = I/cm^2$	Ω @ 75°C	P_Σ	$I = \sqrt{\frac{W}{\Omega}}$	ΔT 50°C $J = I/cm^2$	Weight Fe	Cu	Volume cm^3	A_c cm^2
1	55051	7.19	0.0437	2.12	86	25	0.215	0.216	1.00	617	0.236	0.503	1.46	899	3.1	2.71	1.39	0.113
2	55121	12.3	0.137	2.71	160	25	0.513	0.369	0.848	522	0.563	0.861	1.23	762	6.8	6.3	3.11	0.196
3	55848	17.3	0.259	2.95	257	25	0.897	0.519	0.761	469	0.985	1.211	1.11	683	10	11.3	5.07	0.232
4	55059	21.9	0.466	3.29	336	25	1.27	0.657	0.719	443	1.39	1.533	1.05	647	16	16.3	7.28	0.327
5	55894	30.	1.021	4.51	351	25	1.87	0.924	0.703	433	2.06	2.16	1.02	631	36	23.2	12.4	0.639
6	55586	48.6	1.821	4.39	902	25	4.69	1.46	0.558	344	5.15	3.40	0.812	500	35	59.9	23.3	0.458
7	55071	44.7	1.966	4.77	656	25	3.70	1.34	0.602	371	4.07	3.13	0.877	540	47	47.4	21.0	0.666
8	55076	51.6	2.46	4.88	815	25	4.71	1.55	0.574	353	5.17	3.61	0.814	518	52	61.0	25.7	0.670
9	55083	66.8	4.57	6.02	95[illegible]	25	6.84	2.00	0.541	333	7.50	4.68	0.790	487	92	86.0	39.1	1.06
10	55090	89.4	8.1[illegible]	6.65	1372	25	10.8	2.68	0.498	307	11.8	6.26	0.728	449	131	140	59.5	1.32
11	55439	86.9	8.48	7.48	959	25	8.49	2.60	0.553	341	9.32	6.08	0.807	497	182	109	58.1	1.95
12	55716	100.0	9.38	6.54	1684	25	13.0	3.00	0.480	296	14.3	7.00	0.699	431	133	170	69.0	1.24
13	55110	124.0	13.66	7.09	2125	25	17.8	3.72	0.457	282	19.6	8.68	0.665	410	176	226	93.4	1.44

copper loss ≫ iron loss

Definitions for Table 2-3

Information given is listed by column as:

1. Manufacturer part number
2. Surface area calculated from Figure 2-22
3. Area product effective iron area times window area
4. Mean length turn
5. Total number of turns and wire size using a window utilization factor $K_u = 0.40$
6. Resistance of the wire at 50°C
7. Watts loss is based on Figure 7-2 for a ΔT of 25°C with a room ambient of 25°C surface dissipation times the transformer surface area, total loss is equal to 2 P_{cu}
8. Current calculated from column 6 and 7
9. Current density calculated from column 5 and 8
10. Resistance of the wire at 75°C
11. Watts loss is based on Figure 7-2 for a ΔT of 50°C with a room ambient of 25°C surface dissipation times the transformer surface area, total loss is equal to 2 P_{cu}
12. Current calculated from column 10 and 11
13. Current density calculated from column 5 and 12
14. Effective core weight for silicon plus copper weight in grams
15. Transformer volume calculated from Figure 2-6
16. Core effective cross-section

Table 2-3. Pot core characteristics

	1	2	3	4	5		6	7	8	9	10	11	12	13	14		15	16
	Core	A_t cm^2	A_p cm^4	MLT cm	N	AWG	Ω @ 50°C	P_Σ	$I = \sqrt{\frac{W}{\Omega}}$	ΔT 25°C $J = I/cm^2$	Ω @ 75°C	P_Σ	$I = \sqrt{\frac{W}{\Omega}}$	ΔT 50°C $J = I/cm^2$	Weight f_e	Weight Cu	Volume cm^3	A_c cm^2
1	9 x 5	2.93	0.0065	1.85	25	30	0.175	0.098	0.529	1044	0.192	0.230	0.774	1527	0.8	0.32	0.367	0.10
2	11 x 7	4.35	0.0152	2.2	37	30	0.309	0.130	0.458	904	0.339	0.304	0.670	1322	1.7	0.38	0.662	0.16
3	14 x 8	6.96	0.0393	2.8	74	30	0.787	0.208	0.363	716	0.864	0.487	0.531	1048	3.2	0.98	1.35	0.25
4	18 x 11	11.3	0.114	3.56	143	30	1.934	0.339	0.296	584	2.12	0.791	0.432	853	6.0	2.37	2.78	0.43
5	22 x 13	17.0	0.246	4.4	207	30	3.46	0.510	0.271	535	3.80	1.190	0.396	782	13	4.30	5.17	0.63
6	26 x 16	23.9	0.498	5.2	96	25	0.592	0.717	0.778	479	0.650	1.67	1.13	696	21	7.5	8.65	0.94
7	30 x 19	32.8	1.016	6.0	144	25	1.024	0.984	0.693	427	1.12	2.30	1.01	622	36	12.9	13.9	1.36
8	36 x 22	44.8	2.01	7.3	189	25	1.636	1.34	0.639	394	1.79	3.14	0.937	577	57	20.8	22.0	2.01
9	47 x 28	76.0	5.62	9.3	345	25	3.81	2.28	0.547	337	4.18	5.32	0.798	492	125	48.0	48.6	3.12
10	59 x 36	122.0	13.4	12.0	608	25	8.65	3.66	0.459	283	9.50	8.54	0.670	413	270	109	98.3	4.85

copper loss = iron loss

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Definitions for Table 2-4

Information given is listed by column as:

1. Manufacturer part number
2. Surface area calculated from Figure 2-23
3. Area product effective iron area times window area
4. Mean length turn on one bobbin
5. Total number of turns and wire size for one bobbin using a window utilization factor $K_u = 0.40$
6. Resistance of the wire at 50°C
7. Watts loss is based on Figure 7-2 for a ΔT of 25°C with a room ambient of 25°C surface dissipation times the transformer surface area, total loss is equal to 2 P_{cu}
8. Current calculated from column 6 and 7
9. Current density calculated from column 5 and 8
10. Resistance of the wire at 75°C
11. Watts loss is based on Figure 7-2 for a ΔT of 50°C with a room ambient of 25°C surface dissipation times the transformer surface area, total loss is equal to 2 P_{cu}
12. Current calculated from column 10 and 11
13. Current density calculated from column 5 and 12
14. Effective core weight for silicon plus copper weight in grams
15. Transformer volume calculated from Figure 2-7
16. Core effective cross-section (thickness, 0.014) square stack

Table 2-4. Lamination characteristics

	1	2	3	4	5		6	7	8	9	10	11	12	13	14		15	16
	Core	A_t cm^2	A_p cm^4	MLT cm	N	AWG	Ω @ 50°C	P_Σ	$I = \sqrt{\frac{W}{\Omega}}$	ΔT 25°C $J = I/cm^2$	Ω @ 75°C	P_Σ	$I = \sqrt{\frac{W}{\Omega}}$	ΔT 50°C $J = I/cm^2$	Weight Fe	Cu	Volume cm^3	A_c cm^2
1	EE-3031	4.11	0.0088	1.72	90	30	0.58	0.123	0.323	638	0.645	0.288	0.472	932	1.02	1.02	0.651	0.0502
2	EE-2829	6.63	0.0228	2.33	147	30	1.30	0.199	0.276	546	1.43	0.464	0.403	795	2.19	1.59	1.35	0.0907
3	E1-187	14.4	0.108	3.20	314	30	3.82	0.432	0.237	469	4.19	1.01	0.347	685	7.09	3.08	4.34	0.204
4	EE-2425	23.8	0.293	5.08	498	30	9.61	0.714	0.192	380	10.5	1.67	0.281	555	15.5	9.06	9.22	0.363
5	EE-2627	40.6	0.906	5.79	245	25	1.68	1.22	0.602	371	1.85	2.84	0.876	540	45.8	15.5	19.1	0.816
6	E1-375	47.7	1.23	6.30	350	25	2.62	1.43	0.522	322	2.87	3.34	0.762	470	49.7	24.7	25.3	0.816
7	E1-50	57.7	1.75	7.09	263	25	2.21	1.73	0.625	385	2.43	4.04	0.912	562	90.6	31.7	36.8	1.45
8	E1-21	66.0	2.36	7.57	372	25	3.34	1.98	0.544	335	3.66	4.62	0.793	489	99.3	41.0	39.2	1.45
9	E1-625	90.0	4.29	8.84	503	25	5.27	2.70	0.505	312	5.79	6.30	0.737	455	179	44.4	60.0	2.27
10	E1-75	130.0	8.89	10.6	211	20	0.826	3.90	1.54	296	0.906	9.10	2.24	432	312	105	104.0	3.27
11	E1-87	176.0	16.5	12.3	296	20	1.34	5.28	1.40	270	1.48	12.3	2.04	393	481	135	164.0	4.45
12	E1-100	230.0	28.1	14.5	386	20	2.07	6.90	1.29	249	2.27	16.1	1.88	363	712	241	246.0	5.81
13	E1-112	292.0	44.9	16.0	492	20	2.91	8.76	1.23	237	3.19	20.4	1.79	344	1029	342	350.0	7.34
14	E1-125	361.0	68.7	17.7	625	20	4.09	10.8	1.15	222	4.49	25.3	1.68	324	1414	460	481.0	9.07
15	E1-138	432.0	107.0	19.5	740	20	5.33	13.0	1.10	213	5.85	30.2	1.61	310	1880	680	629.0	11.6
16	E1-150	518.0	143.0	21.2	893	20	6.99	15.5	1.05	203	7.67	36.3	1.54	296	2457	906	829.0	13.1
17	E1-175	704.0	263.0	24.7	1080	20	9.85	21.1	1.034	199	10.8	49.3	1.51	291	3575	2355	1312.0	17.8
18	E1-36	778.0	324.0	26.5	1701	20	16.6	23.3	0.836	161	18.3	54.5	1.22	235	3906	1273	1654.0	15.3
19	E1-19	1093.0	601.0	31.7	2886	20	33.8	32.8	0.696	134	37.1	76.5	1.015	196	4889	3805	2875.0	17.8

copper loss = iron loss

Definitions for Table 2-5

Information given is listed by column as:

1. Manufacturer part number
2. Surface area calculated from Figure 2-24
3. Area product effective iron area times window area
4. Mean length turn on one bobbin
5. Total number of turns and wire size for two bobbins using a window utilization factor $K_u = 0.40$
6. Resistance of the wire at 50°C
7. Watts loss is based on Figure 7-2 for a ΔT of 25°C with a room ambient of 25°C surface dissipation times the transformer surface area, total loss is equal to 2 P_{cu}
8. Current calculated from column 6 and 7
9. Current density calculated from column 5 and 8
10. Resistance of the wire at 75°C
11. Watts loss is based on Figure 7-2 for a ΔT of 50°C with a room ambient of 25°C surface dissipation times the transformer surface area, total loss is equal to 2 P_{cu}
12. Current calculated from column 10 and 11
13. Current density calculated from column 5 and 12
14. Effective core weight for silicon plus copper weight in grams
15. Transformer volume calculated from Figure 2-8
16. Core effective cross-section

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Table 2-5. C-core characteristics

	1	2	3	4	5	6	7	8	9	10	11	12	13	14		15	16
	Core	A_t cm^2	A_p cm^4	MLT cm	N / AWG	Ω @ 50° C	P_Σ	$I = \sqrt{\frac{W}{\Omega}}$	ΔT 25°C $J = \frac{amps}{cm^2}$	Ω @ 75° C	P_Σ	$I = \sqrt{\frac{W}{\Omega}}$	ΔT 50°C $J = \frac{amps}{cm^2}$	Weight f_e	Cu	Volume cm^3	A_c cm^2
1	AL-2	20.9	0.265	3.55	662 30	8.93	0.627	0.187	370	9.81	1.46	0.273	538	12.2	11.1	7.14	0.265
2	AL-3	23.9	0.410	4.18	662 30	10.5	0.717	0.185	365	11.5	1.67	0.269	531	18.1	13.1	9.92	0.410
3	AL-5	33.6	0.767	4.59	946 30	16.5	1.01	0.174	345	18.1	2.35	0.255	503	31.3	20.5	14.06	0.539
4	AL-6	37.5	1.011	5.23	946 30	18.8	1.13	0.172	341	20.6	2.63	0.253	490	41.7	23.4	16.88	0.716
5	AL-124	45.3	1.44	5.50	1317 30	27.5	1.36	0.157	310	30.2	3.17	0.229	452	46.6	34.2	22.50	0.716
6	AL-8	63.4	2.31	5.74	221 20	0.482	1.90	1.404	271	0.529	4.44	2.05	395	67.9	60.0	35.66	0.806
7	AL-9	69.0	3.09	6.38	221 20	0.535	2.07	1.39	268	0.587	4.83	2.03	391	89.2	66.6	41.62	1.077
8	AL-10	74.5	3.85	7.01	221 20	0.588	2.24	1.38	266	0.646	5.22	2.01	387	110.0	73.2	47.55	1.342
9	AL-12	87.0	4.57	7.09	278 20	0.748	2.61	1.32	255	0.821	6.09	1.93	371	111.0	93.2	61.38	1.26
10	AL-135	93.7	5.14	7.36	325 20	0.908	2.81	1.24	240	0.997	6.56	1.81	345	114.0	113.0	69.63	1.26
11	AL-78	98.1	6.07	7.01	312 20	0.831	2.94	1.33	256	0.912	6.87	1.94	374	155.0	103.0	62.83	1.34
12	AL-18	118	7.92	7.61	510 20	1.47	3.55	1.10	211	1.61	8.26	1.60	308	138.0	183.0	94.79	1.25
13	AL-15	120	9.07	8.05	386 20	1.18	3.58	1.23	237	1.30	8.40	1.79	346	205.0	147.0	94.43	1.80
14	AL-16	127	10.8	8.89	386 20	1.30	3.80	1.20	233	1.43	8.89	1.76	340	235.0	162.0	104.95	2.15
15	AL-17	142		10.3	386 20	1.51	4.25	1.185	228	1.66	9.94	1.73	333	314.0	188.0	124.94	2.87
16	AL-19	159	18.0	10.8	511 20	2.10	4.77	1.065	205	2.31	11.1	1.55	299	328.0	261.0	155.44	2.87
17	AL-20	182	22.6	11.5	511 20	2.23	5.46	[illegible]	[illegible]	2.45	12.[illegible]	1.61	310	437.0	278.0	187.08	3.58
18	AL-22	202	28.0	11.5	637 20	2.78	6.05	1.0[illegible]	201	3.05	14.1	1.52	293	489.0	346.0	212.04	3.58
19	AL-23	220	34.9	12.7	637 20	3.07	6.60	1.036	200	3.37	15.4	1.51	291	612.0	382.0	244.67	4.48
20	AL-24	245	40.0	12.0	948 20	4.32	7.35	0.922	179	4.74	17.1	1.35	259	552.0	53[illegible]	[illegible]	3.58

copper loss = iron loss

77-35

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Definitions for Table 2-6

Information given is listed by column as:

1. Manufacturer part number
2. Surface area calculated from Figure 2-25
3. Area product effective iron area times window area
4. Mean length turn on one bobbin
5. Total number of turns and wire size for a single bobbin using a window utilization factor $K_u = 0.40$
6. Resistance of the wire at 50°C
7. Watts loss is based on Figure 7-2 for a ΔT of 25°C with a room ambient of 25°C surface dissipation times the transformer surface area, total loss is P_{cu}
8. Current calculated from column 6 and 7
9. Current density calculated from column 5 and 8
10. Resistance of the wire at 75°C
11. Watts loss is based on Figure 7-2 for a ΔT of 50°C with a room ambient of 25°C surface dissipation times the inductor surface area, total loss is P_{cu}
12. Current calculated from column 10 and 11
13. Current density calculated from column 5 and 12
14. Effective core weight plus copper weight in grams
15. Inductor volume calculated from Figure 2-9
16. Core effective cross-section

Table 2-6. Single-coil C-core characteristics

	1	2	3	4	5	6	7	8	9	10	11	12	13	14		15	16
	Core	$A_t cm^2$	$A_p cm^4$	MLT_{cm}	$N/_{AWG}$	Ω @ 50°C	P_Σ	$I = \sqrt{\frac{W}{\Omega}}$	ΔT 25°C $J = I/cm^2$	Ω @ 75°C	P_Σ	$I = \sqrt{\frac{W}{\Omega}}$	ΔT 50°C $J = I/cm^2$	Weight f_e	Cu	Volume cm^3	$A_c cm^2$
1	AL-2	24.6	0.265	4.47	83/20	0.138	0.737	2.31	445	0.151	1.72	3.37	651	12.2	16.9	10.7	0.264
2	AL-3	27.6	0.410	5.10	83/20	0.158	0.828	2.28	441	0.173	1.93	3.34	644	18.1	19.3	12.5	0.406
3	AL-5	38.1	0.767	5.42	119/20	0.238	1.14	2.18	422	0.26[illegible]	2.67	3.19	615	31.3	29.2	19.7	0.539
4	AL-6	41.9	1.011	6.06	119/20	0.266	1.26	2.17	420	0.292	2.93	3.16	611	41.7	32.6	[illegible].9	0.716
5	AL-124	51.8	1.44	6.56	175/20	0.426	1.55	1.90	368	0.468	3.63	2.78	537	46.6	52.1	30.8	0.716
6	AL-8	72.8	2.31	7.06	255/20	0.669	2.18	1.80	348	0.734	5.10	2.63	508	67.9	81.7	53.5	0.806
7	AL-9	78.4	3.09	7.69	255/20	0.728	2.35	1.79	346	0.799	5.49	2.62	505	89.2	89.0	59.5	1.08
8	AL-10	83.9	3.85	8.33	255/20	0.788	2.52	1.78	345	0.866	5.87	2.60	502	110.0	96.4	65.4	1.34
9	AL-12	101.0	4.57	9.00	327/20	1.09	3.03	1.66	321	1.20	7.07	2.42	468	111.0	134.4	92.1	1.26
10	AL-135	110.0	5.14	9.50	370/20	1.31	3.30	1.58	306	1.43	7.70	2.32	447	114.0	159.0	107.0	1.26
11	AL-78	110.0	6.08	8.15	406/20	1.23	3.30	1.63	316	1.35	7.70	2.38	460	155.0	150.0	81.3	1.34
12	AL-18	142.0	7.87	7.51	564/20	2.14	4.26	1.41	272	2.35	9.94	2.05	396	138.0	260.0	147.0	1.25
13	AL-15	136.0	9.07	10.1	444/20	1.66	4.08	1.56	302	1.83	9.52	2.28	440	205.0	203.0	136.0	1.80
14	AL-16	143.0	10.8	10.7	444/20	1.77	4.29	1.55	300	1.94	10.0	2.27	438	235.0	216.0	147.0	2.15
15	AL-17	158.0	14.4	12.0	444/20	1.97	4.74	1.55	299	2.20	11.1	2.24	433	314.0	241.0	168.0	2.87
16	AL-19	182.0	18.1	13.0	563/20	2.71	5.46	1.41	274	2.97	12.7	2.06	399	328.0	332.0	212.0	2.87
17	AL-20	205.0	22.6	13.6	563/20	2.84	6.15	1.47	284	3.12	14.4	2.14	414	437.0	348.0	259.0	3.58
18	AL-22	228.0	28.0	13.6	704/20	3.56	6.84	1.38	267	3.91	16.0	2.02	390	489.0	435.0	294.0	3.58
19	AL-23	246.0	35.0	15.9	704/20	3.89	7.38	1.37	265	4.27	17.2	2.00	387	612.0	479.0	326.0	4.48
20	AL-24	282.0	40.0	14.6	1026	5.57	8.46	1.23	238	6.11	19.7	1.79	346	552.0	680.0	431.0	3.58

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Definitions for Table 2-7

Information given is listed by column as:

1. Manufacturer part number
2. Surface area calculated from Figure 2-22
3. Area product effective iron area times window area
4. Mean length turn
5. Total number of turns and wire size using a window utilization factor $K_u = 0.40$
6. Resistance of the wire at 50°C
7. Watts loss is based on Figure 7-2 for a ΔT of 25°C with a room ambient of 25°C surface dissipation times the transformer surface area, total loss is equal to 2 P_{cu}
8. Current calculated from column 6 and 7
9. Current density calculated from column 5 and 8
10. Resistance of the wire at 75°C
11. Watts loss is based on Figure 7-2 for a ΔT of 50°C with a room ambient of 25°C surface dissipation times the transformer surface area, total loss is equal to 2 P_{cu}
12. Current calculated from column 10 and 11
13. Current density calculated from column 5 and 12
14. Effective core weight plus copper weight in grams
15. Transformer volume calculated from Figure 2-6
16. Core effective cross-section

Table 2-7. Tape-wound core characteristics

	1	2	3	4	5		6	7	8	9	10	11	12	13	14		15	16
	Core	A_t cm²	A_p cm⁴	MLT cm	N	AWG	Ω @ 50°C	P_Σ	$I = \sqrt{W/\Omega}$	ΔT 25°C $J = I/cm^2$	Ω @ 75°C	P_Σ	$I = \sqrt{W/\Omega}$	ΔT 50°C $J = I/cm^2$	Weight f_e	Cu	Volume cm³	A_c cm²
1	52402	7.26	0.0100	2.05	302	30	2.35	0.218	0.215	425	2.58	0.508	0.313	619	0.63	3.12	1.42	0.022
2	52153	8.29	0.0196	2.22	302	30	2.54	0.249	0.221	436	2.80	0.580	0.322	636	1.31	3.29	1.71	0.053
3	52167	11.1	0.0201	2.21	606	30	5.09	0.333	0.180	357	5.59	0.777	0.263	520	0.80	6.84	2.63	0.022
4	52403	13.5	0.0267	2.30	621	30	5.43	0.405	0.193	381	5.96	0.945	0.281	556	0.88	9.52	3.48	0.022
5	52057	17.4	0.0659	2.53	1017	30	9.78	0.522	0.163	322	10.7	1.22	0.238	471	2.05	13.1	4.98	0.043
6	52000	15.2	0.0787	2.70	606	30	6.22	0.456	0.191	378	6.82	1.06	0.278	550	3.73	7.97	3.99	0.086
7	52063	20.7	0.132	2.85	1017	30	11.0	0.621	0.167	331	12.1	1.45	0.244	48[illegible]	4.47	14.4	6.20	0.086
8	52002	21.8	0.144	2.88	1114	30	12.2	0.654	0.163	323	13.4	1.53	0.239	472	4.62	16.0	6.72	0.086
9	52007	27.6	0.380	3.87	982	30	14.4	0.828	0.169	334	15.8	1.93	0.24[illegible]	487	14.5	17.7	9.84	0.257
10	52167	31.5	0.516	4.23	1000	30	16.1	0.945	0.171	338	17.6	2.21	0.250	494	20.9	19.0	11.9	0.343
11	52094	30.4	0.592	4.47	1017	30	17.3	0.912	0.162	321	19.0	2.13	0.237	468	21.8	21.0	12.2	0.386
12	52004	46.1	0.725	4.02	315	20	0.469	1.38	1.20	234	0.515	3.23	1.77	341	13.4	56.[illegible]	21.3	0.171
13	52032	56.5	1.46	4.65	315	20	0.543	1.69	1.25	240	0.596	3.95	1.82	351	29.8	63.7	27.8	0.343
14	52026	61.0	2.18	5.28	315	20	0.616	1.83	1.22	235	0.676	4.27	1.77	342	44.7	71.3	32.8	0.514
15	52038	65.9	2.91	5.97	315	20	0.6[illegible]7	1.98	1.19	230	0.765	4.61	1.74	334	59.6	79.4	38.3	0.666
16	52035	88.9	4.68	6.33	505	20	1.19	2.67	1.06	204	1.3	6.22	1.55	298	71.5	138.0	59.0	0.686
17	52055	116.0	6.81	6.76	737	20	1.85	3.48	0.970	187	2.0	8.12	1.42	273	83.4	220.0	86.4	0.686
18	52012	110.0	9.35	8.88	505	20	1.66	3.30	0.996	192	1.82	7.70	1.45	280	143.0	235.0	87.4	1.371
19	52017	179.0	12.5	7.51	698	17	0.97	5.37	1.66	160	1.065	12.5	2.33	274	107.0	455.0	163.0	0.686
20	52031	256.0	19.8	8.23	1114	17	1.70	7.68	1.50	145	1.86	17.9	2.19	211	131.0	800.0	272.0	0.686
21	52103	220.0	24.5	8.77	688	17	1.12	6.60	1.72	165	1.23	15.4	2.51	241	238.0	503.0	212.0	1.371
22	52128	304.0	39.4	9.49	1104	17	1.94	9.12	1.53	147	2.13	21.3	2.24	215	286.0	896.0	341.0	1.371
23	52022	256.0	49.1	11.3	688	17	1.44	7.68	1.63	157	1.58	17.9	2.38	229	477.0	629.0	291.0	42
24	52042	347.0	78.7	12.0	1104	17	2.45	10.4	1.45	140	2.69	24.3	2.12	204	572.0	1109.0	453.0	2.742
25	52100	422.0	145.0	15.4	1089	17	3.11	12.7	1.43	138	3.41	29.5	2.08	200	1117.0	1342.0	633.0	5.142
26	52112	878.0	510.0	20.3	2871	17	10.8	26.3	1.1	106	11.8	61.5	1.61	155	2205.0	4895.0	1891.0	6.855
27	52426	1014.0	813.0	22.2	2856	17	11.7	24.4	1.02	98.1	12.9	71.0	1.66	159	3814.0	5077.0	2299.0	10.968

copper loss = iron loss

C. TRANSFORMER VOLUME

The volume of a transformer can be related to the area product A_p of a transformer, treating the volume as shown in Figures 2-6 through 2-9 below as solid quantity without subtraction of anything for the core window. Derivation of the relationship is according to the following: volume varies in accordance with the cube of any linear dimension ℓ (designated ℓ^3 below), where area product A_p varies as the fourth power:

$$\text{Vol} = K_1 \ell^3 \tag{2-1}$$

$$A_p = K_2 \ell^4 \tag{2-2}$$

$$\ell^4 = \frac{A_p}{K_2} \tag{2-3}$$

$$\ell = \left(\frac{A_p}{K_2}\right)^{0.25} \tag{2-4}$$

$$\ell^3 = \left[\left(\frac{A_p}{K_2}\right)^{0.25}\right]^3 = \left(\frac{A_p}{K_2}\right)^{0.75} \tag{2-5}$$

$$\text{Vol} = K_1 \left(\frac{A_p}{K_2}\right)^{0.75} \tag{2-6}$$

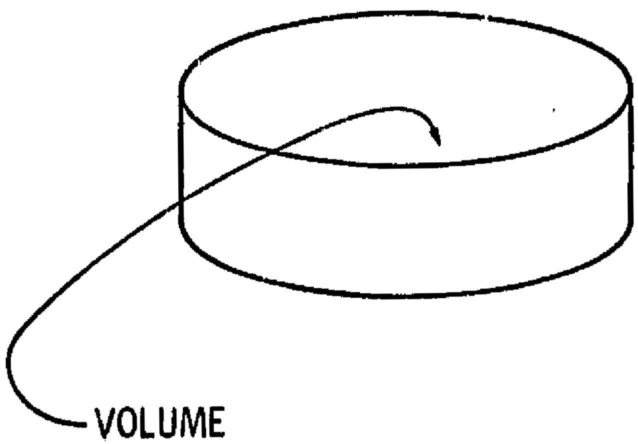

Fig. 2-6. Tape-wound core, powder core, and pot core volume

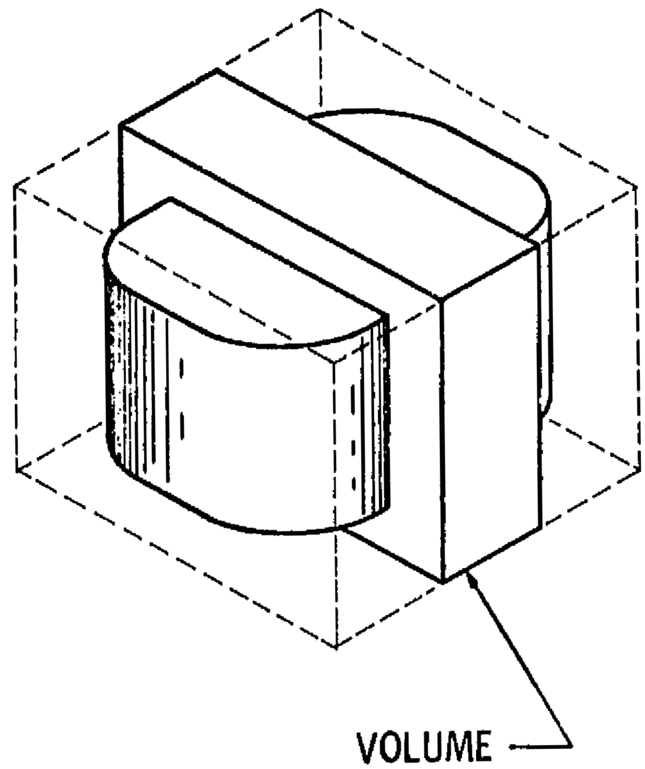

Fig. 2-7. EI Lamination core volume

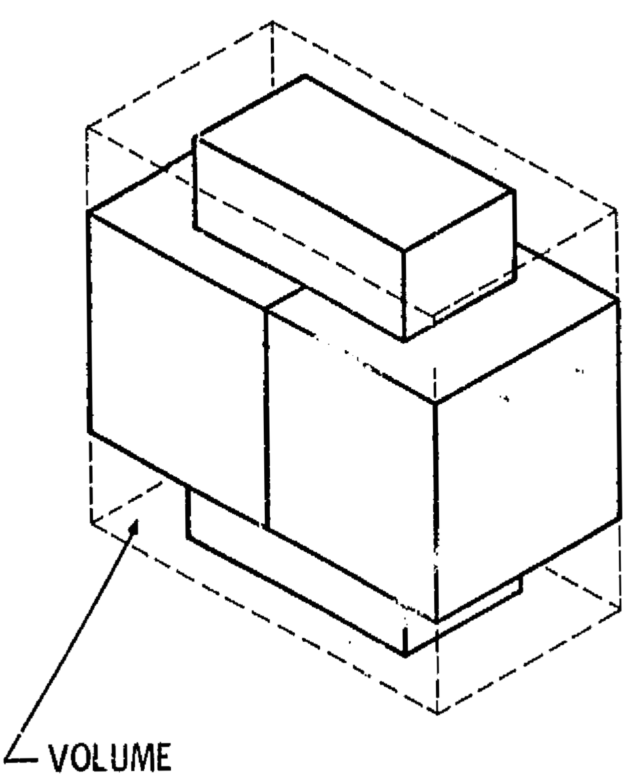

Fig. 2-8. C-core volume

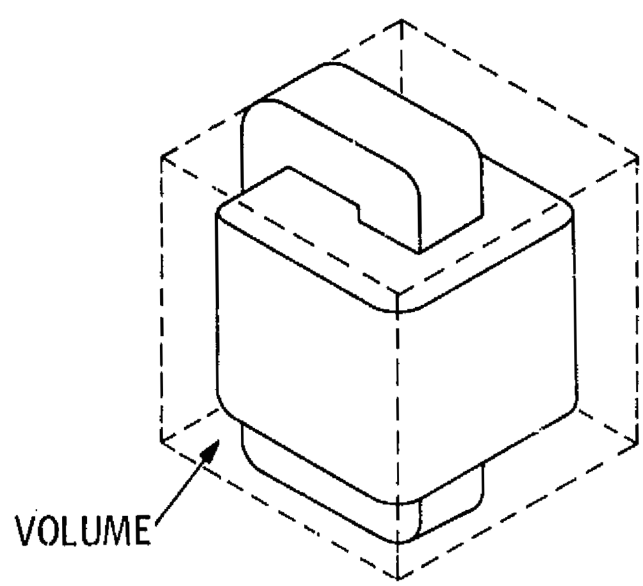

Fig. 2-9. Single-coil C-core volume

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$$K_v = \frac{K_1}{K_2^{0.75}} \quad (2\text{-}7)$$

$$Vol = K_v A_p^{0.75} \quad (2\text{-}8)$$

The volume/area product relationship is

$$Vol = K_v A_p^{0.75}$$

in which K_v is a constant related to core configuration, these values are given in Table 2-8. This constant was obtained by averaging the values in Tables 2-2 through 2-7, column 15.

The relationship between volume and area product A_p for various core types is given in Figures 2-10 through 2-15. It was obtained from the data shown in Tables 2-2 through 2-7, in which the Vol and A_p values are shown in columns 15 for volume, and column 3 for area product.

Table 2-8. Constant K_v

Core type	K_v
Pot core	14.5
Powder core	13.1
Lamination	19.7
C-core	17.9
Single-coil C-core	25.6
Tape-wound core	25.0

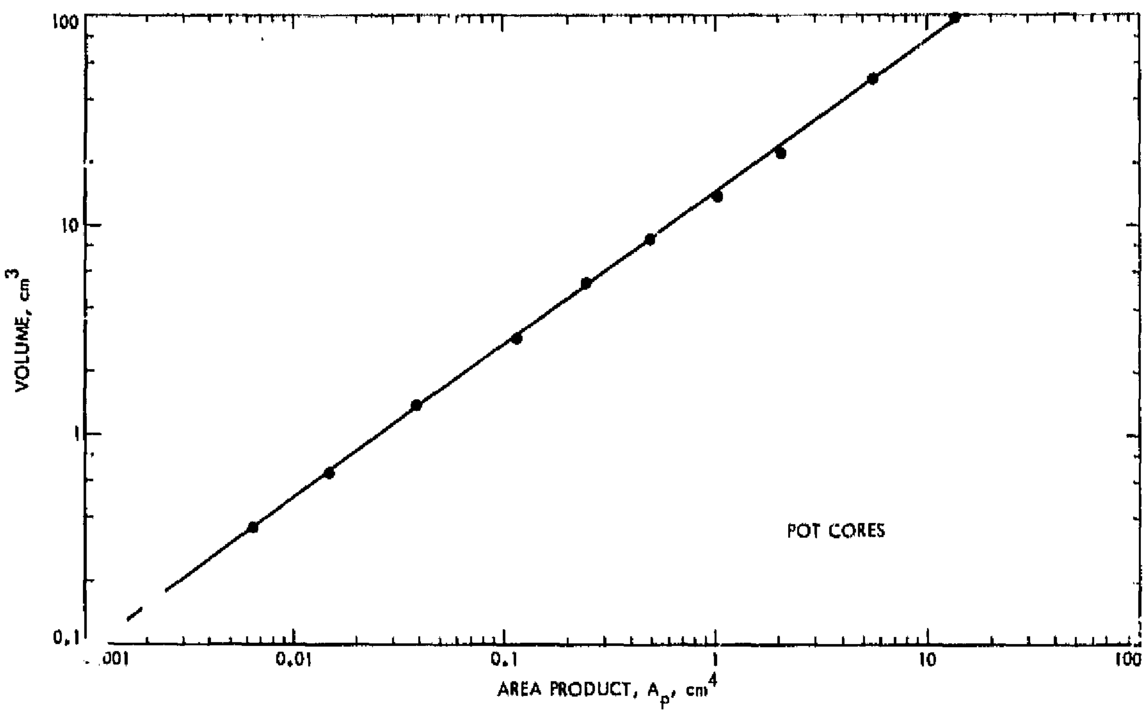

Fig. 2-10. Volume versus area product A_p for pot cores

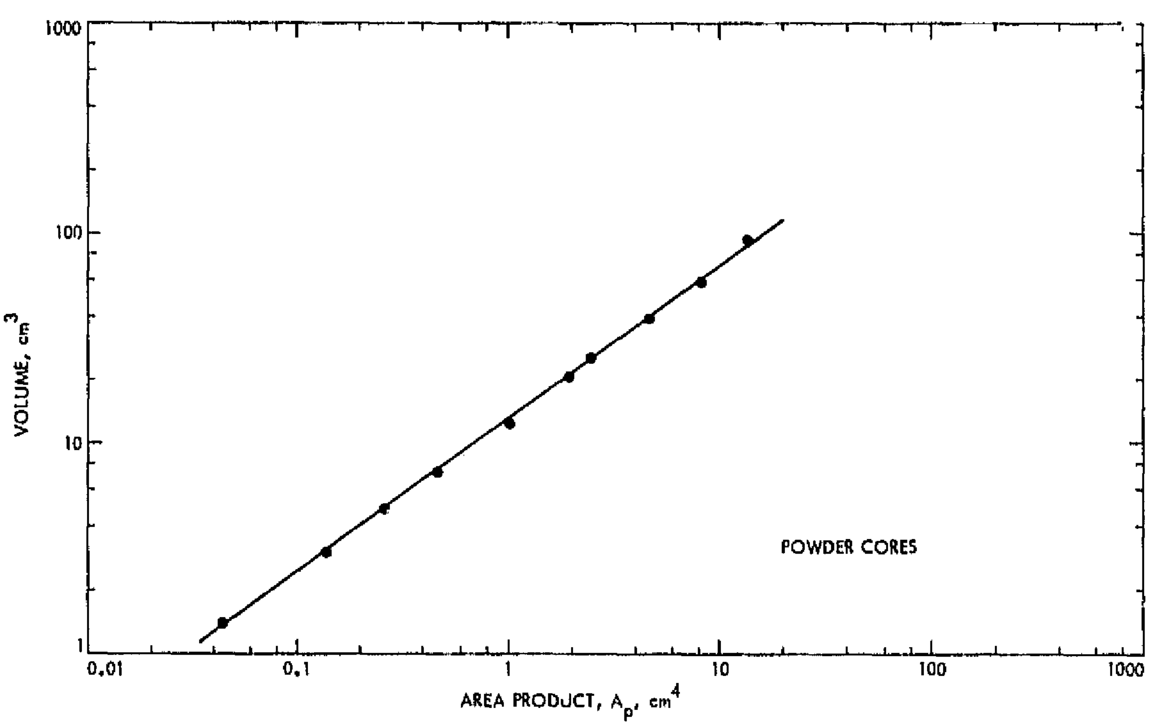

Fig. 2-11. Volume versus area product A_p for powder cores

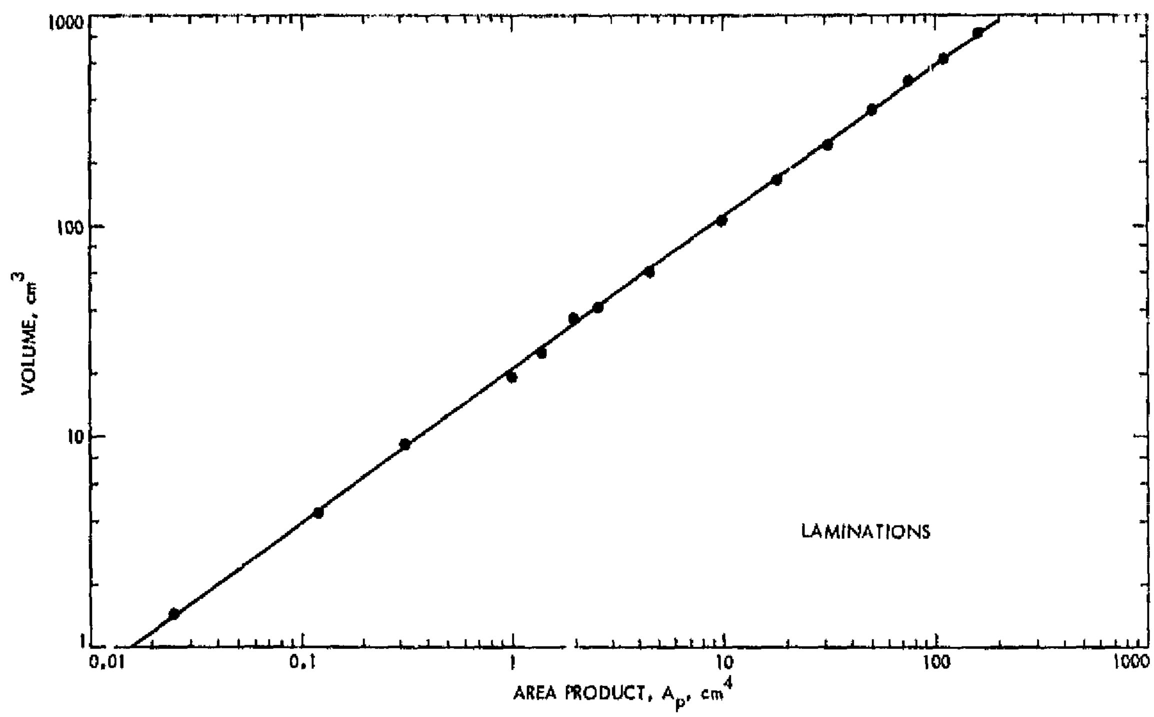

Fig. 2-12. Volume versus area product A_p for laminations

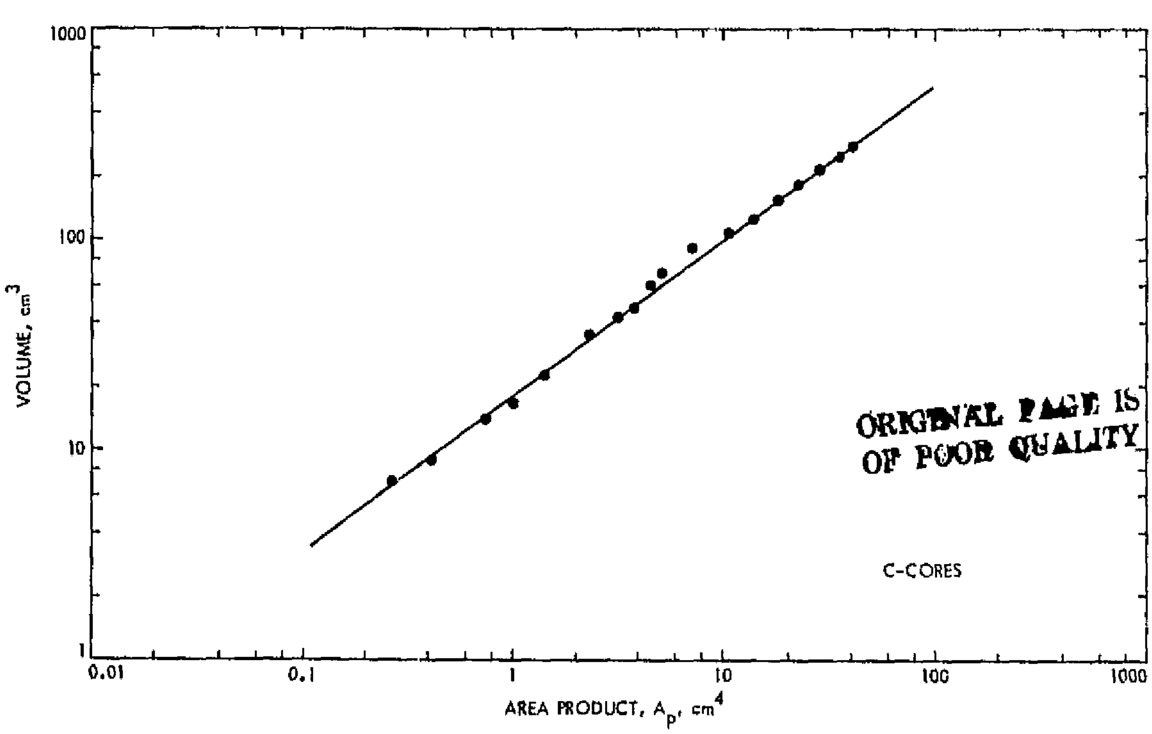

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Fig. 2-13. Volume versus area product A_p for C-cores

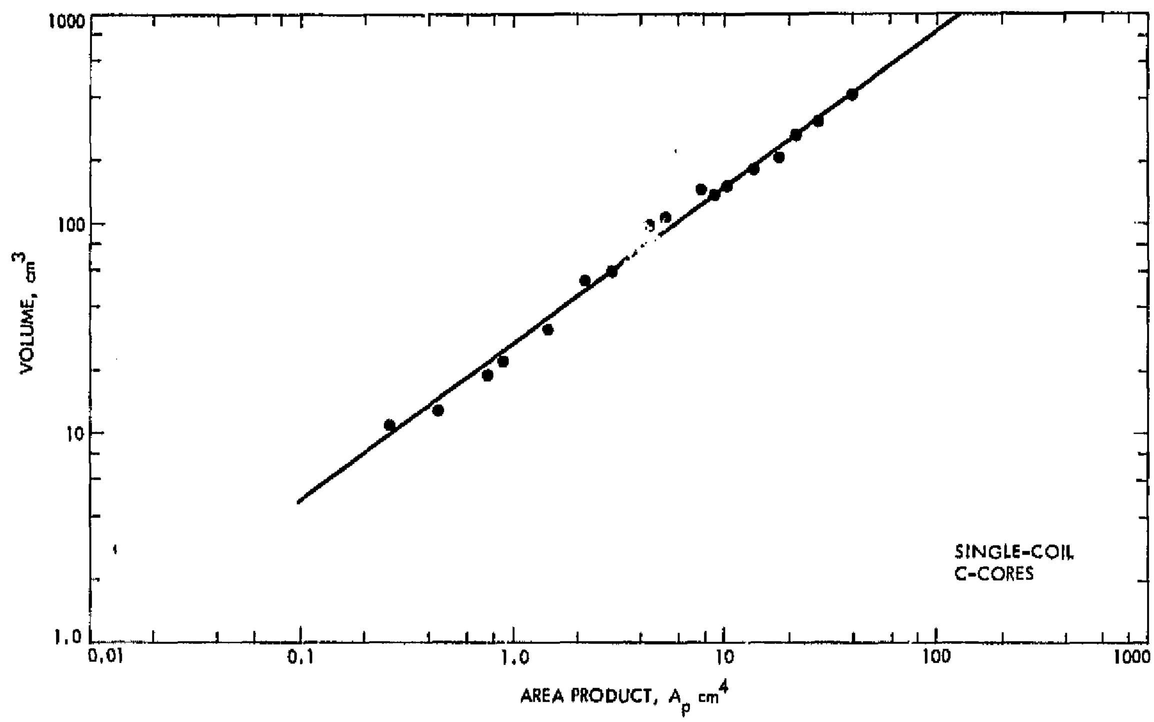

Fig. 2-14. Volume versus area product A_p for single-coil C-cores

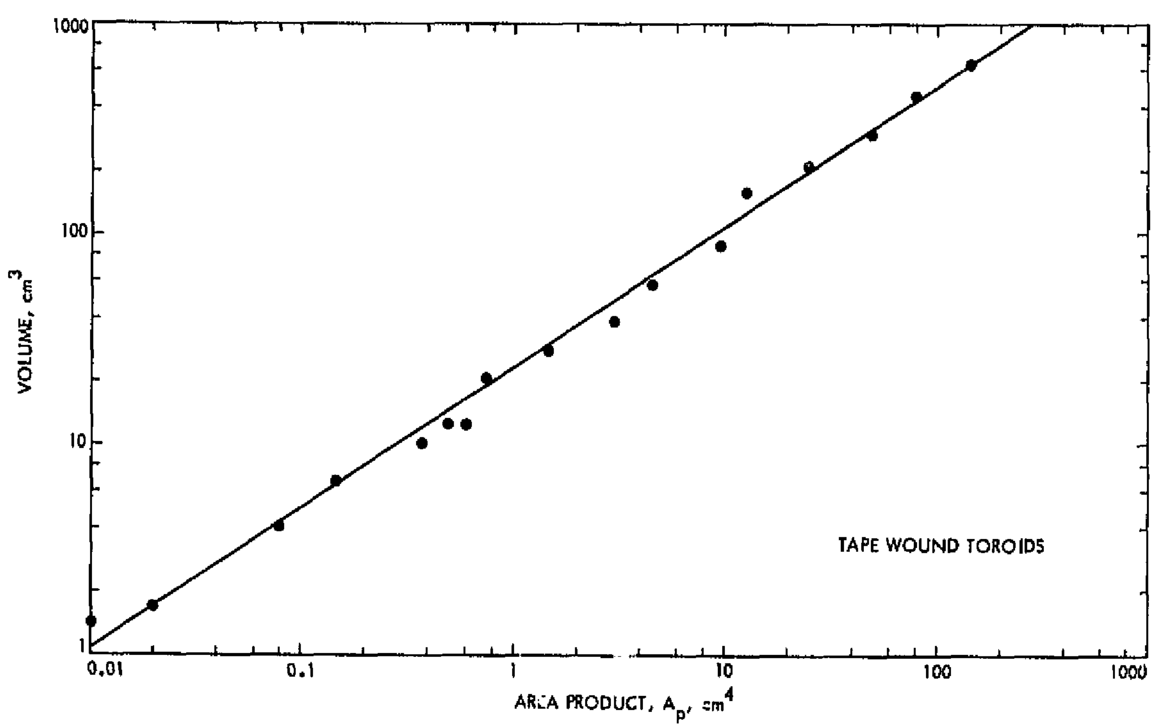

Fig. 2-15. Volume versus area product A_p for tape-wound toroids

IV. TRANSFORMER WEIGHT

The total weight W_t of a transformer can be related to the area product A_p. Derivation of the relationship is according to the following: weight W_t varies in accordance with the cube of any linear dimension ℓ (designated ℓ^3 below), whereas area product A_p varies as the fourth power:

$$W_t = K_3 \ell^3 \tag{2-9}$$

$$A_p = K_2 \ell^4 \tag{2-10}$$

$$\ell^4 = \frac{A_p}{K_2} \tag{2-11}$$

$$\ell = \left(\frac{A_p}{K_2}\right)^{0.25} \tag{2-12}$$

$$\ell^3 = \left[\left(\frac{A_p}{K_2}\right)^{0.25}\right]^3 = \left(\frac{A_p}{K_2}\right)^{0.75} \tag{2-13}$$

$$W_t = K_3 \left(\frac{A_p}{K_2}\right)^{0.75} \tag{2-14}$$

$$K_w = \frac{K_3}{K_2^{0.75}} \tag{2-15}$$

$$W_t = K_w A_p^{0.75} \tag{2-16}$$

The weight/area product relationship

$$W_t = K_w A_p^{0.75}$$

in which K_w is a constant related to core configuration, is shown in Table 2-9, which has been derived by averaging the values in Tables 2-2 through 2-7, column 14.

The relationship between weight and area product A_p for various core types is given in Figures 2-16 through 2-21. It was obtained from the data shown in Tables 2-2 through 2-7, in which the W_t and A_p values are shown in column 14 for weight, and column 3 for area product.

Table 2-9. Constant K_w

Core type	K_w
Pot core	48.0
Powder core	58.8
Lamination	68.2
C-core	66.6
Single-coil C-core	76.6
Tape-wound core	82.3

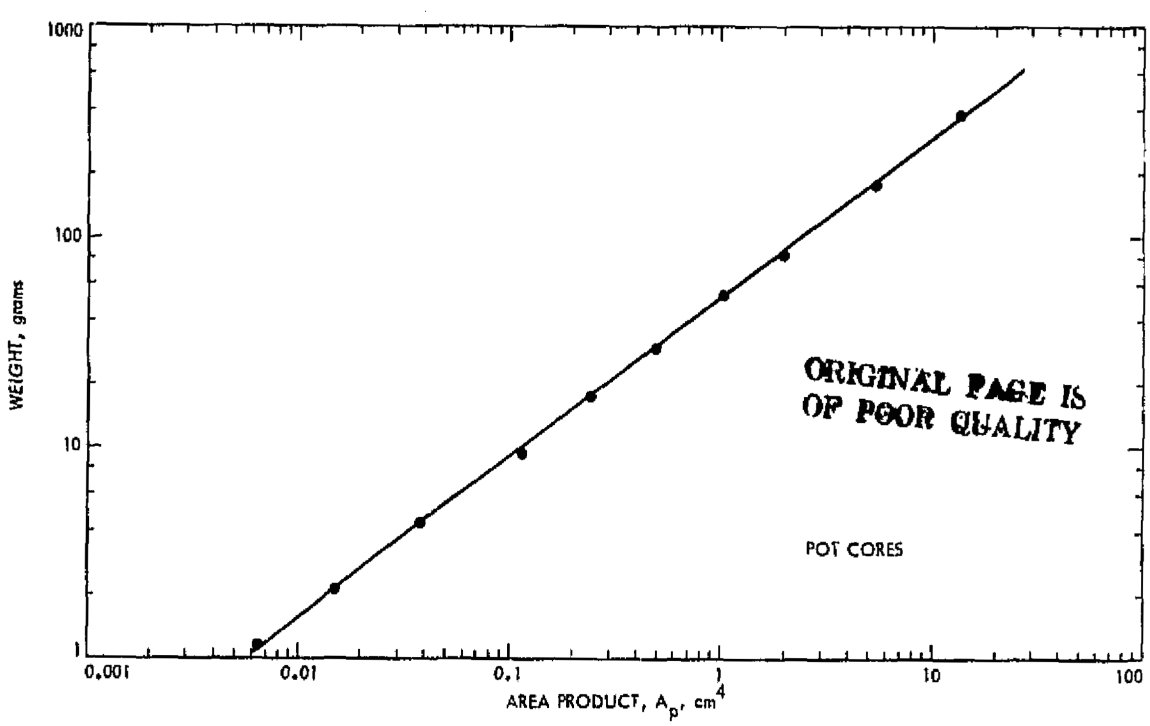

Fig. 2-16. Total weight versus area product A_p for pot cores

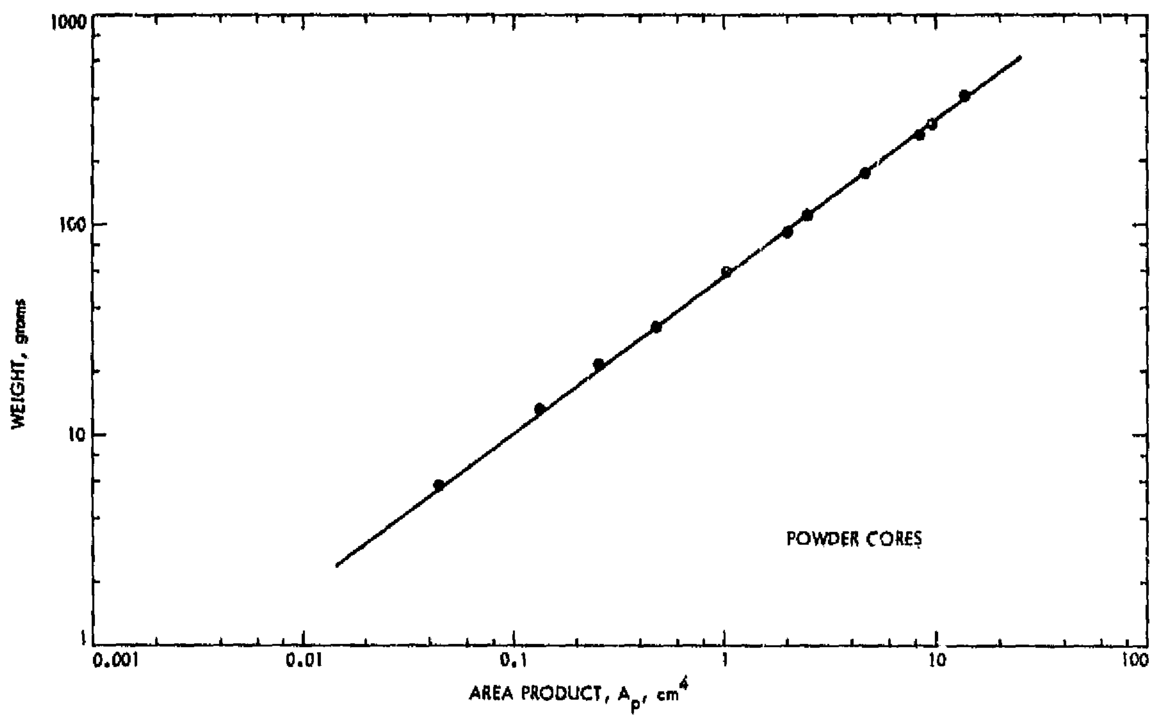

Fig. 2-17. Total weight versus area product A_p for powder cores

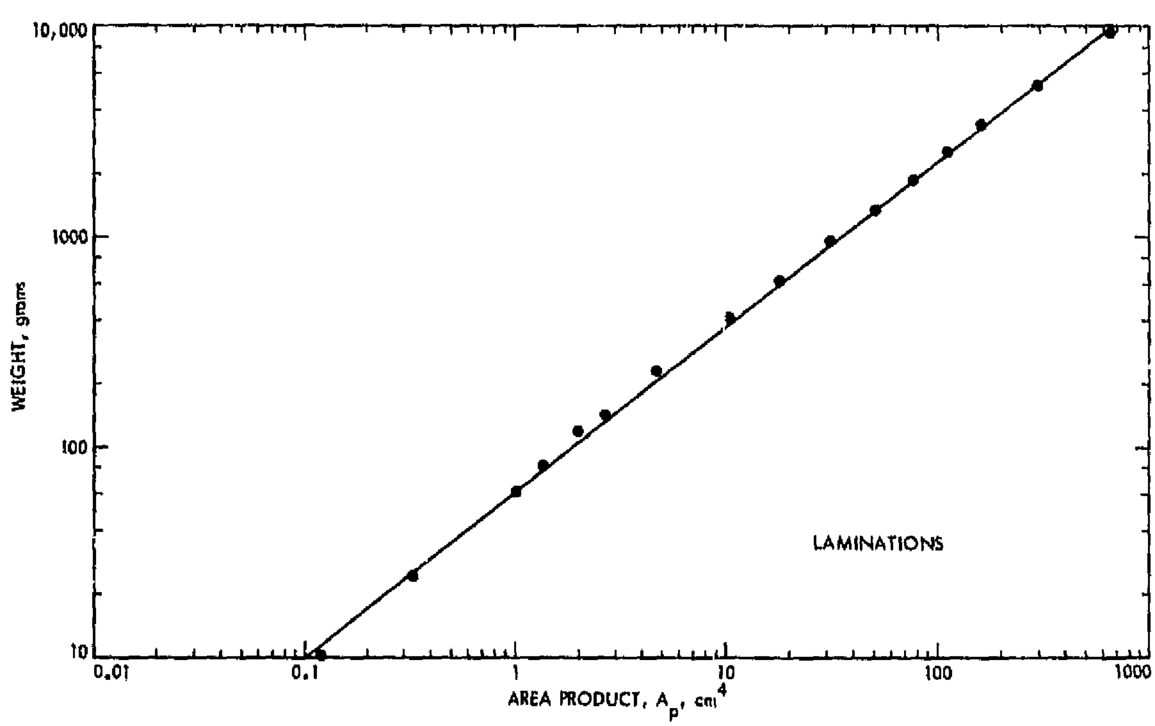

Fig. 2-18. Total weight versus area product A_p for laminations

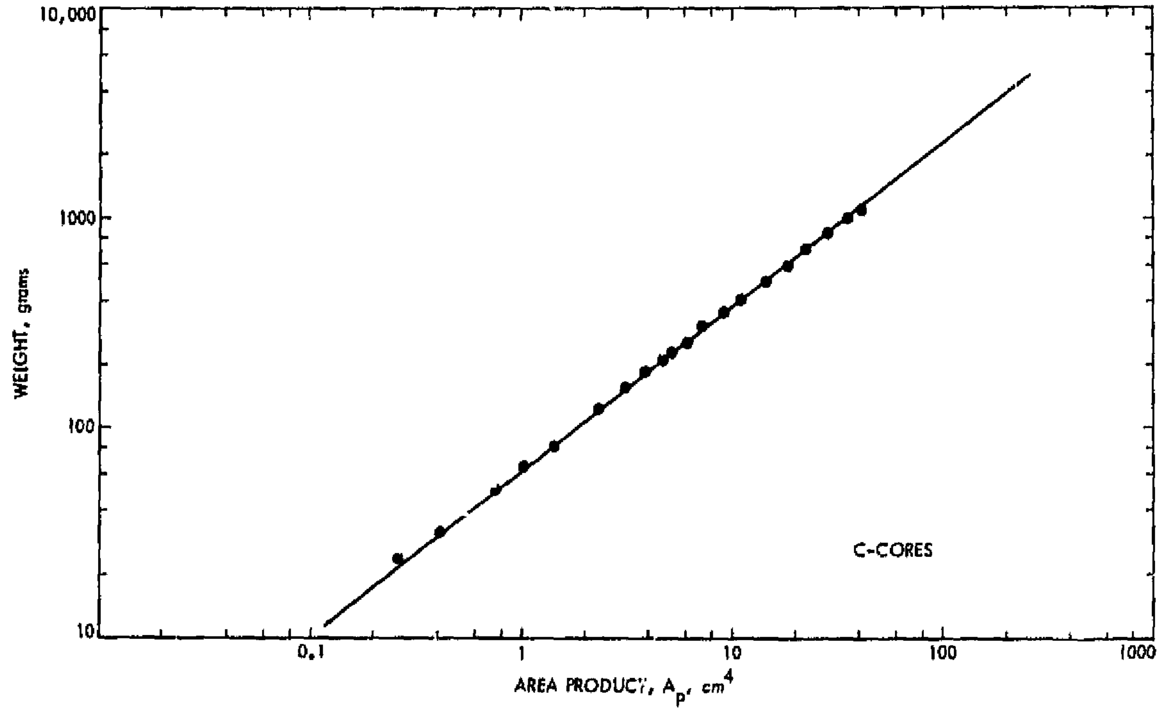

Fig. 2-19. Total weight versus area product A_p for C-cores

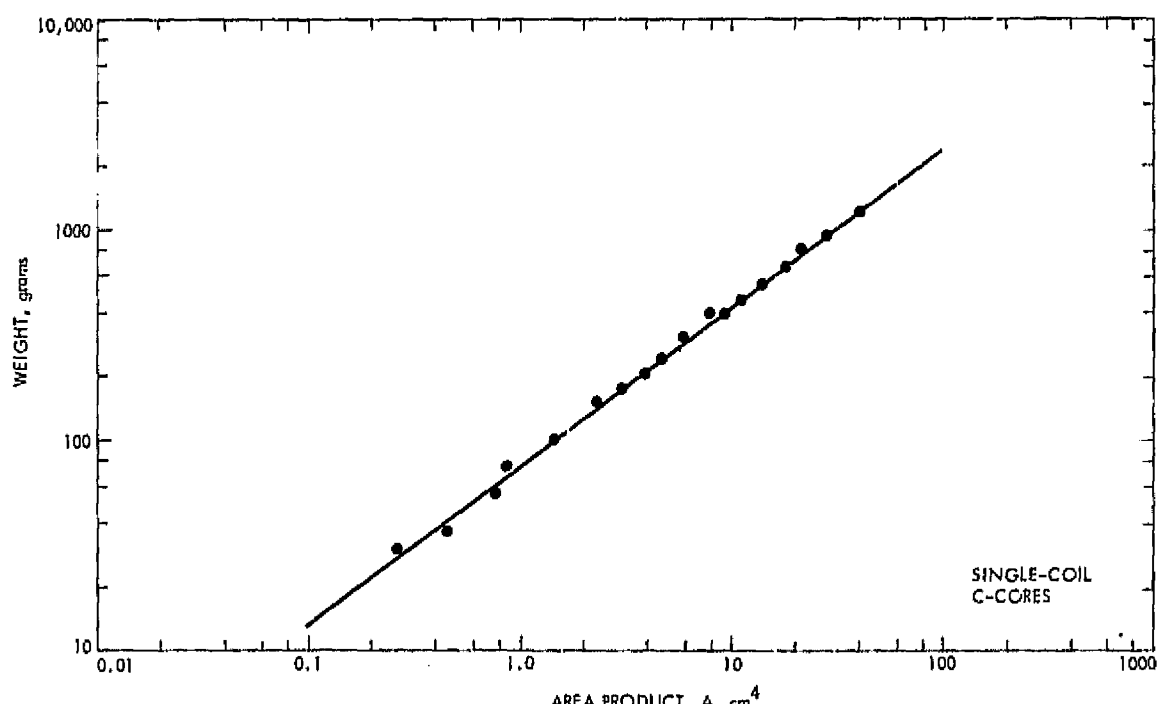

Fig. 2-20. Total weight versus area product A_p for single-coil C-cores

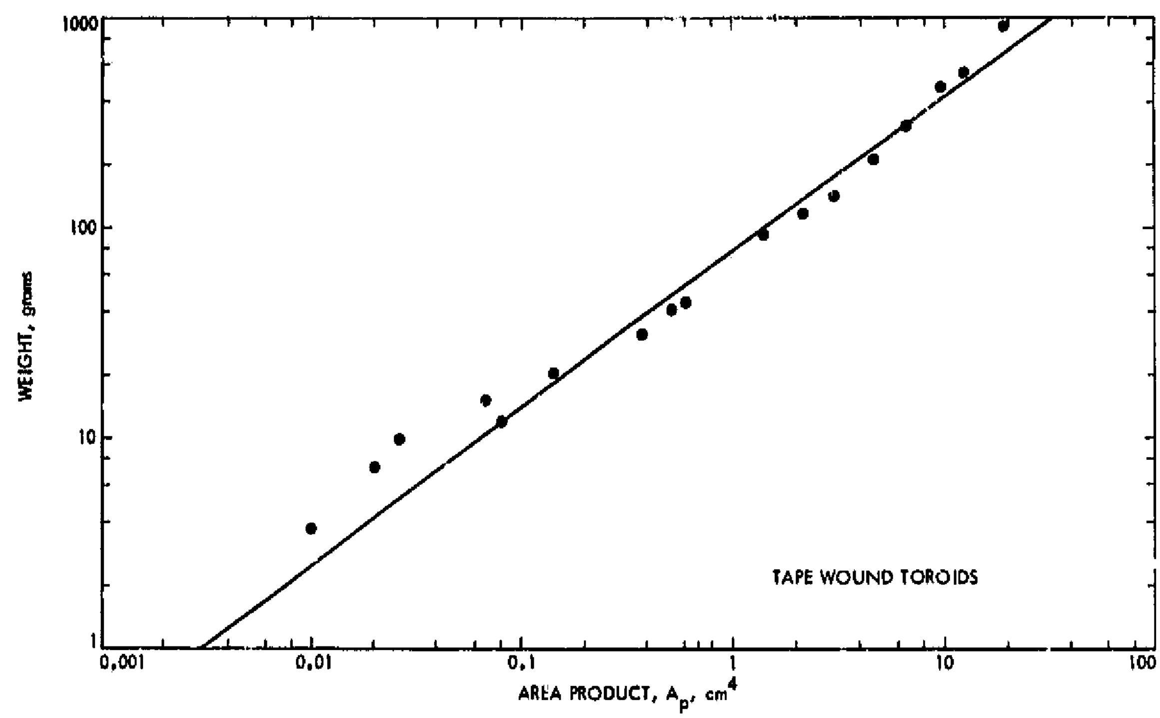

Fig. 2-21. Total weight versus area product A_p for tape-wound toroids

E. TRANSFORMER SURFACE AREA

The surface area A_t of a transformer can be related to the area product A_p of a transformer treating the surface area as shown in Figures 2-22 through 2-25. Derivation of the relationships is in accordance with the square of any linear dimension ℓ (designated ℓ^2 below), where area product varies as the fourth power:

$$A_t = K_4 \ell^2 \tag{2-17}$$

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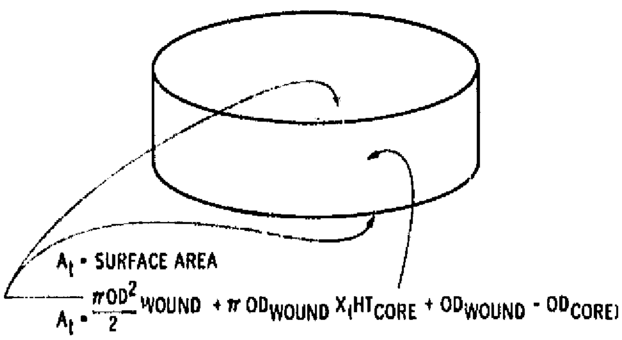

Fig. 2-22. Tape-wound core, powder core, and pot core surface area A_t

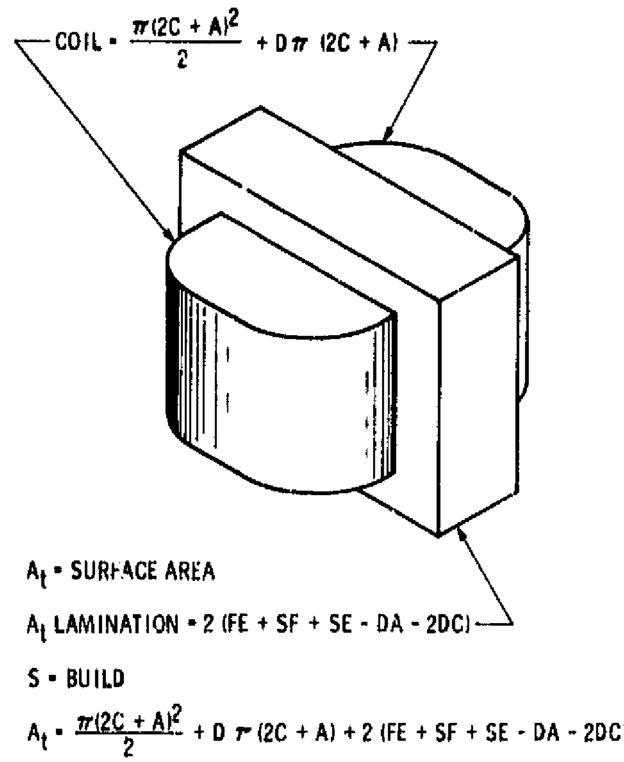

Fig. 2-23. EI lamination surface area A_t

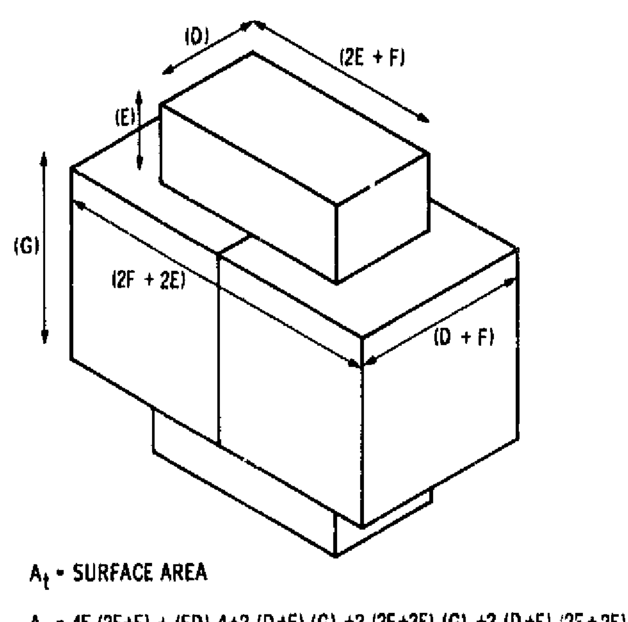

Fig. 2-24. C-core surface area A_t

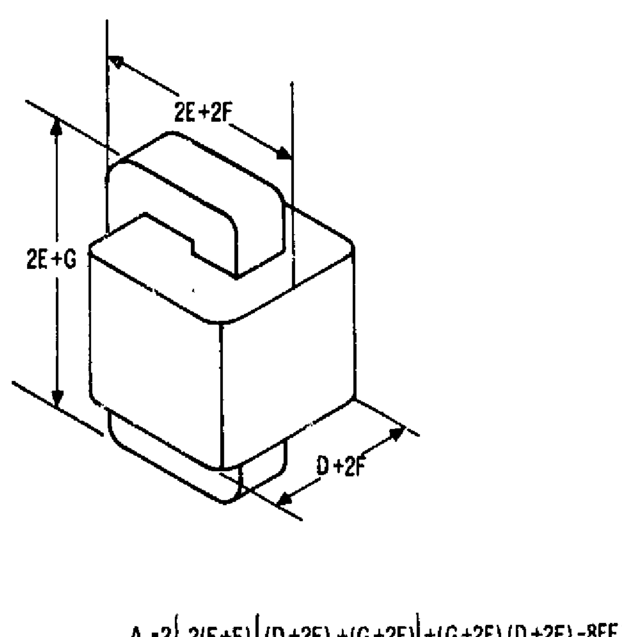

Fig. 2-25. Single-coil C-core surface area A_t

$$A_p = K_2 \ell^4 \tag{2-18}$$

$$\ell^4 = \frac{A_p}{K_2} \tag{2-19}$$

$$\ell = \left(\frac{A_p}{K_2}\right)^{0.25} \tag{2-20}$$

$$\ell^2 = \left[\left(\frac{A_p}{K_2}\right)^{0.25}\right]^2 \tag{2-21}$$

$$\ell^2 = \left(\frac{A_p}{K_2}\right)^{0.5} \tag{2-22}$$

$$A_t = K_4\left(\frac{A_p}{K_2}\right)^{0.5} \tag{2-23}$$

$$K_s = \frac{K_4}{K_2^{0.5}} \tag{2-24}$$

$$A_t = K_s A_p^{0.5} \tag{2-25}$$

The surface area/area product relationship

$$A_t = K_s A_p^{0.5}$$

in which K_s is a constant related to core configuration is shown in Table 2-10, which has been derived by averaging the values in Tables 2-2 through 2-7, column 2.

Table 2-10. Constant K_s

Core type	K_s
Pot core	33.8
Powder core	32.5
Lamination	41.3
C-core	39.2
Single-coil C-core	44.5
Tape-wound core	50.9

The relationship between surface area and area product A_p for various core types is given in Figures 2-26 through 2-31. It was obtained from the data shown in Tables 2-2 through 2-7, in which the A_t and A_p values are shown in columns 2 for surface area, and column 3 for area product.

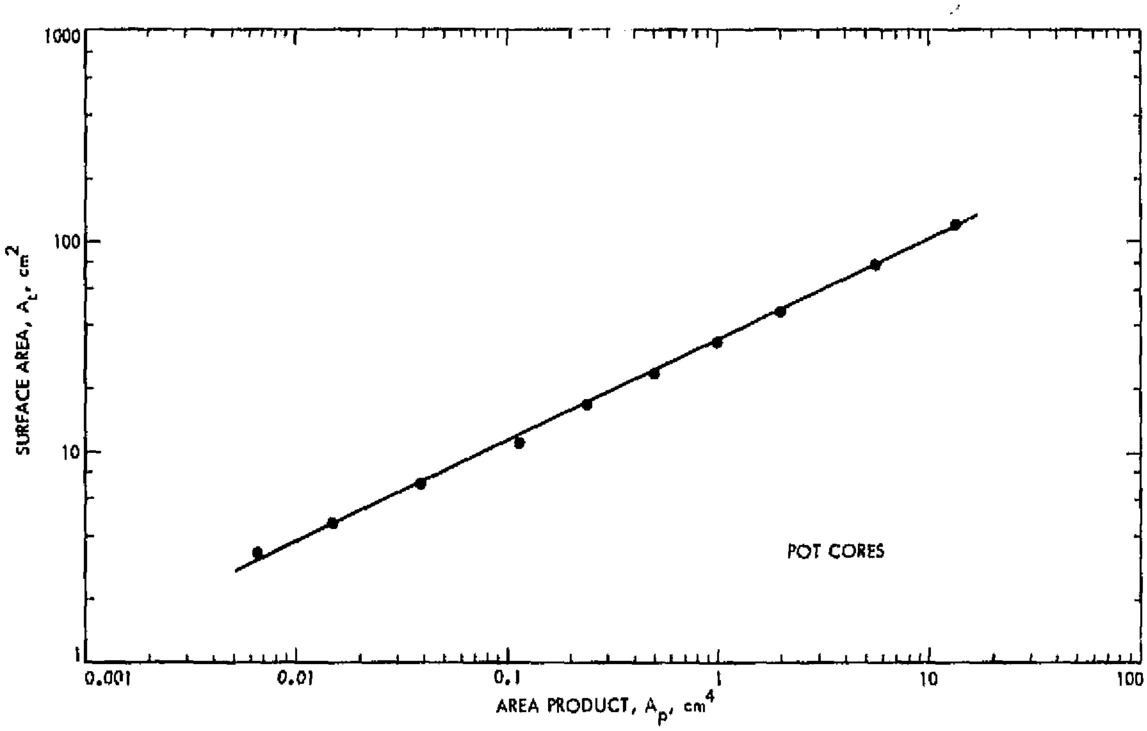

Fig. 2-26. Surface area versus area product A_p for pot cores

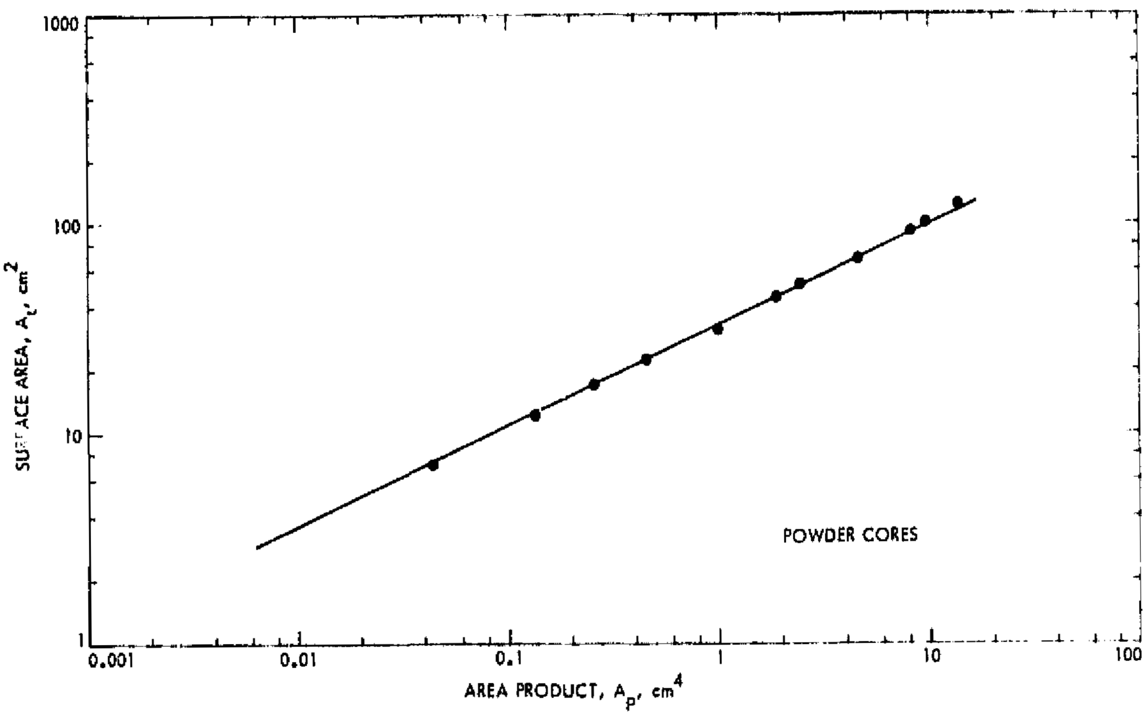

Fig. 2-27. Surface area versus area product A_p for powder cores

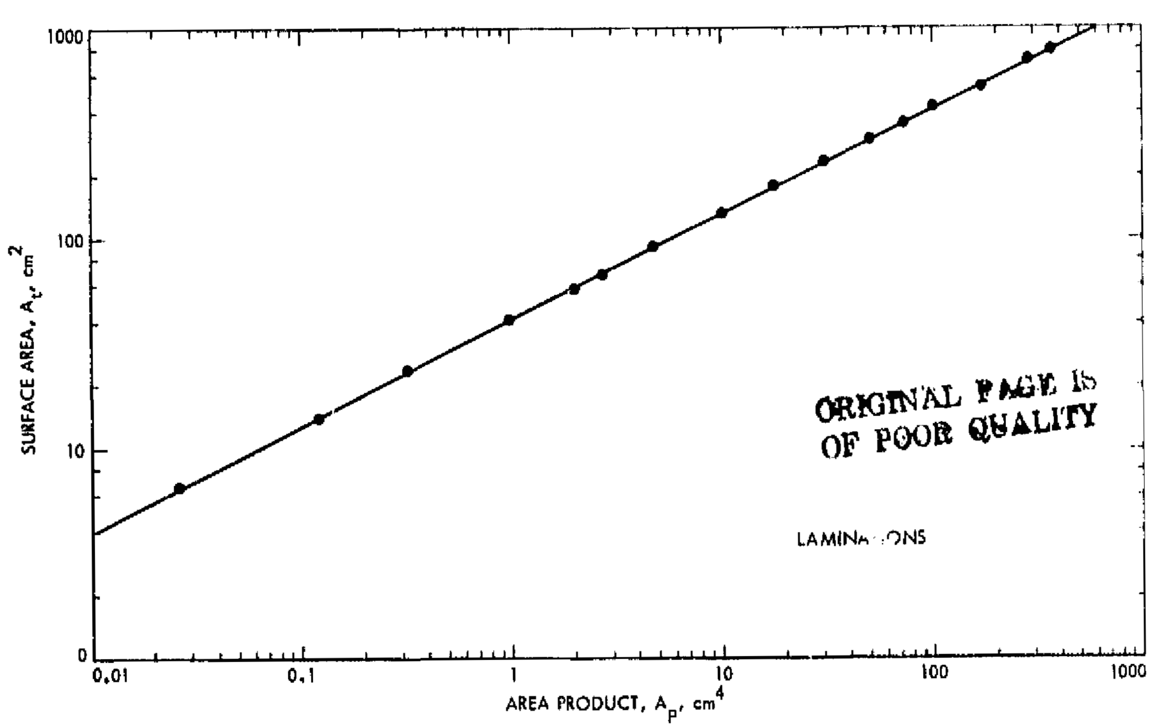

Fig. 2-28. Surface area versus area product A_p for laminations

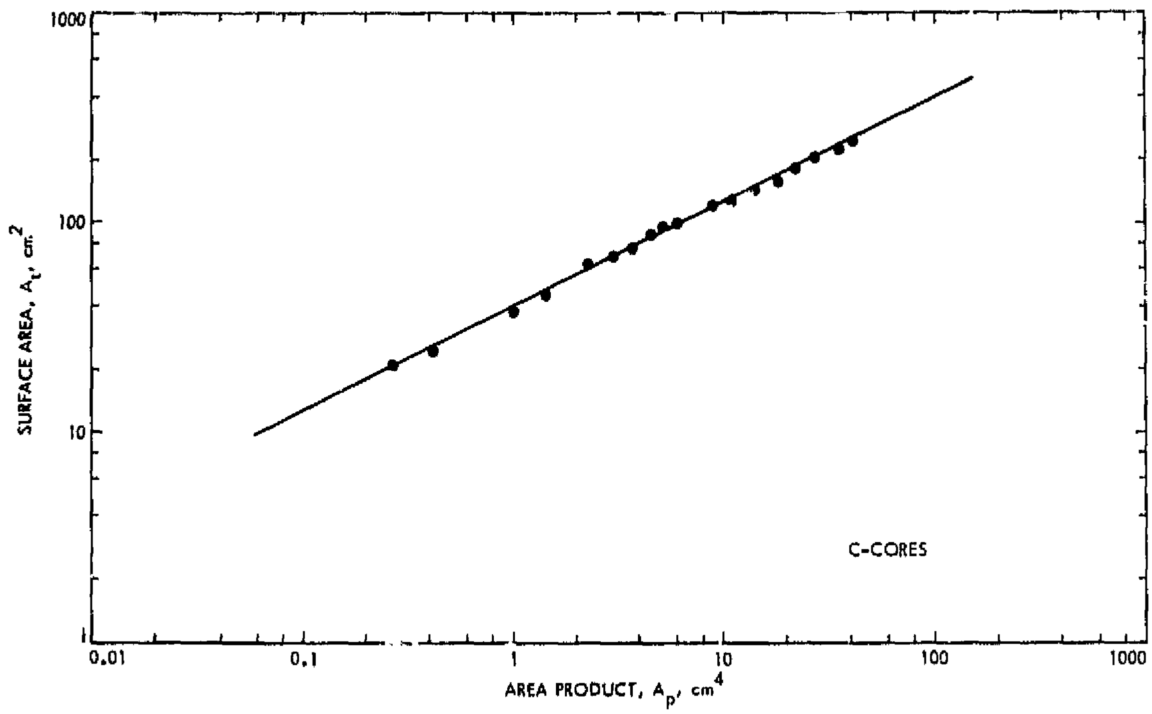

Fig. 2-29. Surface area versus area product A_p for C-cores

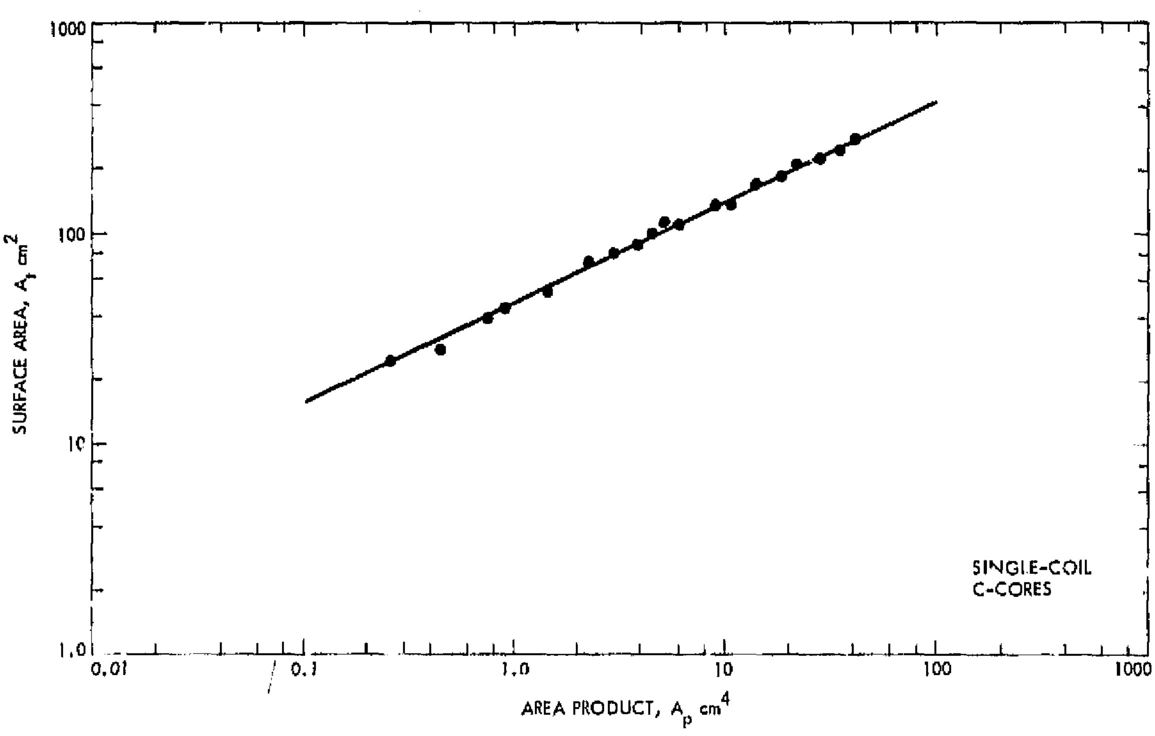

Fig. 2-30. Surface area versus area product A_p for single-coil C-cores

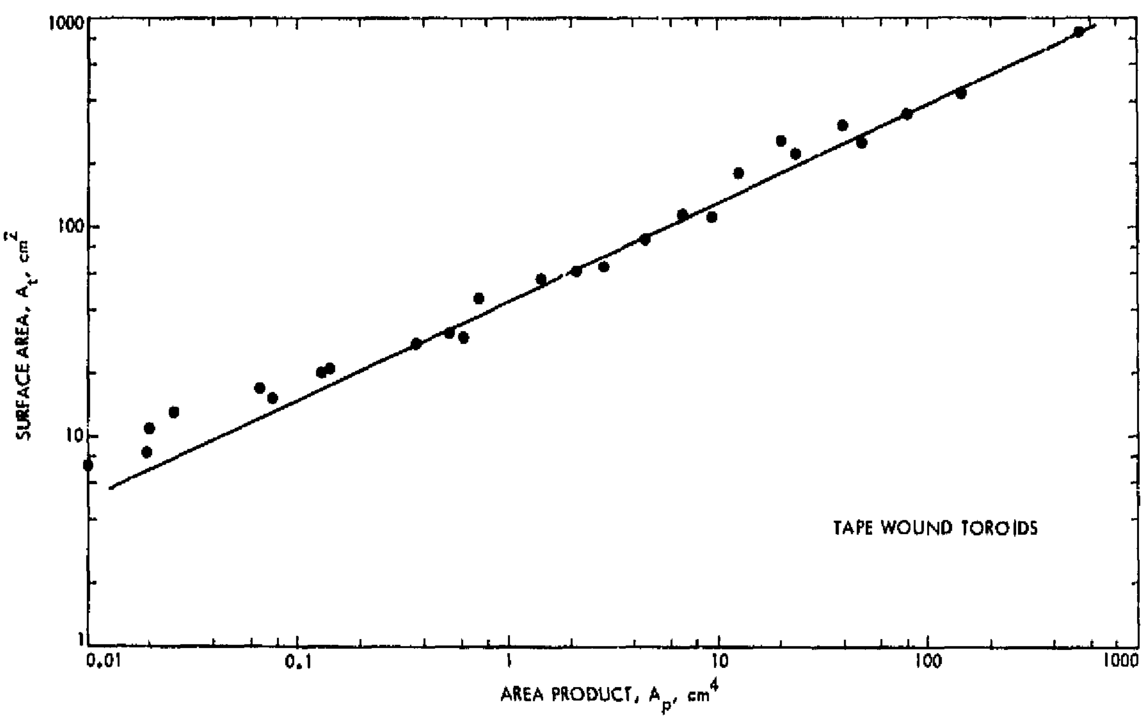

Fig. 2-31. Surface area versus area product A_p for tape-wound toroids

F. TRANSFORMER CURRENT DENSITY

Current density J of a transformer can be related to the area product A_p of a transformer for a given temperature rise.

The relationship of current density J to the area product A_p for a given temperature rise can be derived as follows:

$$A_t = K_s A_p^{0.5} \tag{2-26}$$

$$P_{cu} = I^2 R \tag{2-27}$$

$$I = A_w J \tag{2-28}$$

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$$\therefore P_{cu} = A_w^2 J^2 R \tag{2-29}$$

$$R = \frac{MLT}{A_w} N\rho \tag{2-30}$$

$$\therefore P_{cu} = A_w^2 J^2 \frac{MLT}{A_w} N\rho \tag{2-31}$$

$$P_{cu} = A_w J^2 MLT\ N\rho \tag{2-32}$$

Since MLT has a dimension of length

$$MLT = K_5 A_p^{0.25} \tag{2-33}$$

$$P_{cu} = A_w J^2 K_5 A_p^{0.25} N\rho \tag{2-34}$$

$$A_w N = K_6 W_a = K_3 A_p^{0.5} \tag{2-35}$$

$$P_{cu} = K_6 A_p^{0.5} K_5 A_p^{0.25} J^2 \rho \tag{2-36}$$

$$K_7 = K_6 K_5 \rho \tag{2-37}$$

Assuming the core loss is the same as the copper loss for optimized transformer operation (See Chapter 7).

$$P_{cu} = K_7 A_p^{0.75} J^2 = P_{fe} \tag{2-38}$$

$$P_\Sigma = P_{cu} + P_{fe} \tag{2-39}$$

$$\Delta T = K_8 \frac{P_\Sigma}{A_t} \tag{2-40}$$

$$\Delta T = \frac{2K_8K_7J^2A_p^{0.75}}{K_sA_p^{0.5}} \quad (2\text{-}41)$$

$$K_9 = \frac{2K_8K_7}{K_s} \quad (2\text{-}42)$$

$$\Delta T = K_9J^2A_p^{0.25} \quad (2\text{-}43)$$

$$J^2 = \frac{\Delta T}{K_9A_p^{0.25}} \quad (2\text{-}44)$$

$$K_{10} = \frac{\Delta T}{K_9} \quad (2\text{-}45)$$

$$J^2 = K_{10}A_p^{-0.25} \quad (2\text{-}46)$$

$$J = K_jA_p^{-0.125} \quad (2\text{-}47)$$

The current density/area product relationship*

$$J = K_jA_p^{-0.125}$$

in which K_j is a constant related to core configuration, is shown in Table 2-11, which has been derived by averaging the values in Tables 2-2 through 2-7, columns 9 and 13.

*This is the theoretical value for current density/area product relationship. The empirical values for different core configuration are found in Table 2-1.

Table 2-11. Constant K_j

Core type	$K_j(\Delta 25°)$	$K_j(\Delta 50°)$
Pot core	433	632
Powder core	403	590
Lamination	366	534
C-type core	322	468
Single-coil C-core	395	569
Tape-wound core	250	365

The relationship between current density and area product A_p for a temperature rise of 25°C and 50°C is given in Figures 2-32 through 2-37. It was obtained from the data shown in Tables 2-2 through 2-7, in which the J and A_p values are shown in columns 9 and 13 for current density, and column 3 for area product.

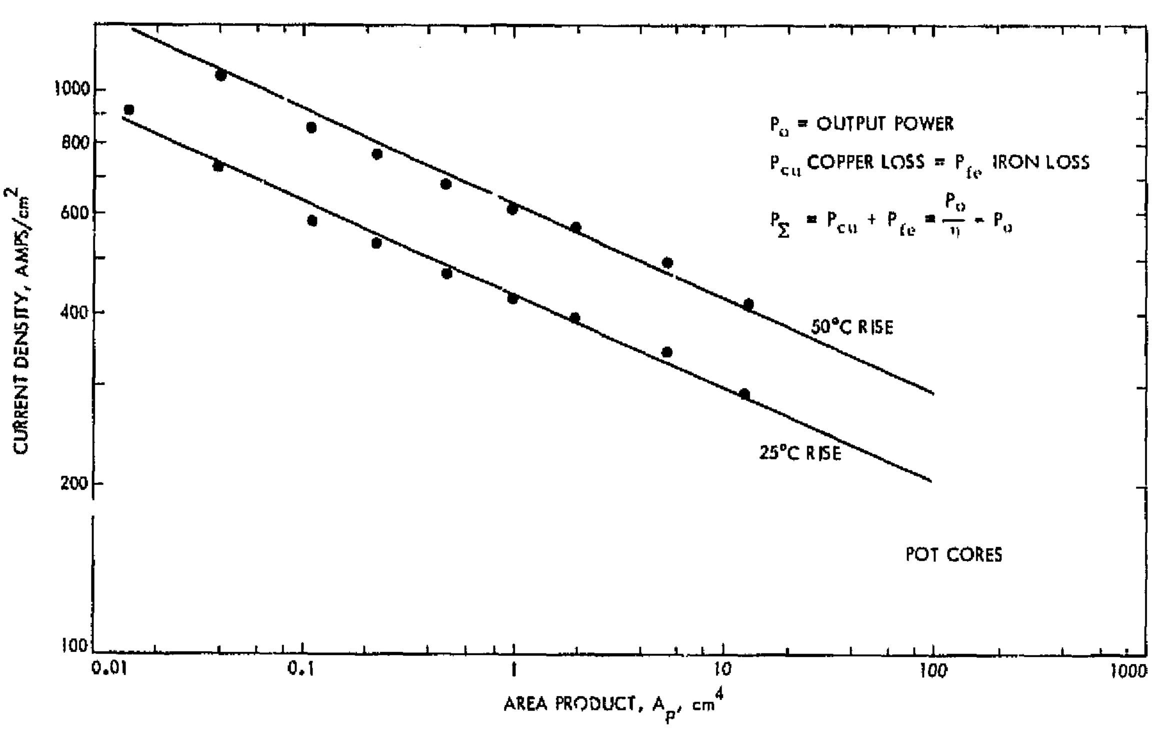

Fig. 2-32. Current density versus area product A_p for a 25°C and 50°C rise for pot cores

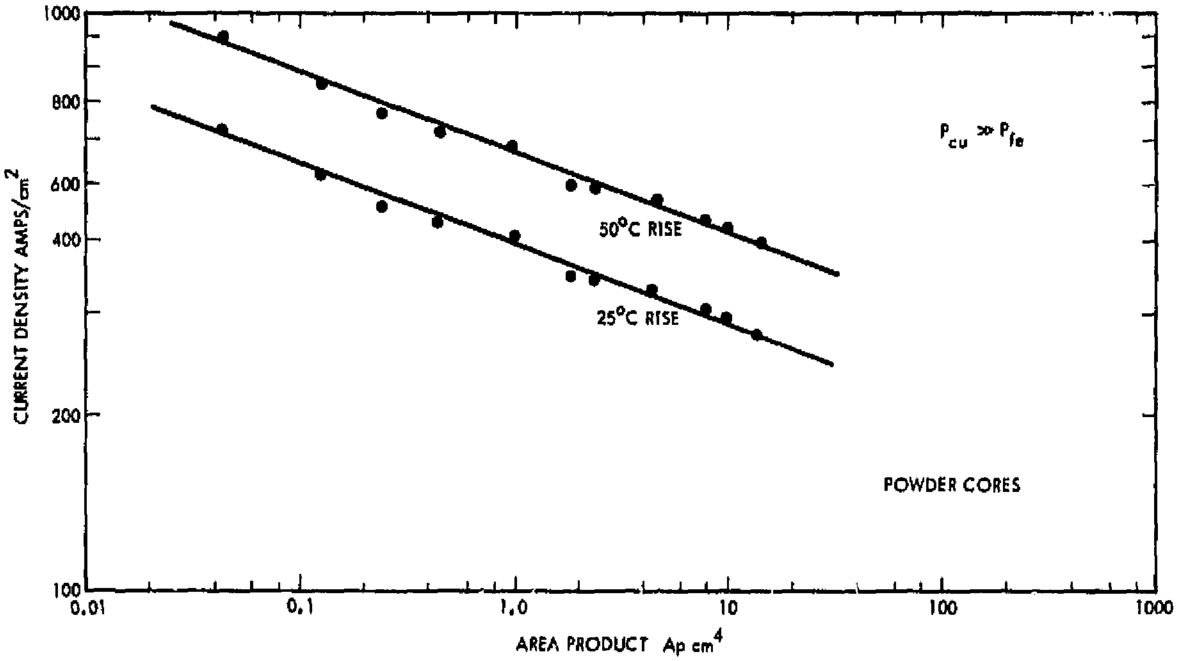

Fig. 2-33. Current density versus area product A_p for a 25°C and 50°C rise for powder cores

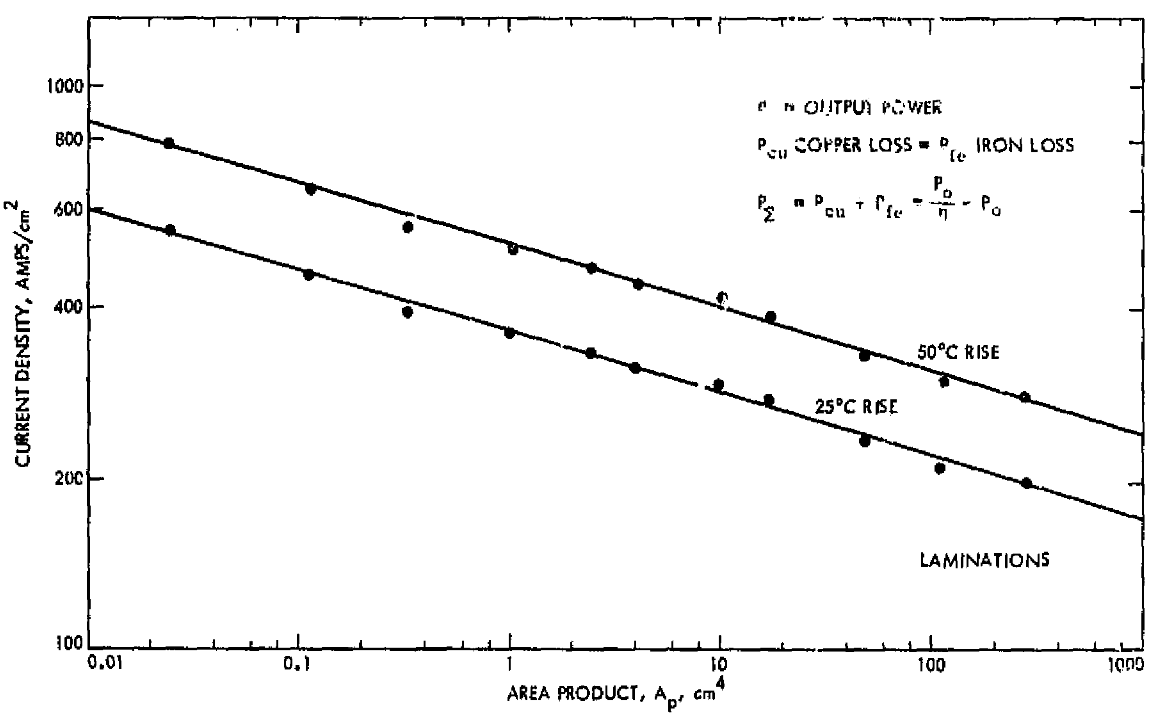

Fig. 2-34. Current density versus area product A_p for 25°C and 50°C rise for laminations

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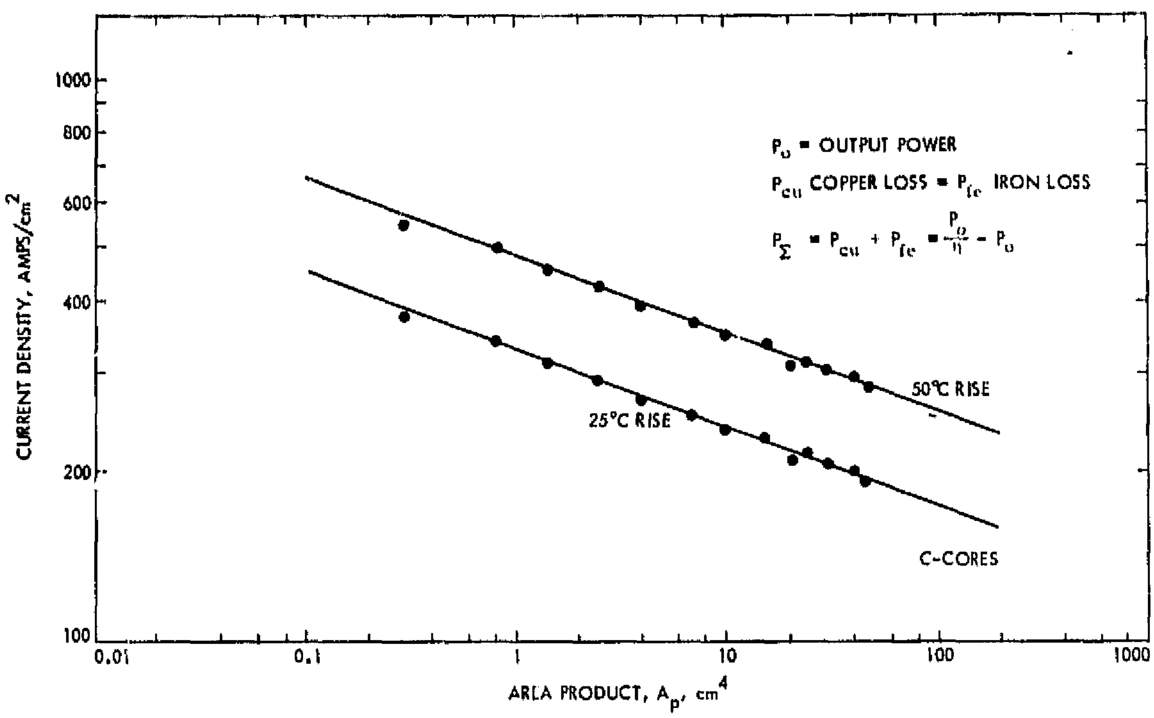

Fig. 2-35. Current density versus area product A_p for 25°C and 50°C rise for C-cores

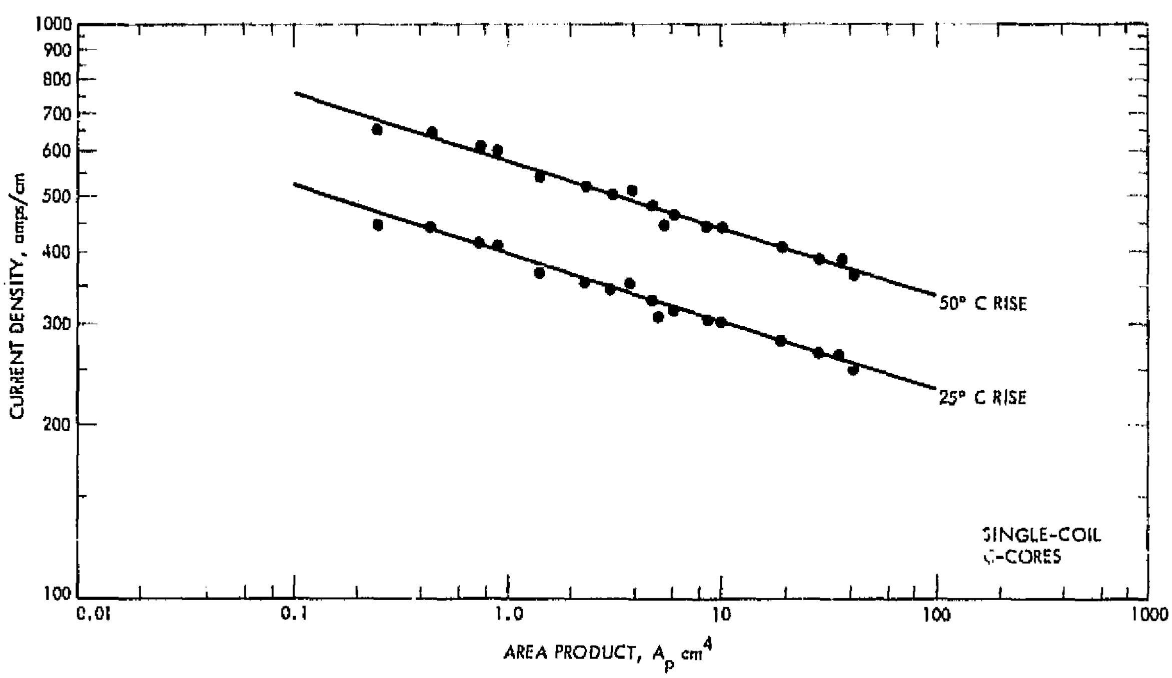

Fig. 2-36. Current density versus area product A for a 25°C and 50°C rise for single-coil C-cores

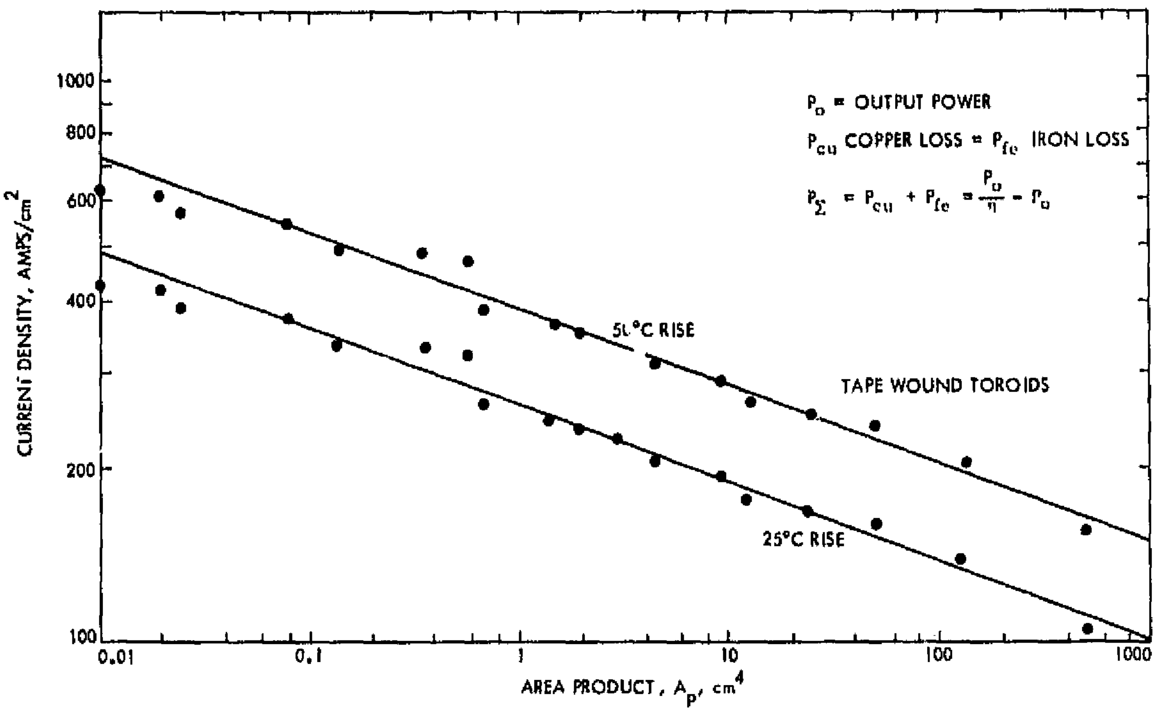

Fig. 2-37. Current density versus area product A_p for 25°C and 50°C rise for tape-wound toroids

CHAPTER III

POWER TRANSFORMER DESIGN

A. INTRODUCTION

The conversion process in power electronics requires the use of transformers, components which frequently are the heaviest and bulkiest item in the conversion circuits. They also have a significant effect upon the overall performance and efficiency of the system. Accordingly, the design of such transformers has an important influence on overall system weight, power conversion efficiency and cost. Because of the interdependence and interaction of parameters, judicious tradeoffs are necessary to achieve design optimization.

B. THE DESIGN PROBLEM GENERALLY

The designer is faced with a set of constraints which must be observed in the design of any transformer. One of these is the output power, P_o, (operating voltage multiplied by maximum current demand) which the secondary winding must be capable of delivering to the load within specified regulation limits. Another relates to minimum efficiency of operation which is dependent upon the maximum power loss which can be allowed in the transformer. Still another defines the maximum permissible temperature rise for the transformer when used in a specified temperature environment.

Other constraints relate to volume occupied by the transformer and particularly in aerospace applications, weight, since weight minimization is an important goal in the design of space flight electronics. Lastly, cost effectiveness is always an important consideration.

Depending upon application, certain of these constraints will dominate. Parameters affecting others may then be traded off as necessary to achieve the most desirable design. It is not possible to optimize all parameters in a single design because of the interaction and interdependence of parameters. For example, if volume and weight are of great significance, reductions in both often can be effected by operating the transformer at a higher frequency but at a penalty in efficiency. When the frequency cannot be raised, reduction in weight and volume may still be possible by selecting a more efficient core

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material, but at a penalty of increased cost. Judicious tradeoffs thus must be effected to achieve the design goals.

A flow chart showing the interrelation and interaction of the various design factors which must be taken into consideration is shown in Figure 3-1.

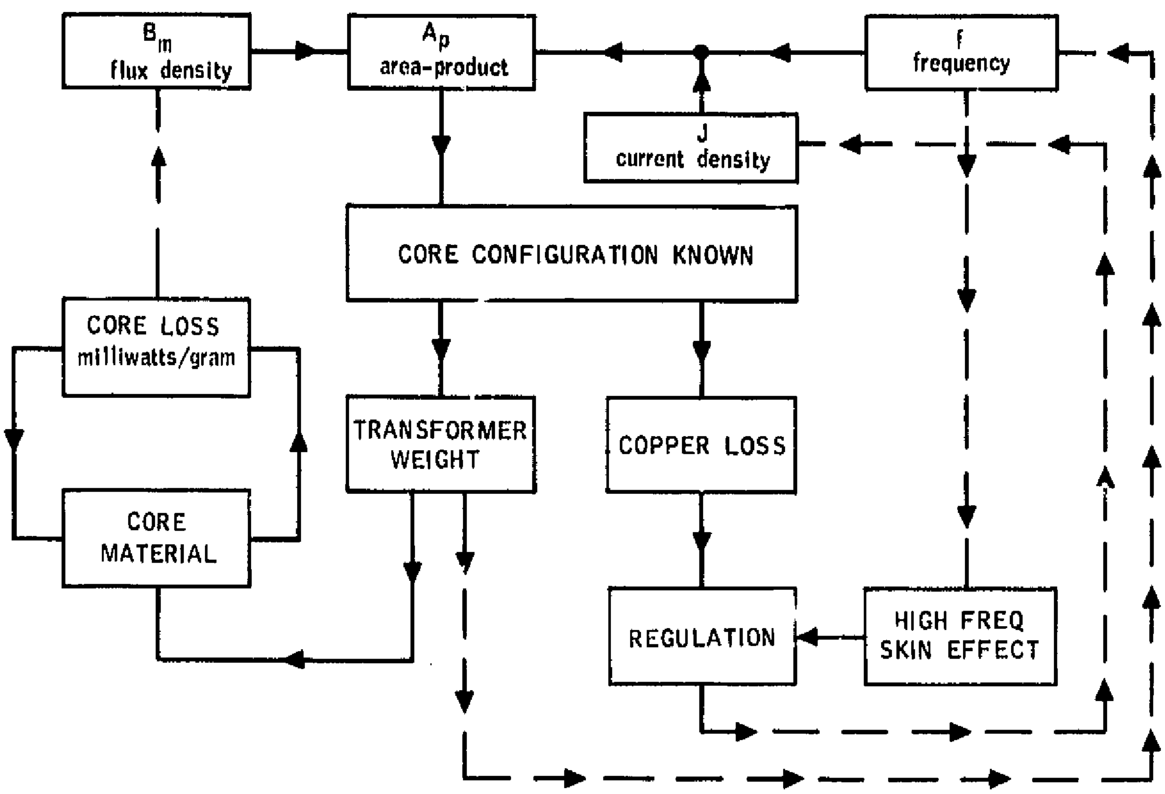

Fig. 3-1. Transformer design factors flow chart

Various transformer designers have used different approaches in arriving at suitable designs. For example, in many cases a rule of thumb is used for dealing with current density. Typically, an assumption is made that a good working level is 1000 circular mils per ampere. This will work in many instances but the wire size needed to meet this requirement may produce a heavier and bulkier transformer than desired or required. The information presented herein makes it possible to avoid the use of this and other rules of thumb and to develop a more economical design with great accuracy.

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C. RELATIONSHIP OF A_p TO TRANSFORMER POWER HANDLING CAPABILITY

According to the newly developed approach, the power handling capability of a core is related to its area product by an equation which may be stated as:

$$A_p = \left(\frac{P_t \times 10^4}{KB_m f K_u K_j}\right)^{1.16} \qquad [\,cm^4\,] \qquad (3\text{-}1)$$

where

K = waveform coefficient
4.0 square wave
4.44 sine wave

B_m = flux density, tesla

f = frequency, Hz

K_u = window utilization factor (see Chapter 6)

K_j = current density coefficient (see Chapter 2)

P_t = apparent power, primary plus secondary

From the above it can be seen that factors such as flux density, frequency of operation, window utilization factor K_u, which defines the maximum space which may be occupied by the copper in the window, and the constant K_j, which is related to temperature rise, all have an influence on the transformer area product. The constant K_j is a new parameter that gives the designer control of the copper loss. Derivation is set forth in detail in Chapter 2. The derivation for area product A_p is set forth in detail at the end of this chapter Appendix 3.A.

D. OUTPUT POWER VS INPUT POWER VS APPARENT POWER CAPABILITY

Output power (P_o) is of greatest interest to the user. To the transformer designer it is the apparent power (P_t) which is associated with the geometry of the transformer that is of greater importance. Assume, for the sake of simplicity, the core of an isolation transformer has but two windings in the window area (W_a), a primary and a secondary. Also assume that the window area (W_a) is divided up in proportion to the power handling capability of the windings using equal current density. The primary winding handles P_{in} and the secondary handles P_o to the load. Since the power transformer has to be designed to accommodate the primary P_{in} and secondary P_o, then:

$$P_t = P_{in} + P_o \tag{3-2}$$

$$P_{in} = \frac{P_o}{\eta} \tag{3-3}$$

$$P_t = \frac{P_o}{\eta} + P_o \tag{3-4}$$

$$P_t = P_o\left(\frac{1}{\eta} + 1\right) \tag{3-5}$$

The designer must be concerned with the apparent power handling capability, P_t, of the transformer core and windings. P_t may vary by a factor ranging from 2 to 2.828 times the input power, P_{in}, depending upon the type of circuit in which the transformer is used. If the current in the rectifier transformer becomes interrupted, its effective RMS value changes. Transformer size, thus, is not only determined by the load demand but, also, by application because of the different copper losses incurred due to current waveform (see Chapter 7, Fig. 7-20).

For example, for a load of one watt, compare the power handling capabilities required for each winding (neglecting transformer and diode losses so that $P_{in} = P_o$) for the full-wave bridge circuit of Fig. 3-2, the full-wave center-tapped secondary circuit of Fig. 3-3, and the push-pull center-tapped full-wave circuit in Fig. 3-4, where all windings have the same number of turns (N).

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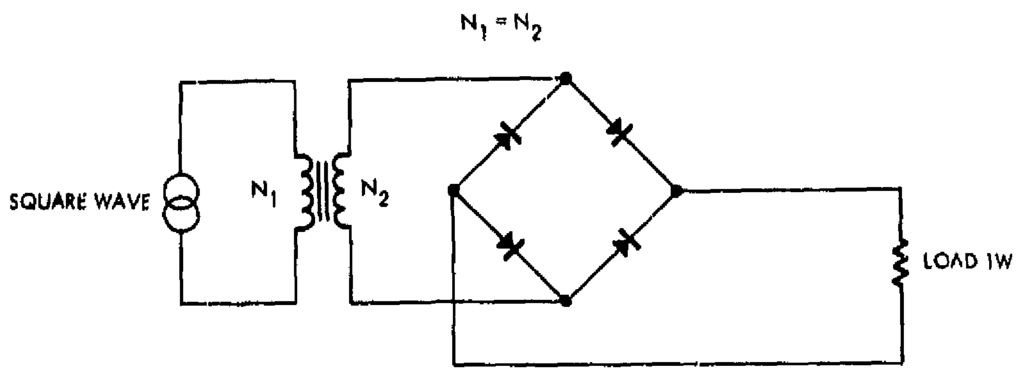

Fig. 3-2. Full-wave bridge circuit

The total apparent power P_t for the circuit shown in Fig. 3-2 is 2 watts. This is shown in the following equation:

$$P_t = \overbrace{(I_{N1}\ E_{N1})}^{P_{in}} + \overbrace{(I_{N2}\ E_{N2})}^{P_o} \tag{3-6}$$

$$P_t = 2\ P_{in} \tag{3-7}$$

in which I_{N1} and I_{N2} are the currents associated with the primary and secondary windings, respectively, and E_{N1} and E_{N2} are the voltages across the primary and secondary windings, respectively.

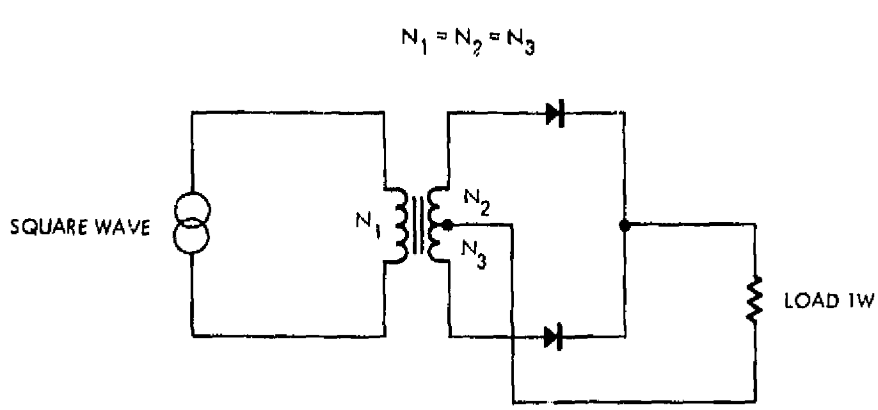

Fig. 3-3. Full-wave, center-tapped circuit

The total power P_t for the circuit shown in Fig. 3-3 increased 20.7% due to the distorted wave form of the interrupted current flowing in the secondary winding. This is shown in the following equation:

$$P_t = (I_{N1} E_{N1}) + \left[(0.707 I_{N2} E_{N2}) + (0.707 I_{N3} E_{N3})\right] \qquad (3\text{-}8)$$

$$P_t = P_{in} + 0.707 P_{in} + 0.707 P_{in} = 2.414 P_{in} \qquad (3\text{-}9)$$

Rewriting equation 3-5 to incorporate the RMS rating,

$$P_t = P_o \left(\frac{1}{\eta} + \sqrt{2}\right) \qquad (3\text{-}10)$$

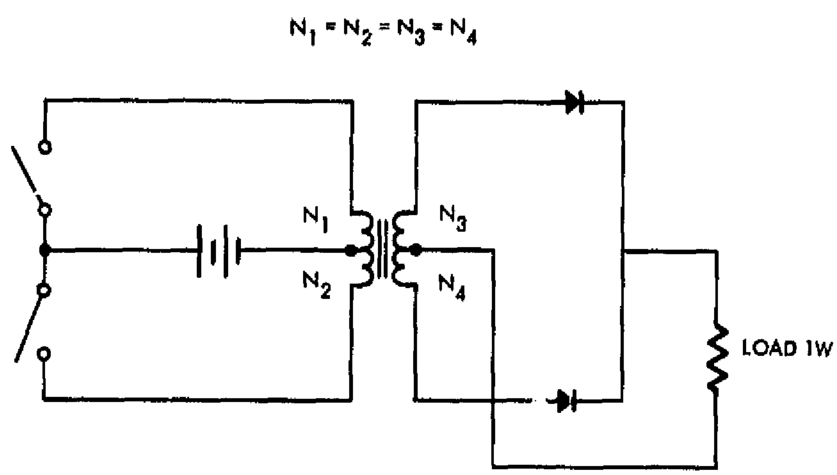

Fig. 3-4. Push-pull, full-wave, center-tapped circuit

The total power P_t for the circuit shown in Figure 3-4, which is typical of a dc to dc converter, increases to 2.828 times P_{in} because of the interrupted current flowing in both the primary and secondary windings since

$$N_1 = N_2 = N_3 = N_4 ,$$

$$P_t = \left[(0.707 I_{N1} E_{N1}) + (0.707 I_{N2} E_{N2})\right] + \left[(0.707 I_{N3} E_{N3}) + (0.707 I_{N4} E_{N4})\right] \quad (3\text{-}11)$$

$$P_t = 0.707 P_{in} + 0.707 P_{in} + 0.707 P_{in} + 0.707 P_{in} = 2.828 P_{in} \quad (3\text{-}12)$$

Again,

$$P_t = P_o \left(\frac{\sqrt{2}}{\eta} + \sqrt{2}\right) \quad (3\text{-}13)$$

Thus the circuit configuration in which the transformer is to be used must be considered by the designer when sizing the transformer.

Rather than discuss the various methods used by transformer designers, the author believes it will be more useful to consider typical design problems and to work out solutions using the approach based upon the newly formulated relationships.

E. A 2.5-kHz TRANSFORMER DESIGN PROBL... AS AN EXAMPLE

Assume a specification for a transformer design as shown in Fig. 3-2, requiring the following:

(1) E_o, 10 volts

(2) I_o, 2.0 amperes

(3) E_{in}, 50 volts

(4) f, 2500 Hz (square wave)

(5) Maximum temperature rise, 25°C

(6) Transformer efficiency, 95%

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Assuming the bridge rectifier of Fig. 3-2 and using the efficiency const... int of 95%:

Definitions for Table 3-1

Information given is listed by column as:

1. Manufacturer part number
2. Surface area calculated from Chapter 2, Fig. 2-24
3. Area product effective iron area times window area
4. Mean length turn on one bobbin
5. Total number of turns and wire size for two bobbins using a window utilization factor $K_u = 0.40$
6. Resistance of the wire at 50°C
7. Watts loss is based on Fig. 7-2 for a ΔT of 25°C with a room ambient of 25°C surface dissipation times the transformer surface area, total loss is equal to $2\ P_{cu}$
8. Current calculated from column 6 and 7
9. Current density calculated from column 5 and 8
10. Resistance of the wire at 75°C
11. Watts loss is based on Fig. 7-2 for a ΔT of 50°C with a room ambient of 25°C surface dissipation times the transformer surface area, total loss is equal to $2\ P_{cu}$
12. Current calculated from column 10 and 11
13. Current density calculated from column 5 and 12
14. Effective core weight in grams
15. Copper weight in grams
16. Transformer volume calculated from Chapter 2, Fig. 2-8
17. Core effective cross-section

Table 3-1. C-core characteristics

	1	2	3	4	5	6	7	8	9	10	11	12	13	14	15	16	17
	Core	A_t cm^2	A_p cm^4	MLT cm	N / AWG	Ω @ 50° C	$P_Σ$	$I = \sqrt{\frac{W}{Ω}}$	ΔT 25°C $J = \frac{amps}{cm^2}$	Ω @ 75° C	$P_Σ$	$I = \sqrt{\frac{W}{Ω}}$	ΔT 50°C $J = \frac{amps}{cm^2}$	Core Wt grams	Cu Wt grams	Volume cm^3	A_c cm^2
1	AL-2	20.9	0.265	3.55	662 30	8.93	0.627	0.187	470	9.81	1.46	0.273	538	12.23	11.1	7.14	0.265
2	AL-3	23.9	0.410	4.18	662 30	10.5	0.717	0.185	365	11.5	1.67	0.259	531	18.12	13.06	8.92	0.410
3	AL-5	33.6	0.767	4.59	946 30	16.5	1.01	0.174	345	18.1	2.35	0.255	503	31.3	20.50	14.06	0.539
4	AL-6	37.5	1.011	5.23	946 30	18.8	1.15	0.172	341	20.6	2.63	0.254	490	41.7	23.40	16.88	0.716
5	AL-124	45.3	1.44	5.50	1317 30	27.5	1.36	0.157	310	30.2	3.17	0.229	452	46.6	34.20	22.50	0.716
6	AL-8	63.4	2.31	5.74	221 20	0.482	1.90	1.404	271	0.529	4.44	2.05	395	67.9	59.95	35.66	0.806
7	AL-9	69.0	3.09	6.38	221 20	0.535	2.07	1.39	268	0.587	4.83	2.03	351	89.2	[illegible]	41.62	1.077
8	L-10	74.5	3.85	7.01	221 20	0.588	2.24	1.38	266	0.646	5.22	2.01	387	110	73.2	47.55	1.342
9	AL-12	87.0	4.57	7.09	275 20	0.748	2.61	1.32	255	0.821	6.09	1.93	371	111	93.2	61.38	1.26
10	AL-135	93.7	5.14	7.36	325 20	0.908	2.81	1.24	[illegible]	0.997	6.56	1.81	345	114	113	69.63	1.26
11	AL-78	98.1	6.07	7.01	312 20	0.831	2.94	1.33	25[illegible]	0.912	6.87	1.94	374	155	103	62.83	1.34
12	AL-18	118	7.92	7.61	510 20	1.47	3.55	1.10	211	1.61	8.26	1.60	308	138	183	94.79	1.25
13	AL-15	120	9.07	8.05	386 20	1.18	3.56	1.23	237	1.30	8.40	1.79	346	205	147	94.43	1.80
14	AL-16	127	10.8	8.89	386 20	1.30	3.80	1.20	233	1.43	8.89	1.76	340	235	162	104.95	2.15
15	AL-17	142	14.4	10.3	386 20	1.51	4.25	1.185	228	1.66	9.94	1.73	333	314	188	124.94	2.87
16	AL-19	159	18.0	10.8	511 20	2.10	4.77	1.065	205	2.31	11.1	1.55	299	328	261	155.44	2.87
17	AL-20	182	22.6	11.5	511 20	2.23	5.46	1.106	213	2.45	12.7	1.61	310	437	278	187.08	3.58
18	AL-22	202	28.0	11.5	637 20	2.78	6.05	1.043	201	3.05	14.1	1.52	293	489	346	212.04	3.58
19	AL-23	220	34.9	12.7	637 20	3.07	6.60	1.036	200	3.37	15.4	1.51	291	612	382	244.67	4.48
20	AL-24	245	40.0	12.0	948 20	4.32	7.35	0.922	178	4.74	17.1	1.35	259	552	538	280.91	3.58

copper loss = iron loss

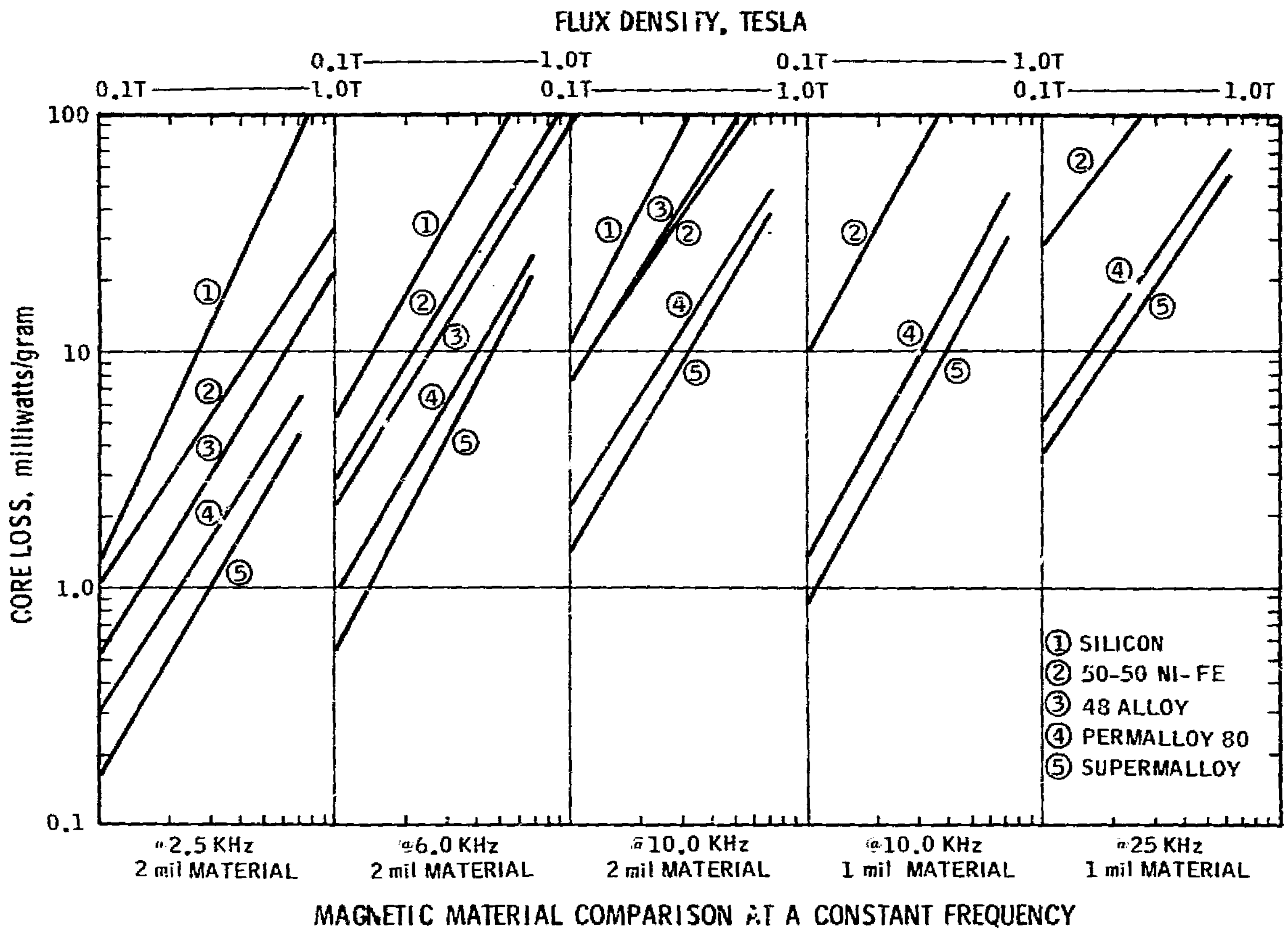

Fig. 3-5. Magnetic material comparison at a constant frequency

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Step No. 1. Calculate the apparent power P_t from equation 3-5, allowing for 1.0 volt diode drop (V_d) assumed:

$$P_t = P_o \left(\frac{1}{\eta} + 1\right)$$

$$P_t = I_o (E_o + V_d) \times \left(\frac{1}{\eta} + 1\right)$$

$$P_t = 2 (10 + 2) \times \left(\frac{1}{0.95} + 1\right)$$

$$P_t = 49.3 \quad [\text{watts}]$$

Step No. 2. Calculate the area product A_p from equation 3-1:

$$A_p = \left(\frac{P_t \times 10^4}{K B_m f K_u K_j}\right)^{1.16} \quad [\text{cm}^4]$$

Assuming

$$K = 4.0$$

$$B_m = 0.3 \quad [\text{tesla}]$$

$$K_u = 0.4 \text{ (Chapter 6)}$$

$$K_j = 323 \text{ (Chapter 2)}$$

$$A_p = \left(\frac{(49.3) \times 10^4}{(4.0)(0.3)(2500)(0.4)(323)}\right)^{1.16}$$

or

$$A_p = 1.32 \quad [\text{cm}^4]$$

After the A_p has been determined, the geometry of the transformer can be evaluated as described in Chapter 2 for weight, for surface area, and for volume, and appropriate changes made, if required. Having established the configuration, it is then necessary to determine the core material to complete core selection.

<u>Step No. 3.</u> Select a C-core from Table 3-1 with a value of A_p closest to the one calculated.

$$\text{AL-124 with an } A_p = 1.44 \qquad [\text{cm}^4]$$

<u>Step No. 4.</u> Calculate the total transformer losses P_Σ:

$$P_\Sigma = \frac{P_o}{\eta} - P_o \qquad [\text{watts}]$$

$$P_\Sigma = \frac{24}{0.95} - 24$$

$$P_\Sigma = 1.26 \qquad [\text{watts}]$$

Maximum efficiency is realized when the copper (winding) losses are equal to the iron (core) losses (see Chapter 7):

$$P_{cu} = P_{fe}$$

and therefore

$$P_{cu} = \frac{P_\Sigma}{2}$$

and thus

$$P_{cu} = \frac{1.26}{2}$$

$$P_{cu} = 0.63 = P_{fe}$$

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Step No. 5. Select the core weight from Table 3-1, column 14, then calculate the core loss in milliwatts per gram:

$$\text{AL-124} \quad W_t = 46.6 \text{ grams}$$

$$\frac{P_{fe}}{W_t} \times 10^3 = \text{milliwatts/g}$$

$$\frac{0.63}{46.6} \times 10^3 = \text{milliwatts/g}$$

13.5 milliwatts/g

Step No. 6. Select the proper magnetic material in Fig. 3-5, reading from the 2.5 kHz frequency curve for a flux density of 0.3 tesla. The magnetic material that comes closest to 13.5 milliwatts per gram is silicon steel, with approximately 12 milliwatts per gram. With a weight of 46.6 grams, the total core loss is 560 milliwatts, which meets the requirement of the design.

Step No. 7. Calculate the number of primary turns using Faraday's law, equation 3.A-1.*

$$N_p = \frac{E_p \times 10^4}{4 B_m A_c f}$$

The iron cross section A_c is found in Table 3-1, column 17:

$$A_c = 0.716 \qquad [cm^2]$$

Thus

$$N_p = \frac{(50) \times 10^4}{(4)(0.3)(0.716)(2500)}$$

*See Appendix 3.A, at the end of Chapter 3.

or

$$N_p = 233 \text{ turns (primary)}$$

Step No. 8. Calculate the current density J from ecuation 3.A-17:

$$J = K_j A_p^{-0.14}$$

(The value for K_j is found in Table 2-1.)

$$J = (323)(1.32)^{-0.14}$$

$$J = 307 \qquad [A/cm^2]$$

Step No. 9. Calculate the primary current I_p and wire size A_w:

$$I_p = \frac{P_t}{2\,E_p} \qquad [A]$$

$$I_p = \frac{(49.3)}{(2)(50)}$$

$$I_p = 0.493 \qquad [A]$$

The bare wire size $A_{w(B)}$ for the primary is

$$A_{w(B)} = \frac{I_p}{J}$$

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$$A_{w(B)} = \frac{0.493}{307}$$

$$A_{w(B)} = 0.001606 \qquad [cm^2]$$

Step No. 10. Select the wire area A_w in Table 6-1 for equivalent (AWG) wire size, column A.

$$\text{AWG No. } 25 = 0.001623 \qquad \left[cm^2\right]$$

The rule is that when the calculated wire size does not fall close to those listed in the table, the next smaller size should be selected.

Step No. 11. Calculate the resistance of the primary winding, using Table 6-1, column C, and Table 3-1, column 4, for the MLT:

$$R_p = MLT \times N \times (\text{column C}) \times \zeta \times 10^{-6} \qquad [\Omega]$$

$$R_p = (5.5)(233)(1062)(1.098) \times 10^{-6}$$

$$R_p = 1.49 \qquad [\Omega]$$

Step No. 12. Calculate the primary copper loss P_{cu}:

$$P_{cu} = I_p^2 R_p \qquad [\text{watts}]$$

$$P_{cu} = (0.493)^2 (1.49)$$

$$P_{cu} = 0.362 \qquad [\text{watts}]$$

Step No. 13. Calculate the secondary turns:

$$N_s = \frac{N_p}{E_p}(E_s)$$

$$E_s = 10 + 2\,V_d$$

$$N_s = \frac{(233)}{(50)}(12)$$

$$N_s = 56$$

<u>Step No. 14.</u> Calculate the wire size $A_{w(B)}$ for the secondary winding:

$$A_{w(B)} = \frac{I_s}{J}$$

$$A_{w(B)} = \frac{(2)}{(307)}$$

$$A_{w(B)} = 0.00651 \quad [cm^2]$$

<u>Step No. 15.</u> Select the wire area A_w in Table 6-1 for equivalent (AWG) wire size, column A:

$$\text{AWG No. } 19 = 0.00653 \quad [cm^2]$$

The rule is that when the calculated wire size does not fall close to those listed in the table, the next smaller size should be selected.

<u>Step No. 16.</u> Calculate the resistance of the secondary winding, using Table 6-1, column C, and Table 3-1, column 4, for the MLT.

$$R_s = MLT \times N \times (\text{column C}) \times \zeta \times 10^{-6} \quad [\Omega]$$

$$R_s = (5.5)(56)(264)(1.098) \times 10^{-6}$$

$$R_s = 0.0893 \quad [\Omega]$$

<u>Step No. 17.</u> Calculate the secondary copper loss P_{cu}:

$$P_{cu} = I_s^2 R_s$$

$$P_{cu} = (2.0)^2 (0.0813)$$

$$P_{cu} = 0.357 \quad [\text{watts}]$$

Step No. 18. Summarize the losses and compare with the total losses P_Σ:

Primary	P_{cu}	$= 0.362$	[watts]
Secondary	P_{cu}	$= 0.357$	[watts]
Core	P_{fe}	$= 0.560$	[watts]
Total	P_Σ	$= 1.279$	[watts]

The total power loss in the transformer is 1.279 watts, which will effectively meet the required 95% efficiency.

From Chapter 7, the surface area A_t required to dissipate waste heat (expressed as watts loss per unit area) is

$$A_t = \frac{P_\Sigma}{\psi}$$

where

$\psi = 0.03 \text{ W/cm}^2$ at 25°C rise

Referring to Table 3-1, column 1, for the AL-124 size core, the surface area A_t is 45.3 cm^2:

$$\psi = \frac{P_\Sigma}{A_t}$$

and thus

$$\psi = \frac{1.279}{45.3}$$

or

$$\psi = 0.0282 \qquad [W/cm^2]$$

which will produce the required temperature rise.

F. A 10-kHz TRANSFORMER DESIGN PROBLEM AS AN EXAMPLE

Assume a specification for a transformer design, as shown in Fig. 3-3, requiring the following:

(1) E_o, 56 volts

(2) I_o, 1.79 amperes

(3) E_{in}, 200 volts

(4) f, 10 kHz (square wave)

(5) Maximum temperature rise, 25°C

(6) Transformer efficiency, 98%

assuming the full-wave, center-taped rectifier of Fig. 3-3 and using the efficiency constraint of 98%.

Step No. 1. Calculate the apparent power P_t from equation 3-10, allowing for 1.0 volt diode drop (V_d) assumed:

$$P_t = P_o \left(\frac{1}{\eta} + \sqrt{2}\right)$$

$$P_t = I_o (E_o + V_d) \times \left(\frac{1}{\eta} + \sqrt{2}\right)$$

$$P_t = 1.79 (56 + 1) \times \left(\frac{1}{0.98} + 1.41\right)$$

$$P_t = 248 \qquad [\text{watts}]$$

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Step No. 2. Calculate the area product A_p from equation 3-1:

$$A_p = \left(\frac{P_t \times 10^4}{K B_m f K_u K_j}\right)^{1.16} \quad [cm^4]$$

assuming

$K = 4.0$

$B_m = 0.3$ [tesla]

$K_u = 0.4$ (Chapter 6)

$K_j = 323$ (Chapter 2)

$$A_p = \left(\frac{248 \times 10^4}{(4.0)(0.3)(10^4)(0.4)(323)}\right)^{1.16}$$

or

$$A_p = 1.72 \quad [cm^4]$$

after the A_p has been determined, the geometry of the transformer can be evaluated as described in Chapter 2 for weight, for surface area, and for volume, and appropriate changes made, if required. Having established the configuration, it is then necessary to determine the core material to complete core selection.

Step No. 3. Select a C-core from Table 3-1 with a value of A_p closest to the one calculated:

AL-8 with an $A_p = 2.31$ [cm^4]

Step No. 4. Calculate the total transformer losses P_Σ:

$$P_\Sigma = \frac{P_o}{\eta} - P_o \qquad \text{[watts]}$$

$$P_\Sigma = \left(\frac{102}{0.98}\right) - (102)$$

$$P_\Sigma = 2.08 \qquad \text{[watts]}$$

Maximum efficiency is realized when the copper (winding) losses are equal to the iron (core) losses (see Chapter 7) which is expressed as

$$P_{cu} = P_{fe}$$

and therefore

$$P_{cu} = \frac{P_\Sigma}{2}$$

and thus

$$P_{cu} = \frac{2.08}{2}$$

$$P_{cu} = 1.04 \qquad \text{[watts]}$$

Step No. 5. Select the core weight from Table 3-1, Column 14, then calculate the core loss in milliwatts per gram:

$$\text{AL-8} \quad W_t = 66.6 \text{ grams}$$

$$\frac{P_{fe}}{W_t} \times 10^3 = \text{milliwatts/g}$$

$$\frac{1.04}{66.6} \times 10^3 = \text{milliwatts/g}$$

15.6 milliwatts/g

Step No. 6. Select the proper magnetic material in Fig. 3-5, reading from the 10-kHz frequency curve with a density of 0.3 tesla. The magnetic material that comes closest to 15.6 milliwatts per gram is Permalloy 80, with approximately 12 milliwatts per gram. When nickel steel is used, Table 7-1 provides a weight correction factor.

The weight from Table 3-1 is multiplied by the weight correction factor:

$$66.6 \times 1.144 = 76.2 \qquad [\text{grams}]$$

With a weight of 76.2 grams the total core loss is

$$12 \times 76.2 \times 10^{-3} = 0.914 \qquad [\text{watts}]$$

Step No. 7. Calculate the number of primary turns using Faraday's law, equation 3.A-1:

$$N_p = \frac{E_p \times 10^4}{4 B_m A_c f}$$

The iron cross section A_c is found in Table 3-1, column 17:

$$A_c = 0.806 \qquad [\text{cm}^2]$$

$$N_p = \frac{(200) \times 10^4}{(4)(0.3)(0.806)(10^4)}$$

$$N_p = 207 \text{ turns (primary)}$$

Step No. 8. Calculate the current density J from equation 3. A-17:

$$J = K_j A^{-0.14}$$

The value for K_j is found in Table 2-1:

$$J = (323)(2.31)^{-0.14}$$

$$J = 287 \qquad [A/cm^2]$$

Step No. 9. Calculate the primary current I_p and wire size A_w:

$$I_p = \frac{I_o (E_o + V_d)}{E_p \eta} \qquad [A]$$

$$I_p = \frac{1.79 (56 + 1)}{(200)(0.98)}$$

$$I_p = 0.520 \qquad [A]$$

The bare wire size for the primary is

$$A_{w(B)} = \frac{I_p}{J} \qquad [cm^2]$$

$$A_{w(B)} = \frac{0.520}{287}$$

$$A_{w(B)} = 0.00181 \qquad [cm^2]$$

Step No. 10. Select the wire area $A_{w(B)}$ in Table 6-1 for equivalent (AWG) wire size, column A:

$$\text{AWG No. } 25 = 0.001623 \qquad [cm^2]$$

The rule is that when the calculated wire size does not fall close to those listed in the table, the next smaller size should be selected.

Step No. 11. Calculate the resistance of tne primary winding, using Table 6-1, column C, and Table 3-1, column 4, for the MLT:

$$R_p = MLT \times N \times (\text{column C}) \times \zeta \times 10^{-6} \quad [\Omega]$$

$$R_p = (5.74)(207)(1062)(1.098) \times 10^{-6}$$

$$R_p = 1.38 \quad [\Omega]$$

Step No. 12. Calculate the primary copper loss P_{cu}:

$$P_{cu} = I_p^2 R_p \quad [\text{watts}]$$

$$P_{cu} = (0.520)^2 (1.38)$$

$$P_{cu} = 0.373 \quad [\text{watts}]$$

Step No. 13. Calculate the secondary turns:

$$N_s = \frac{N_p}{E_p}(E_s)$$

$$E_s = 56 + 1\ V_d$$

$$N_s = \frac{(207)}{(200)}(57)$$

$$N_s = 59 \text{ turns secondary}$$

Step No. 14. Calculate the wire size $A_{w(B)}$ for the secondary winding (see equation 3-8):

$$A_{w(B)} = \frac{I_o\,(0.707)}{J} \qquad [cm^2]$$

$$A_{w(B)} = \frac{1.79\,(0.707)}{287}$$

$$A_{w(B)} = 0.0044 \qquad [cm^2]$$

Step No. 15. Select the bare wire area $A_{w(B)}$ in Table 6-1 for equivalent (AWG) wire size, column A:

$$\text{AWG No. } 21 = 0.00411 \qquad [cm^2]$$

The rule is that when the calculated wire size does not fall close to those listed in the table, the next smaller size should be selected.

Step No. 16. Calculate the resistance of the secondary winding, using Table 6-1, column C, and Table 3-1, column 4, for the MLT:

$$R_s = MLT \times N \times (\text{column C}) \times \zeta \times 10^{-6}$$

$$R_s = (5.74)(59)(419)(1.098) \times 10^{-6}$$

$$R_s = 0.156 \qquad [\Omega]$$

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Step No. 17. Calculate the total secondary copper loss P_{cu}, N_2 plus N_3 (see Fig. 3-3):

$$P_{cu} = (I_o \times 0.707)^2 R_s + (I_o \times 0.707)^2 R_s \qquad [\text{watts}]$$

$$P_{cu} = 2\,(1.79 \times 0.707)^2\, 0.156$$

$$P_{cu} = 0.499 \qquad \text{[watts]}$$

Step No. 18. Summarize the losses and compare with the total losses P_Σ:

Primary	P_{cu}	= 0.373	[watts]
Secondary	P_{cu}	= 0.499	[watts]
Core	P_{fc}	= 1.07	[watts]
Total	P_Σ	= 1.942	[watts]

The total power loss in the transformer is 1.942 watts, which will meet the required 98% efficiency.

From Chapter 7, the surface area A_t required to dissipate waste heat (expressed as watts loss per unit area) is

$$A_t = \frac{P_\Sigma}{\psi}$$

where

$$\psi = 0.03\ \text{W/cm}^2 \text{ at } 25°\text{C rise}$$

Referring to Table 3-1, column 1, for the AL-8 size core, the surface area A_t is 63.4 cm^2:

$$\psi = \frac{P_\Sigma}{A_t}$$

$$\psi = \frac{1.942}{63.4}$$

$$\psi = 0.0306 \quad [W/cm^2]$$

which will produce the required temperature rise.

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1. McLyman, C., Design Parameters of Toroidal and Bobbin Magnetics, Technical Memorandum 33-651, pages 12-15, Jet Propulsion Laboratory, Pasadena, Calif.

2. Blume, L.F., Transformer Engineering, John Wiley & Sons Inc., New York, N.Y. 1938. Pages 272-282.

3. Terman, F.E., Radio Engineers Handbook, McGraw-Hill Book Co., Inc., New York 1943. Pages 28-37.

APPENDIX 3.A

TRANSFORMER POWER HANDLING CAPABILITY

The power handling capability of a transformer can be related to A_p a quantity (which is the W_aA_c product where W_a is the available core window area in cm^2 and A_c is the effective cross-sectional area of the core in cm^2), as follows.

A form of the Faraday law of electromagnetic induction much used by transformer designers states:

$$E = K B_m A_c N f \times 10^{-4} \qquad (3.A-1)$$

(The constant K is taken at 4 for square wave and at 4.44 for sine wave operation.)

It is convenient to restate this expression as:

$$N A_c = \frac{E \times 10^4}{4 B_m f} \qquad (3.A-2)$$

for the following manipulation.

By definition the window utilization factor is:

$$K_u = \frac{N A_w}{W_a} \qquad (3.A-3)$$

and this may be restated as:

$$N = \frac{K_u W_a}{A_w} \qquad (3.A-4)$$

If both sides of the equation are multiplied by A_c, then:

$$N A_c = \frac{K_u W_a A_c}{A_w} \quad (3.A\text{-}5)$$

From equation 3.A-2:

$$\frac{K_u W_a A_c}{A_w} = \frac{E \times 10^4}{4 B_m f} \quad (3.A\text{-}6)$$

Solving for $W_a A_c$:

$$W_a A_c = \frac{E A_w \times 10^4}{4 B_m f K_u} \quad (3.A\text{-}7)$$

By definition, current density $J = \text{amp/cm}^2$ which may also be stated:

$$J = \frac{I}{A_w} \quad (3.A\text{-}8)$$

which may also be stated as:

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$$A_w = \frac{I}{J} \quad (3.A\text{-}9)$$

It will be remembered that transformer efficiency is defined as:

$$\eta = \frac{P_o}{P_{in}} \text{ and } P_{in} = E I \quad (3.A\text{-}10$$

Rewriting equation 3.A-7 as:

$$EA_w = 4B_m fK_u W_a A_c 10^{-4} = \frac{EI}{J} \quad (3.A\text{-}11$$

and since:

$$\frac{EI}{J} = \frac{P_{in}}{J} = \frac{P_o}{J\eta} \qquad (3.A-12)$$

then:

$$W_aA_c\Big|_{total} = W_aA_c\Big|_{Primary} + W_aA_c\Big|_{Secondary}$$

$$W_aA_c\ {}_{total} = \frac{P_o \times 10^4}{\eta J\ 4B_m f K_u} + \frac{P_o \times 10^4}{4B_m f K_u J} = \frac{P_o \times 10^4}{4B_m f K_u J}(1/\eta + 1) \qquad (3.A-13)$$

and since

$$P_t = \frac{P_o}{\eta} + P_o \qquad (3.A-14)$$

then

$$W_aA_c = \frac{P_t \times 10^4}{4B_m f K_u J} \qquad (3.A-15)$$

$$A_p = \frac{P_t \times 10^4}{4\,B_m\,f\,J\,K_u} \qquad (3.A-16)$$

Combining the equation from Table 2-1,

$$J = K_j\,A_p^{-0.14} \qquad (3.A-17)$$

yielding

$$A_p = \frac{P_t \times 10^4}{4\, B_m f K_u \left(K_j A_p^{-0.14}\right)} \qquad (3.A\text{-}18)$$

$$A_p^{0.86} = \frac{P_t \times 10^4}{4\, B_m f K_u K_j} \qquad (3.A\text{-}19)$$

$$A_p = \left(\frac{P_t \times 10^4}{4\, B_m f K_u K_j}\right)^{1.16} \qquad \left[cm^4\right] \qquad (3.A\text{-}20)$$

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CHAPTER IV

SIMPLIFIED CUT CORE INDUCTOR DESIGN

A. INTRODUCTION

Designers have used various approaches in arriving at suitable inductor designs. For example, in many cases a rule of thumb used for dealing with current density is that a good working level is 1000 circular mils per ampere. This is satisfactory in many instances; however, the wire size used to meet this requirement may produce a heavier and bulkier inductor than desired or required. The information presented herein will make it possible to avoid the use of this and other rules of thumb and to develop a more economical and a better design.

B. CORE MATERIAL

Designers have routinely tended to specify moly permalloy powder core materials for filter inductors used in high frequency power converters and pulse-width modulated (PWM) switched regulators because of the availability of manufacturers' literature containing tables, graphs and examples which simplify the design task. Use of these cores may not result in an inductor design optimized for size and weight. For example as shown in Figure 4-1, moly permalloy powder cores operating with a dc bias of 0.3 tesla have only about 80% of original inductance with very rapid falloff at higher densities. In contrast, the steel core has approximately four times the useful flux density capability while retaining 90% of the original inductance at 1.2 tesla.

There are significant advantages to be gained by the use of C cores and cut toroids fabricated from grain-oriented silicon steel, despite such disadvantages as the need for banding and gapping materials, banding tools, mounting brackets and winding mandrels.

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*See Reference 1.

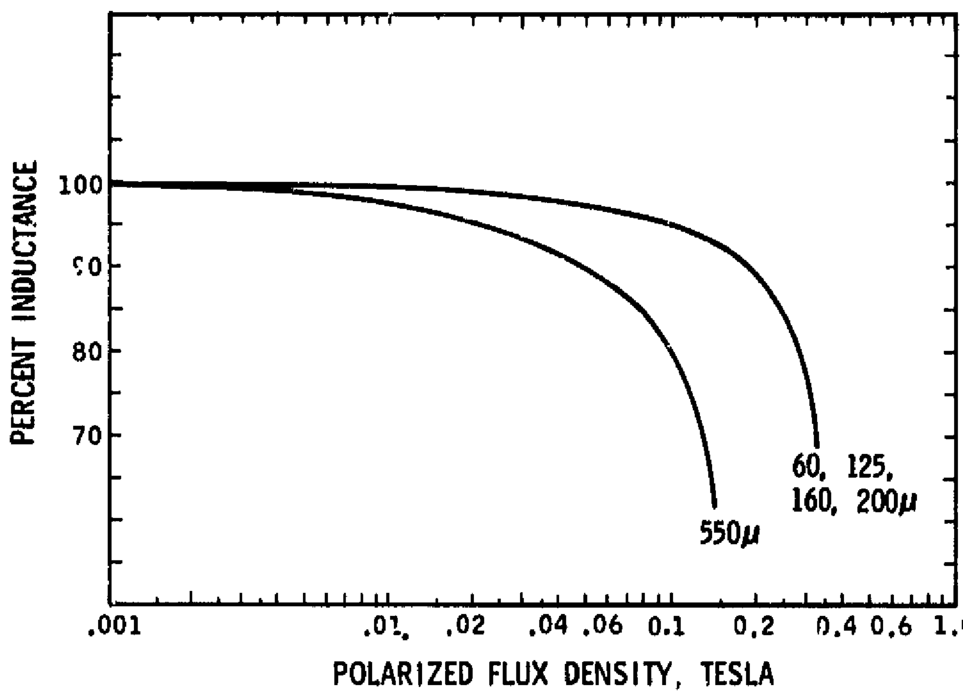

Fig. 4-1. Inductance vs dc bias for moly permalloy cores.

Grain-oriented silicon steels provide greater flexibility in the design of high frequency inductors because the air gap can be adjusted to any desired length and because the relative permeability is high even at high dc flux density. Such steels can develop flux densities of 1.6 tesla, with useful linearity to 1.2 tesla. Moly permalloy* cores carrying dc current on the other hand have useful flux density capabilities to only about 0.3 tesla.

C. RELATIONSHIP OF A_p TO INDUCTOR ENERGY HANDLING CAPABILITY

According to the newly developed approach the energy handling capability of a core is related to its area product A_p by a equation which may be stated as follows:

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$$A_p^* = \left(\frac{2(Eng) \times 10^4}{B_m K_u K_j}\right)^{1.16} \quad [cm^4] \quad (4\text{-}1)$$

K_j = current density coefficient (See Chapter 2.)

K_u = window utilization factor (See Chapter 6.)

B_m = flux density, tesla

Eng = energy, watt seconds

From the above it can be seen that factors such as flux density, window utilization factor K_u (which defines the maximum space which may be occupied by the copper in the window) and the constant K_j (which is related to temperature rise), all have an influence on the inductor area product. The constant K_j is a new parameter that gives the designer control of the copper loss. Derivation is set forth in detail in Chapter 2.

D. FUNDAMENTAL CONSIDERATIONS

The design of a linear reactor depends upon four related factors.

1. Desired inductance
2. Direct current
3. Alternating current ΔI
4. Power loss and temperature rise

With these requirements established, the designer must determine the maximum values for B_{dc} and for B_{ac} which will not produce magnetic saturation, and must make tradeoffs which will yield the highest inductance for a given volume. The core material which is chosen dictates the maximum flux density which can be tolerated for a given design. Magnetic saturation values for different core materials are shown in Table 4-1 as follows.

*Deviation is set forth in detail in Appendix 4.A at the end of this chapter.

Table 4-1. Magnetic material

Material Type		Flux Density (tesla)
Magnesil	3% Si, 97% Fe	1.6
Orthonol	50% Ni, 50% Fe	1.5
48 Alloy	48% Ni, 50% Fe	1.2
Permalloy	79% Ni, 17% Fe, 4% Mo	0.75

It should be remembered that maximum flux density depends upon $B_{dc} + B_{ac}$ in manner shown in Figure 4-2.

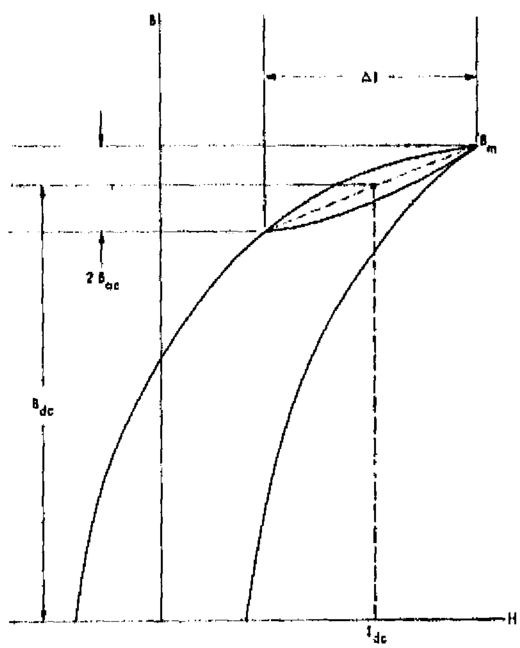

Fig. 4-2. Flux density versus $I_{dc} + \Delta I$

$$B_{max} = B_{dc} + B_{ac} \quad [\text{tesla}]$$

$$B_{dc} = \frac{0.4\pi N I_{dc} \times 10^{-4}}{l_g + \frac{l_m}{\mu_r}} \quad [\text{tesla}] \qquad (4\text{-}2)$$

$$B_{ac} = \frac{0.4\pi N \frac{\Delta I}{2} \times 10^{-4}}{l_g + \frac{l_m}{\mu_r}} \quad [\text{tesla}] \qquad (4\text{-}3)$$

Combining Eqs. (4-2) and (4-3),

$$B_{max} = \frac{0.4\pi N I_{dc} \times 10^{-4}}{l_g + \frac{l_m}{\mu_r}} + \frac{0.4\pi N \frac{\Delta I}{2} \times 10^{-4}}{l_g + \frac{l_m}{\mu_r}} \quad \text{[tesla]} \quad (4\text{-}4)$$

The inductance of an iron-core inductor carrying dc and having an air gap may be expressed as:

$$L = \frac{0.4\pi N^2 A_c \times 10^{-8}}{l_g + \frac{l_m}{\mu_r}} \quad \text{[henry]} \quad (4\text{-}5)$$

Inductance is dependent on the effective length of the magnetic path which is the sum of the air gap length (l_g) and the ratio of the core mean length to relative permeability (l_m/μ_r).

When the core air gap (l_g) is large compared to relative permeability (l_m/μ_r), because of the high relative permeability (μ_r) variations in μ_r do not substantially effect the total effective magnetic path length or the inductance. The inductance equation then reduces to:

$$L = \frac{0.4\pi N^2 A_c \times 10^{-8}}{l_g} \quad \text{[henry]} \quad (4\text{-}6)$$

Final determination of the air gap size requires consideration of the effect of fringing flux which is a function of gap dimension, the shape of the pole faces, and the shape, size and location of the winding. Its net effect is to shorten the air gap.

Fringing flux decreases the total reluctance of the magnetic path and therefore increases the inductance by a factor F to a value greater than that

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calculated from equation 4-6. Fringing flux* is a larger percentage of the total for larger gaps. The fringing flux factor is:

$$F = \left(1 + \frac{l_g}{\sqrt{A_c}} \log_e \frac{2G}{l_g}\right) \qquad (4\text{-}7)$$

where G is a dimension defined in Chapter 2. (This equation is also valid for laminations.)

Equation (4-7) is plotted in Figure 4-3 below.

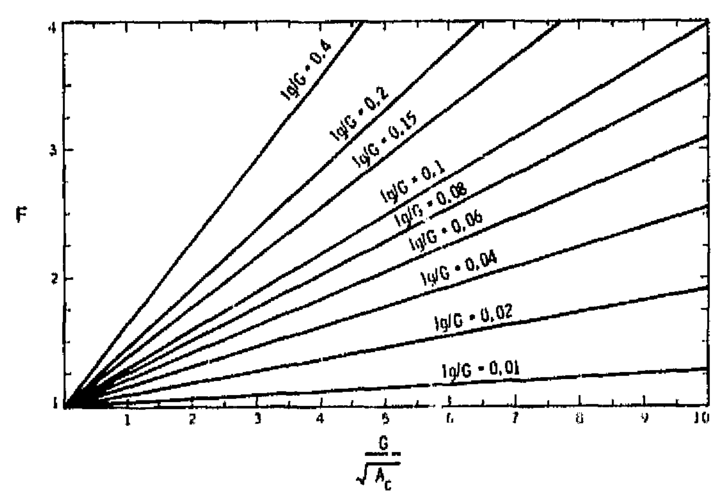

Fig. 4-3. Increase of reactor inductance with flux fringing at the gap.

Inductance L computed in equation (4-6) does not include the effect of fringing flux. The value of inductance L' corrected for fringing flux is:

$$L' = \frac{0.4\pi N^2 A_c F \times 10^{-8}}{l_g} \quad [\text{henry}] \qquad (4\text{-}8)$$

*See Reference 2.

Effective permeability may be calculated from the following expression:

$$\mu_\Delta = \frac{\mu_m}{1 + \frac{l_g}{l_m}\mu_m} \qquad \text{,-9)}$$

μ_m = core material permeability

Curves which have been plotted for values of l_g/l_m from 0 to 0.005 are shown in Figure 4-4.

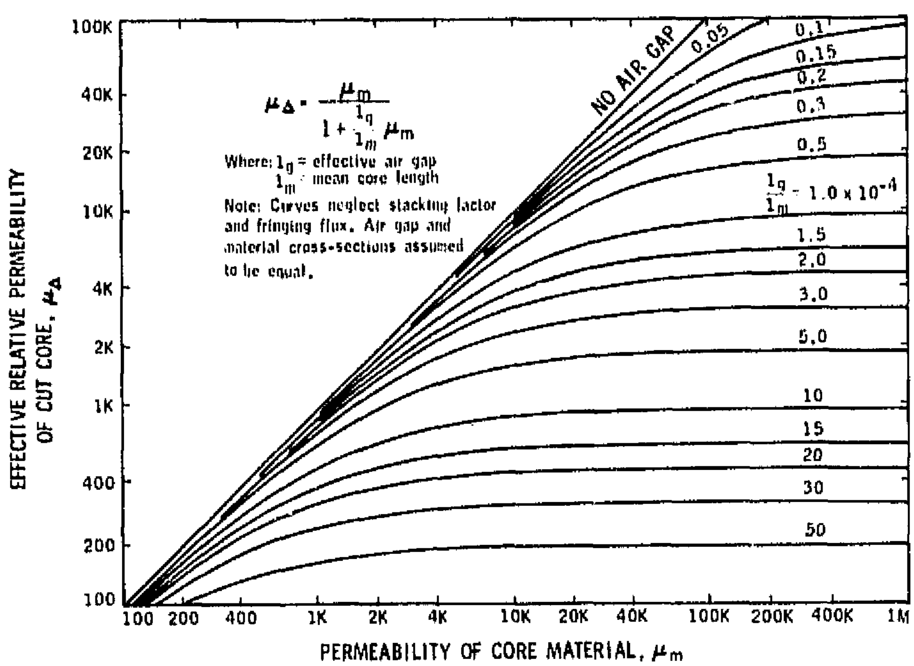

Fig. 4-4. Effective permeability of cut core vs permeability of the material

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The effective design permeability for a butt core joint structure for material permeabilities ranging from 100 to 1,000,000 are shown. Effective permeability variation as a function of core geometry is shown in the curves plotted in Figure 4-5.

After establishing the required inductance and the dc bias current which will be encountered, dimensions can be determined. This requires consideration of the energy handling capability which is controlled by the area product A_p. The energy handling capability of a core is derived from

$$\frac{LI^2}{2} = \text{Energy} \qquad [\text{watt seconds}] \qquad (4\text{-}10)$$

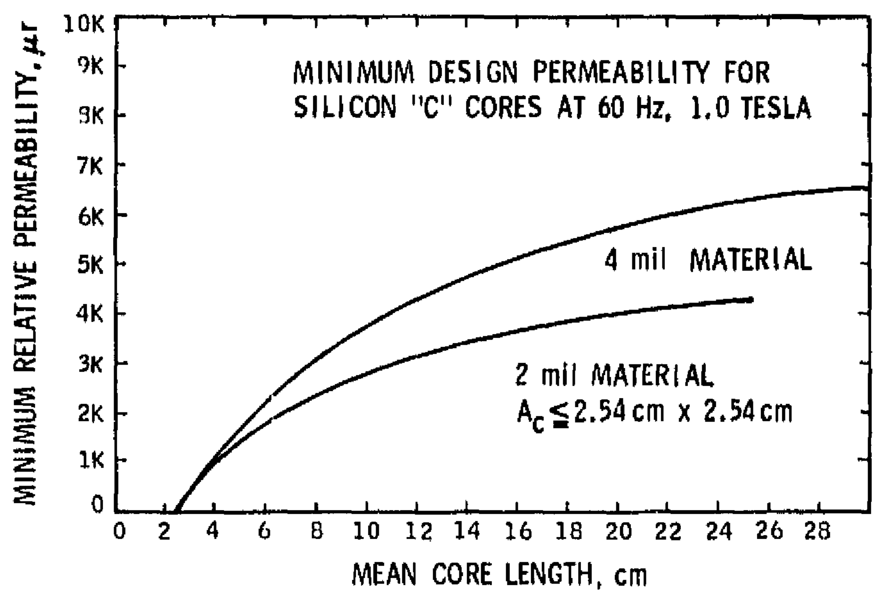

Fig. 4-5. Minimum design permeability

and

$$A_P = \left(\frac{2(\text{Eng}) \times 10^4}{B_m K_u K_j}\right)^{1.16} \qquad [\text{cm}^4] \qquad (4\text{-}11)$$

in which:

B_m = maximum flux density (B_{dc} + B_{ac})

K_u = 0.4 (Chapter 6)

K_j = (See Chapter 2)

Eng = energy, watt seconds

E. DESIGN EXAMPLE

For a typical design example, assume:

1. Inductance 0.015 henrys
2. dc current 2 amp
3. ac current 0.1 amp
4. 25°C rise
5. Frequency 20 KHz

The procedure would then be as follows:

Step No. 1. Calculate the energy involved from equation (4-10):

$$Eng = \frac{LI^2}{2} \qquad (4\text{-}12)$$

$$Eng = \frac{0.015(2.0)^2}{2}$$

$$Eng = 0.030 \qquad [\text{watt seconds}]$$

Step No. 2. Calculate the area product A_p from equation (4-1):

$$A_p = \left(\frac{2(Eng) \times 10^4}{B_m K_u K_j}\right)^{1.16} \qquad [cm^4]$$

$$A_p = \left(\frac{2(0.03) \times 10^4}{(1.2)(0.4)(395)}\right)^{1.16} = 3.80 \qquad [cm^4]$$

A core which has an area product closest to the calculated value is the AL-10 which is described in Table 2-6, Chapter 2, and Appendix 4B. That size core has an area product A_p of 3.85 cm^4 (A_c = 1.34 eff. cm^2 and W_a = 2.87 cm^2).

After the A_p has been determined, the geometry of the inductor can be evaluated as described in Chapter 2 for weight, surface area, volume, and appropriate changes made, if required.

Step No. 3. Determine the current density from:

$$J = K_j A_p^{-0.14} \qquad (4\text{-}13)^*$$

$$J = 395(3.80)^{-0.14} = 328 \text{ amps/cm}^2$$

Step No. 4. Determine the wire size from:

$$\text{Wire size} = \frac{I_{dc}}{\text{amp/cm}^2}$$

$$\text{Wire size} = \frac{2}{328} = 0.00609 \qquad [\text{cm}^2]$$

Select the wire size from Table 6-1, column A, Chapter 6. The rule is that when the calculated wire size does not fall close to those listed in the table, the next smallest size should be selected.

The closest wire size to 0.00609 is AWG No. 20

$$\text{Area} = 0.005188 \text{ (bare)} \qquad [\text{cm}^2]$$

Step No. 5. Calculate the number of turns.

The number of turns per square cm for No. 20 wire is 98.9 based on 60% wire fill factor data taken from Table 6-1, Chapter 6, column J.

$$\text{effective window} \times \text{turns/cm}^2$$

$$2.58 \times 98.9 = 255$$

$$\text{Total number of turns} = 255$$

*Derivation of equation (4-13) is shown in Chapter 2.

Step No. 6. The air gap dimension is determined from equation (4-6) by solving for l_g as follows:

$$l_g = \frac{0.4\pi \ N^2 A_c \times 10^{-8}}{L} \qquad (4\text{-}14)$$

$$l_g = \frac{1.26(255)^2(1.342) \times 10^{-8}}{(0.015)}$$

$$l_g = 0.0733 \qquad [cm]$$

Gap spacing is usually maintained by inserting Kraft paper. However this paper is available only in mil thicknesses. Since l_g has been determined in cm, it is necessary to convert as follows:

$$cm \times 393.7 = mils \text{ (inch system)}$$

Substituting values:

$$0.0733 \times 393.7 = 28.8 \qquad [mils]$$

An available size of paper is 15 mil sheet. Two thicknesses would therefore be used, giving equal gaps in both legs.

The effect of fringing flux upon inductance can now be considered. As mentioned, the data shown in Figure 4-3 were developed to show graphically the effect of gap length l_g variation on fringing flux. In order to use this data, the ratio of l_g to window length G must be determined. For the AL-10 size, Table 4.B-8 shows a G value of 3.015 cm. Therefore:

$$\frac{l_g}{G} = \frac{0.0733}{3.015} = 0.0243 \qquad [cm]$$

and accordingly

$$\frac{G}{\sqrt{A_c}} = \frac{3.015}{1.16} = 2.60$$

The fringing flux factor F from Figure 4-3 may be stated:

$$F = 1.28$$

The recalculated number of turns can be determined by rewriting equation 4-8:

$$N = \sqrt{\frac{l_g L}{0.4\pi A_c F \times 10^{-8}}}$$

and by inserting the known values

$$N = \sqrt{\frac{(0.0733)(0.015)}{(1.26)(1.342)(1.28) \times 10^{-8}}} = 226$$

Step No. 7. Calculate the ac and dc flux density from equation (4-4)

$$B_{max} = \frac{0.4\pi N\left(I_{dc} + \frac{\Delta I}{2}\right) \quad 10^{-4}}{l_g} \qquad \text{[tesla]}$$

$$B_{max} = \frac{(1.26)(226)(2 + 0.05) \times 10^{-4}}{(0.0733)} \qquad \text{[tesla]}$$

$$B_{max} = 0.793 \qquad \text{[tesla]}$$

Step No. 8. Calculate core loss. This may be determined from Figure 4-6, in conjunction with the equation below:

$$B_{ac} = \frac{0.4\pi N \frac{\Delta I}{2} \times 10^{-4}}{l_g} \qquad \text{[tesla]}$$

$$B_{ac} = \frac{(1.26)(226)(0.05) \times 10^{-4}}{(0.0733)} \qquad [\text{tesla}]$$

$$B_{ac} = 0.0194 \qquad [\text{tesla}]$$

The ac core loss for this value can be found by reference to the graph shown in Figure 4-6 which is based upon solutions of the following expression for various operating frequencies:

$$P_{fe} = \frac{\text{milliwatts}}{\text{gram}} \times W_t$$

Referring to Table 4.B-8 for the AL-10 size core, the weight of the core is 110 grams. The core loss in milliwatts per gram is obtained from:

$$P_{fe} = (2.1)(110) = 230 \qquad [\text{milliwatts}]$$

Step No. 9. Calculate copper loss and temperature rise.

The resistance of a winding is the mean length turn in cm multiplied by the resistance in micro ohms per cm and the total number of turns. Referring to Table 4.B-8 for the AL-10 size core for the mean length per turn (MLT) and the wire table (Chapter 6) for the resistance of No. 20 wire then:

$$R = MLT \times N \times (\text{Column C}) \times \zeta \times 10^{-6} \qquad [\Omega]$$

$$R = 8.33 \times 226 \times 332 \times 1.098 \times 10^{-6}$$

$$R = 0.686 \qquad [\Omega]$$

Since power loss is $P_{cu} = I^2R$,

$$P_{cu} = (2)^2(0.625) = 2.75 \qquad [\text{watts}]$$

$$P_{\Sigma} = P_{cu} + P_{fe}$$

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$$P_\Sigma = 2.74 + 0.165$$

$$P_\Sigma = 2.90 \qquad \text{[watts]}$$

From Chapter 7 the surface area A_t required to dissipate waste heat (expressed as watts loss per unit area) is:

$$A_t = \frac{P_\Sigma}{\Psi}$$

$$\Psi = 0.03 \text{ W/cm}^2 \text{ at } 25°\text{C rise}$$

Referring to Table 4.B-8 for the AL-10 size core, the surface area A_t is 79.39 cm^2.

$$\Psi = \frac{P_\Sigma}{A_t}$$

$$\Psi = \frac{2.90}{79.39} = 0.0365 \qquad [\text{W/cm}^2]$$

which will produce the required temperature rise.

(In a test sample made to prove out this example, the measured inductance was found to be 0.0159 hy with a resistance of 0.600 ohms at 25°C and a resistance of 0.647 Ω at 45°C.)

With the reduction in turns resulting from consideration of fringing flux in some cases the designer may be able to increase the wire size and reduce the copper loss.

This completes the explanation of the example.

Much of the information which the designer needs can only be found in a scattered variety of texts and other literature. To make this information more conveniently available, helpful data has been gathered together and reproduced in Appendix 4.B which contains 20 tables and 22 figures. The index has been prepared to make it possible for the designer to readily locate specific information.

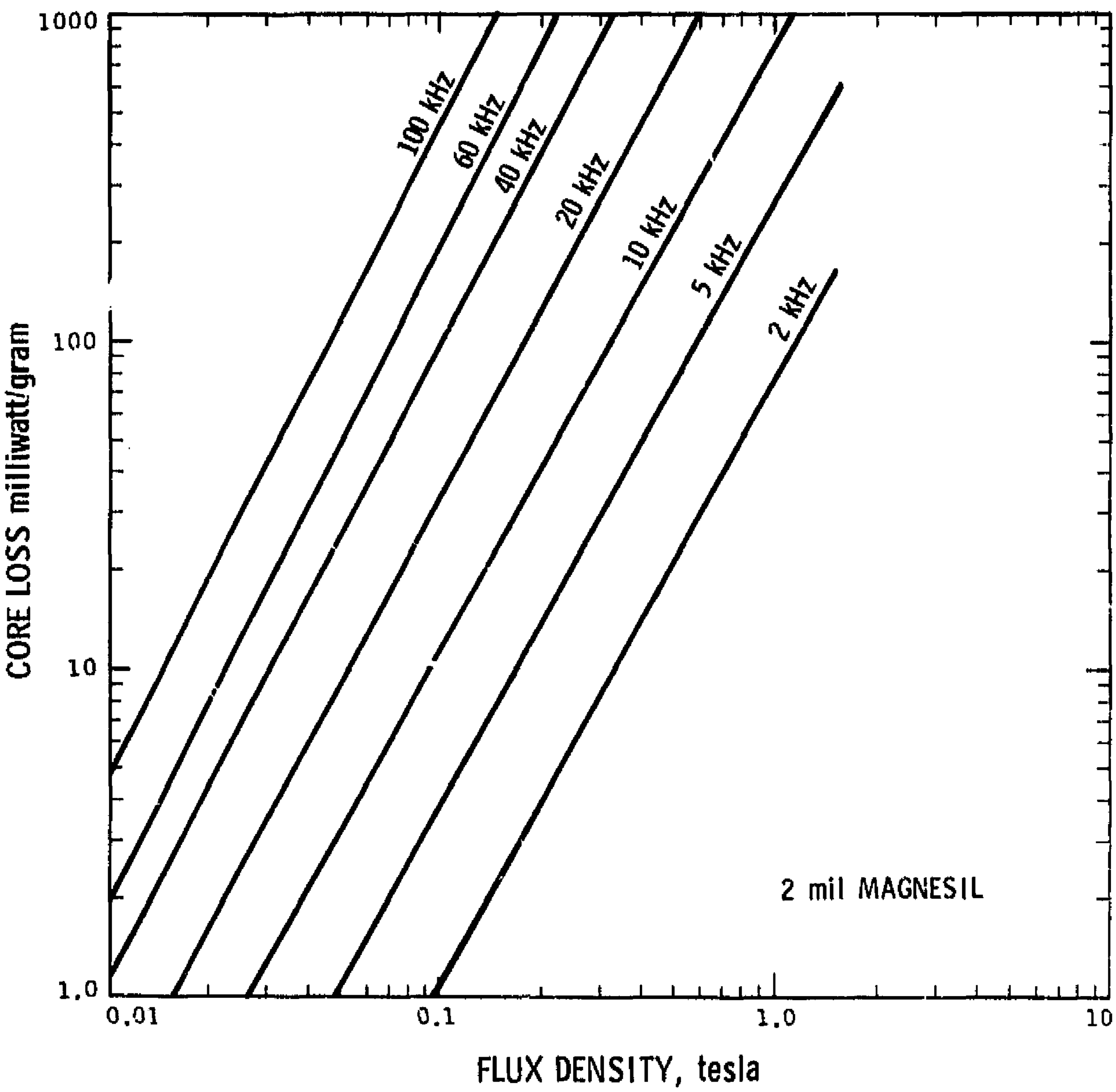

Fig. 4-6. Design curves showing maximum core loss for 2 mil silicon "C" cores

APPENDIX 4-A

LINEAR REACTOR DESIGN WITH AN IRON CORE

After calculating the inductance and dc current, select the proper size core with a given $LI^2/2$. The energy handling capability of an inductor can be determined by its area product A_p of which, W_a is the available core window area in cm^2 and A_c is the core effective cross sectional area cm^2. The W_aA_c or area product A_p relationship is obtained by solving $E = LdI/dt$ as follows:*

$$E = L\frac{dI}{dt} = N\frac{d\phi}{dt} \qquad (4.A\text{-}1)$$

$$L = N\frac{d\phi}{dI} \qquad (4.A\text{-}2)$$

$$\phi = B_m A_c' \qquad (4.A\text{-}3)$$

$$B_m = \frac{\mu_o NI}{l_g' + \frac{l_m'}{\mu_r}} \qquad (4.A\text{-}4)$$

$$\phi = \frac{\mu_o NI A_c'}{l_g' + \frac{l_m'}{\mu_r}} \qquad (4.A\text{-}5)$$

$$\frac{d\phi}{dI} = \frac{\mu_o N A_c'}{l_g' + \frac{l_m'}{\mu_r}} \qquad (4.A\text{-}6)$$

$$L = N\frac{d\phi}{dI} = \frac{\mu_o N^2 A_c'}{l_g' + \frac{l_m'}{\mu_r}} \qquad (4.A\text{-}7)$$

*Symbols marked with a prime (such as H') are mks (meter kilogram second) units.

$$\text{Energy} = \frac{1}{2} LI^2 = \frac{\mu_o N^2 A_c' I^2}{2\left(l_g' + \frac{l_m'}{\mu_r}\right)} \tag{4.A-8}$$

If B_m is specified,

$$I = \frac{B_m\left(l_g' + \frac{l_m'}{\mu_r}\right)}{\mu_o N} \tag{4.A-9}$$

$$\text{Eng} = \frac{\mu_c N^2 A_c'}{2\left(l_g' + \frac{l_m'}{\mu_r}\right)} \left(\frac{B_m\left(l_g' + \frac{l_m'}{\mu_r}\right)}{\mu_o N}\right)^2 \tag{4.A-10}$$

$$\text{Eng} = \frac{B_m{}^2\left(l_g' + \frac{l_m'}{\mu_r}\right) A_c'}{2\mu_o} \tag{4.A-11}$$

$$I = \frac{K_u W_a' J'}{N} = \frac{\left(B_m\left(l_g' + \frac{l_m'}{\mu_r}\right)\right)}{\mu_o N} \tag{4.A-12}$$

Solving for $(l_g' + l_m'/\mu_r)$

$$\left(l_g' + \frac{l_m'}{\mu_r}\right) = \frac{\mu_o K_u W_a' J'}{B_m} \tag{4.A-13}$$

Substituting into the energy equation

$$\text{Eng} = \frac{B_m^{\,2}\left(\dfrac{\mu_o K_u W_a' J'}{B_m}\right) A_c'}{2\mu_o} \qquad (4.A\text{-}14)$$

$$\text{Eng} = \frac{B_m^{\,2} A_c'}{2\mu_o} \times \frac{\mu_o K_u W_a' J'}{B_m} \qquad (4.A\text{-}15)$$

$$\text{Eng} = \frac{B_m K_u W_a' A_c' J'}{2} \qquad (4.A\text{-}16)$$

let

W_a = window area, cm^2

A_c = core area, cm^2

J = current density, amps/cm^2

H = magnetizing force, amp turn/cm

l_g = air gap, cm

l_m = magnetic path length, cm

$W_a' = W_a \times 10^{-4}$

$A_c' = A_c \times 10^{-4}$

$J' = J \times 10^4$

$l_m' = l_m \times 10^{-2}$

$l_g' = l_g \times 10^{-2}$

$H' = H \times 10^2$

Substituting into the energy equation

$$\mathrm{Eng} = \frac{W_a A_c B_m J K_u}{2} \times 10^{-4} \qquad (4.A\text{-}17)$$

Solving for $A_p = W_a A_c$

$$A_p = \frac{2(\mathrm{Eng})}{B_m J K_u} \times 10^4 \qquad (4.A\text{-}18)$$

Combining equation from Table 2-1,

$$J = K_j A_p^{-0.14} \qquad (4.A\text{-}19)$$

yielding:

$$A_p = \frac{2(\mathrm{Eng}) \times 10^4}{K_u B_m \left(K_j A_p^{-0.14}\right)} \qquad (4.A\text{-}20)$$

$$A_p^{0.86} = \frac{2(\mathrm{Eng}) \times 10^4}{K_u B_m K_j} \qquad (4.A\text{-}21)$$

$$A_p = \left(\frac{2(\mathrm{Eng}) \times 10^4}{K_u B_m K_j}\right)^{1.16} \left[\mathrm{cm}^4\right] \qquad (4.A\text{-}22)$$

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APPENDIX 4.B

C CORE AND BOBBIN MAGNETIC AND DIMENSIONAL SPECIFICATION

A. Definitions for Tables 4.B-1 through 4.B-20

Tables 4.B-1 through 4.B-20* show magnetic and dimensional specifications for twenty C cores. The information is listed by line as:

1 Manufacture and part number
2 Units
3 Ratio of the window area over the iron area
4 Product of the window area times the iron area
5 Window area W_a gross
6 Iron area A_c effective
7 Mean magnetic path length lm
8 Core weight of silicon steel multiplied by the stacking factor
9 Copper weight single bobbin
10 Mean length turn
11 Ratio of G dimension divided by the square root of the iron area (Ac)
12 Ratio of the W_a (eff)/W_a
13 Inductor overall surface area A_t
14-17 "C" core dimensions
18 Bobbin manufacturer and part number**†
19 Bobbin inside winding length†
20 Bobbin inside build†
21 Bobbin winding area length times build†
22 Bracket manufacturer and part number††

B. Nomographs for 20 C core sizes

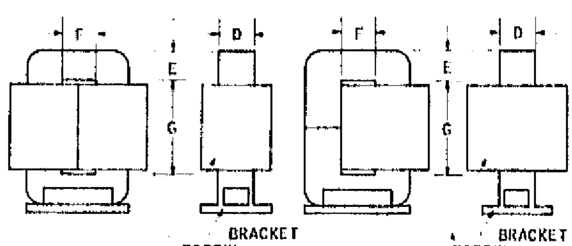
Figures 4.B-1 through 4.B-20 are graphs for 20 different "C" cores. The nomographs display resistance, number of turns, and wire size at a fill factor of K_2 = 0.60. These graphs are included to provide a close approximation for breadboarding purposes.

*References 3, 4.

**The first number in front of the part number indicates the number of bobbins.
†Dorco Electronics, 15533 Vermont Ave., Paramount, Calif. 90723.
††Hallmark Metals, 610 West Foothill Blvd., Glendora, Calif. 91740.

Table 4.B-1. "C" core AL-2

"C" CORE	AL 2			
	ENGLISH		METRIC	
Wa/Ac			3.32	
Wa x Ac	0.0073	in^4	0.265	cm^4
Wa	0.156	in^2	1.008	cm^2
Ac (effective)	0.041	in^2	0.264	cm^2
lm	2.233	in	5.671	cm
CORE WT	0.027	lb	12.23	grams
COPPER WT	0.371	lb	16.87	grams
* MLT FULLWOUND	1.76	in	4.47	cm
$G/\sqrt{Ac}$			3.08	
Wa (effective) /Wa			0.836	
A_T	3.80	in^2	24.56	cm^2
D	0.250	in	0.635	cm
E	0.187	in	0.474	cm
F	0.250	in	0.635	cm
G	0.625	in	1.587	cm
BOBBIN	DORCO ELECTRONICS * 1-L-2			
LENGTH	0.580	in	1.473	cm
BUILD	0.225	in	0.571	cm
* Wa (effective)	0.130	in^2	0.841	cm^2
BRACKET	HALLMARK METALS * 04 010 03			



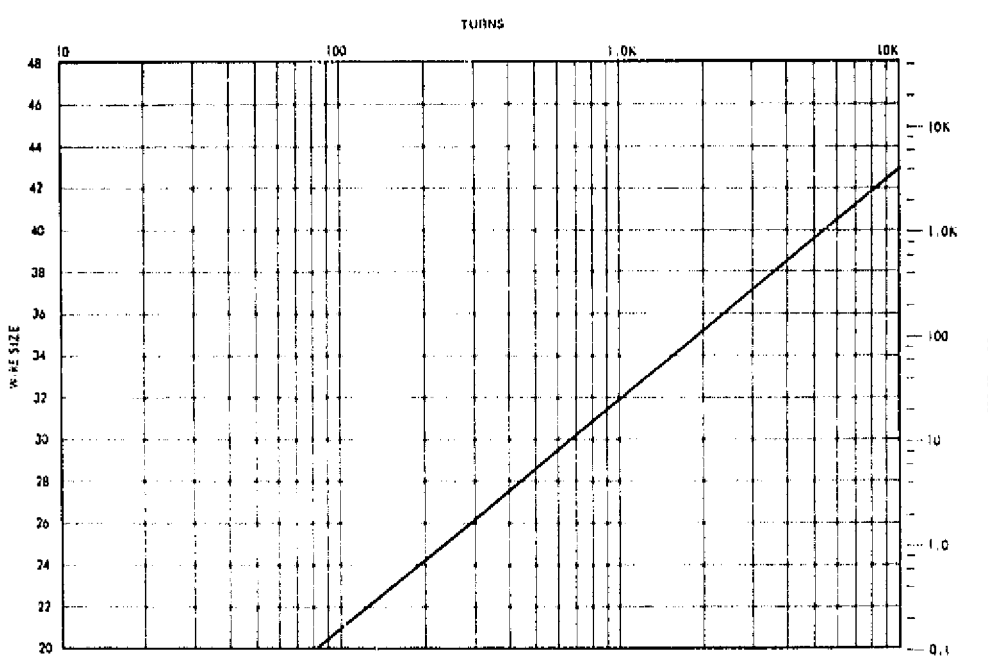
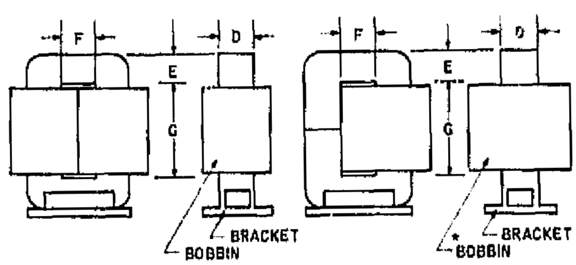


Fig. 4.B-1. Wiregraph for "C" core AL-2

Table 4.B-2. "C" core AL-3

"C" CORE	AL-3			
	ENGLISH		METRIC	
Wa/Ac			2.23	
Wa x Ac	0.0098	in^4	0.410	cm^4
Wa	0.156	in^2	1.006	cm^2
Ac (effective)	0.063	in^2	0.406	cm^2
lm	2.233	in	5.671	cm
CORE WT	0.04	lb	18.12	grams
COPPER WT	0.042	lb	19.26	grams
* MLT FULLWOUND	2.01	in	5.10	cm
$G/\sqrt{Ac}$			2.49	
Wa (effective) /Wa			0.836	
A_T	4.27	in^2	27.58	cm^2
D	0.375	in	0.952	cm
E	0.187	in	0.474	cm
F	0.250	in	0.635	cm
G	0.625	in	1.587	cm
BOBBIN	DORCO ELECTRONICS *		1-L-3	
LENGTH	0.580	in	1.473	cm
BUILD	0.225	in	0.571	cm
* Wa (effective)	0.130	in^2	0.841	cm^2
BRACKET	HALLMARK METALS *		06-010-03	



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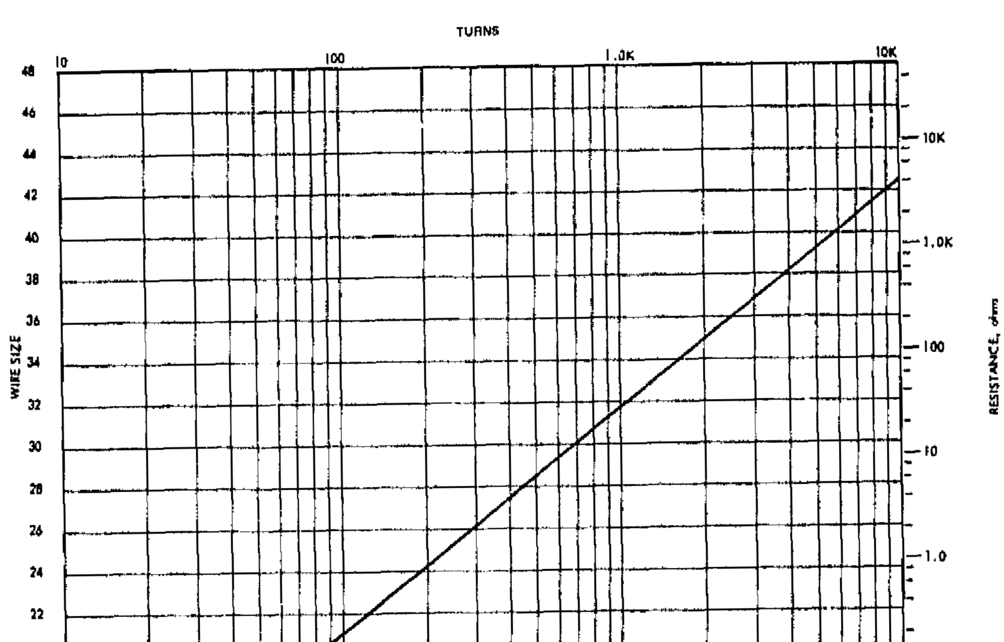
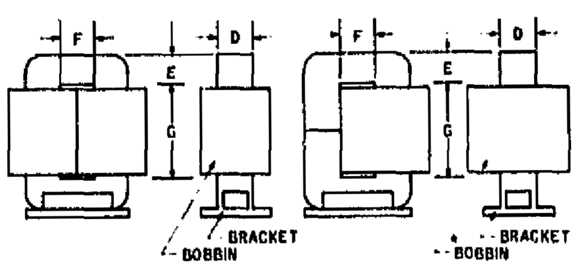


Fig. 4.B-2. Wiregraph for "C" core AL-3

Table 4.B-3. "C" core AL-5

"C" CORE	AL-5			
	ENGLISH		METRIC	
Wa/Ac			2.33	
Wa x Ac	0.018	in^4	0.787	cm^4
Wa	0.219	in^2	1.423	cm^2
Ac (effective)	0.0838	in^2	0.539	cm^2
lm	2.933	in	7.45	cm
CORE WT	0.067	lb	30.4	grams
COPPER WT	0.0643	lb	29.2	grams
* MLT FULLWOUND	2.13	in	5.42	cm
$G/\sqrt{Ac}$			3.028	
Wa (effective) /Wa			0.843	
A_T	5.90	in^2	38.1	cm^2
D	0.375	in	0.952	cm
E	0.250	in	0.635	cm
F	0.250	in	0.635	cm
G	0.875	in	2.22	cm
BOBBIN	DORCO ELECTRONICS *		1-L-5	
LENGTH	0.830	in	2.11	cm
BUILD	0.225	in	0.571	cm
* Wa (effective)	0.188	in^2	1.20	cm^2
BRACKET	HALLMARK METALS *		06-012-04	



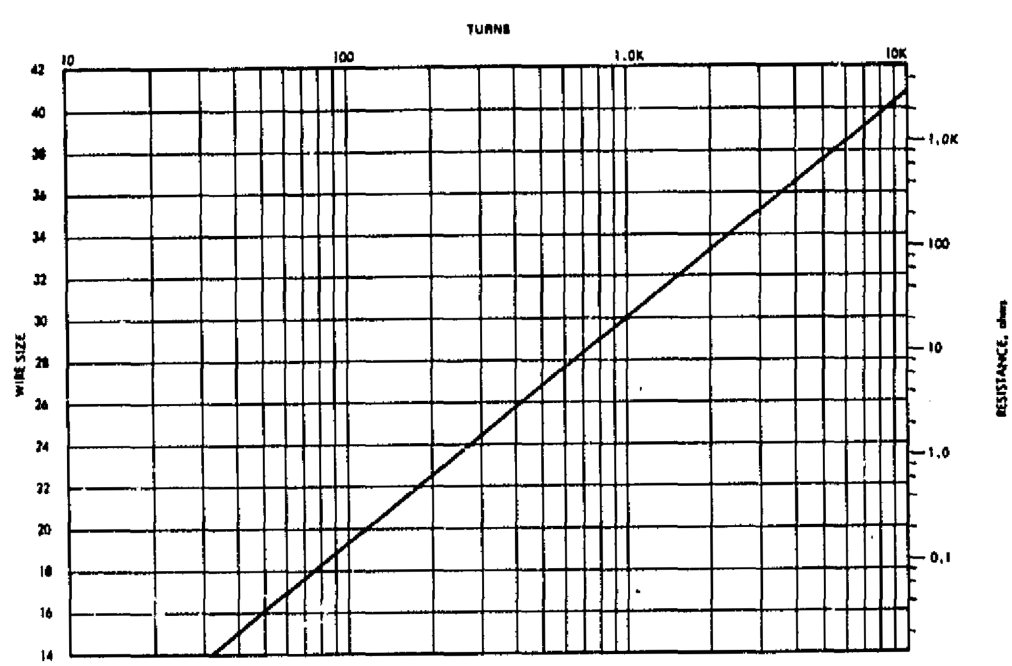
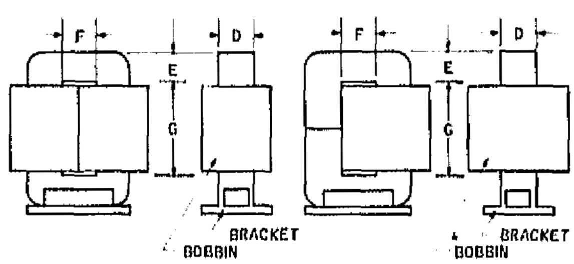


Fig. 4.B-3. Wiregraph for "C" core AL-5

Table 4.B-4. "C" core AL-6

"C" CORE	AL-6			
	ENGLISH		METRIC	
Wa/Ac			1.75	
Wa x Ac	0.024	in^4	1.011	cm^4
Wa	0.219	in^2	1.413	cm^2
Ac (effective)	0.111	in^2	0.716	cm^2
lm	2.933	in	7.45	cm
CORE WT	0.091	lb	41.2	grams
COPPER WT	0.0719	lb	32.6	grams
* MLT FULLWOUND	2.38	in	6.00	cm
$G/\sqrt{Ac}$			2.03	
Wa (effective) /Wa			0.843	
A_T	6.50	in^2	41.9	cm^2
D	0.500	in	1.27	cm
E	0.250	in	0.635	cm
F	0.250	in	0.635	cm
G	0.875	in	2.22	cm
BOBBIN	DORCO ELECTRONICS *		I-L-6	
LENGTH	0.830	in	2.11	cm
BUILD	0.225	in	0.571	cm
* Wa (effective)	0.186	in^2	1.20	cm^2
BRACKET	HALLMARK METALS *		08-012-04	



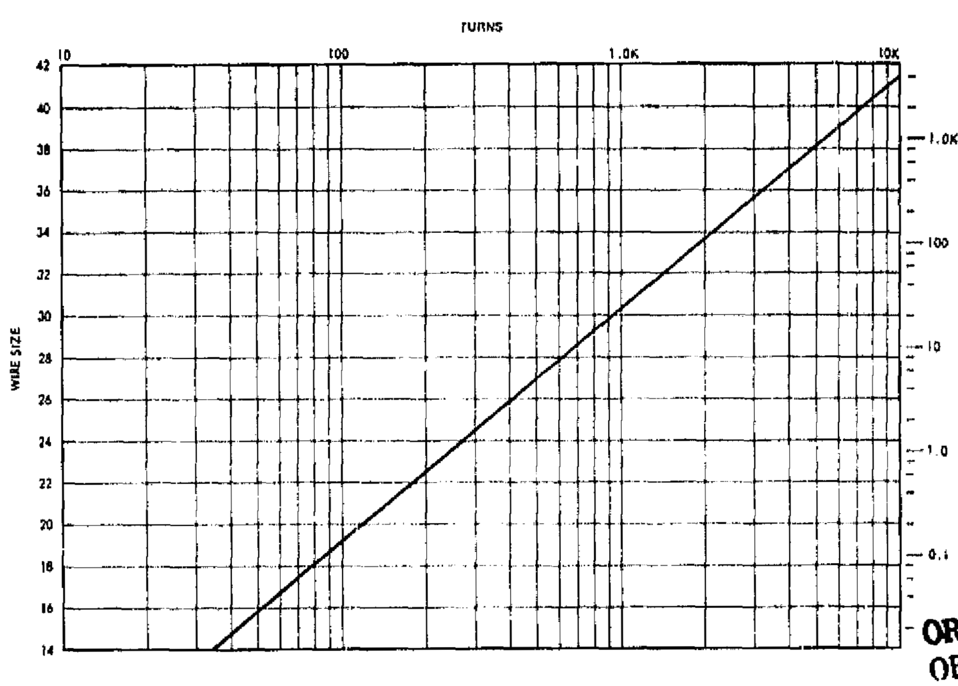


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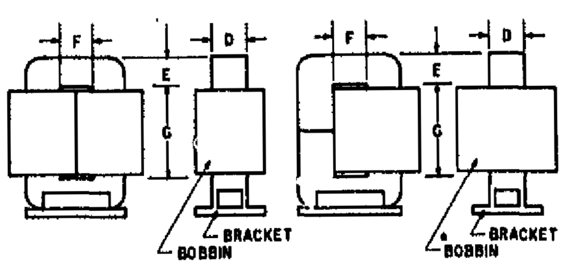
Fig. 4.B-4. Wiregraph for "C" core AL-6

Table 4.B-5. "C" core AL-124

"C" CORE	AL-124			
	ENGLISH		METRIC	
Wa/Ac			2.80	
Wa x Ac	0.0347	in^4	1.44	cm^4
Wa	0.313	in^2	2.02	cm^2
Ac (effective)	0.111	in^2	0.718	cm^2
lm	3.306	in	8.40	cm
CORE WT	0.103	lb	46.7	grams
COPPER WT	0.115	lb	52.13	grams
* MLT FULLWOUND	2.59	in	6.58	cm
$G/\sqrt{Ac}$			3.00	
Wa (effective) /Wa			0.876	
A_T	8.03	in^2	51.79	cm^2
D	0.500	in	1.27	cm
E	0.250	in	0.635	cm
F	0.313	in	0.795	cm
G	1.00	in	2.54	cm
BOBBIN	DORCO ELECTRONICS # 1-L-124			
LENGTH	0.955	in	2.426	cm
BUILD	0.288	in	0.731	cm
* Wa (effective)	0.275	in^2	1.77	cm^2
BRACKET	HALLMARK METALS # 08-013-04			



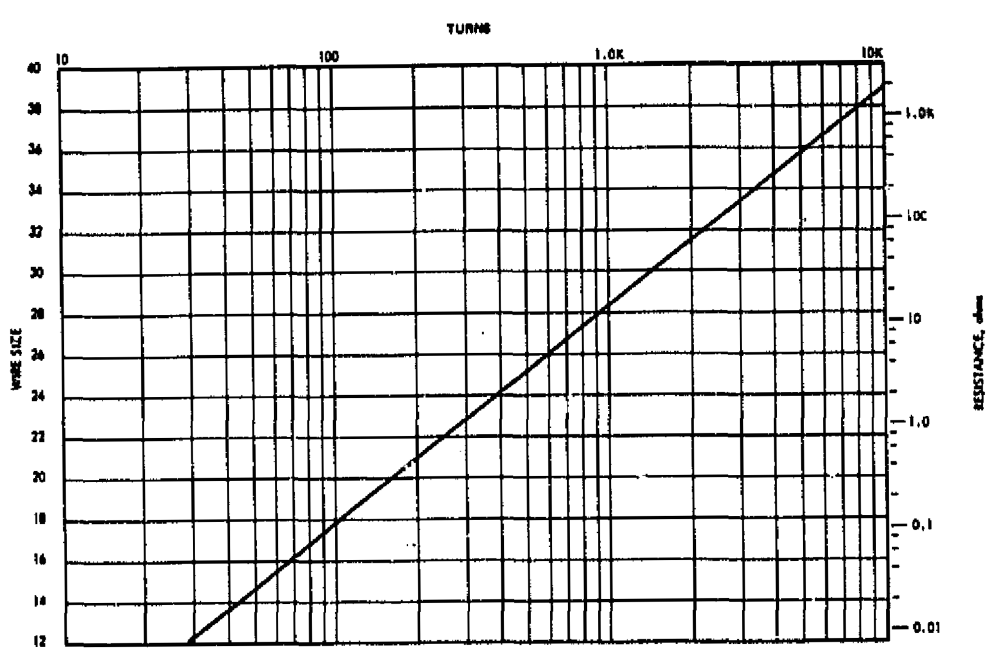
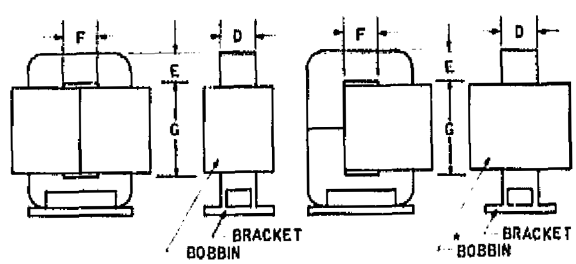


Fig. 4.B-5. Wiregraph for "C" core AL-124

Table 4.B-6. "C" core AL-8

"C" CORE	AL-8			
	ENGLISH		METRIC	
Wa/Ac			3.16	
Wa x Ac	0.068	in^4	2.31	cm^4
Wa	0.446	in^2	2.87	cm^2
Ac (effective)	0.126	in^2	0.808	cm^2
lm	4.198	in	10.66	cm
CORE WT	0.147	lb	66.59	grams
COPPER WT	0.180	lb	81.7	grams
* MLT FULLWOUND	2.77	in	7.06	cm
$G/\sqrt{Ac}$			3.36	
Wa (effective) /Wa			0.898	
A_T	11.29	in^2	72.8	cm^2
D	0.376	in	0.952	cm
E	0.376	in	0.952	cm
F	0.376	in	0.952	cm
G	1.187	in	3.015	cm
BOBBIN	DORCO ELECTRONICS *		1-L-8	
LENGTH	1.142	in	2.9	cm
BUILD	0.350	in	0.889	cm
* Wa (effective)	0.399	in^2	2.578	cm^2
BRACKET	HALLMARK METALS *		06-102-08	



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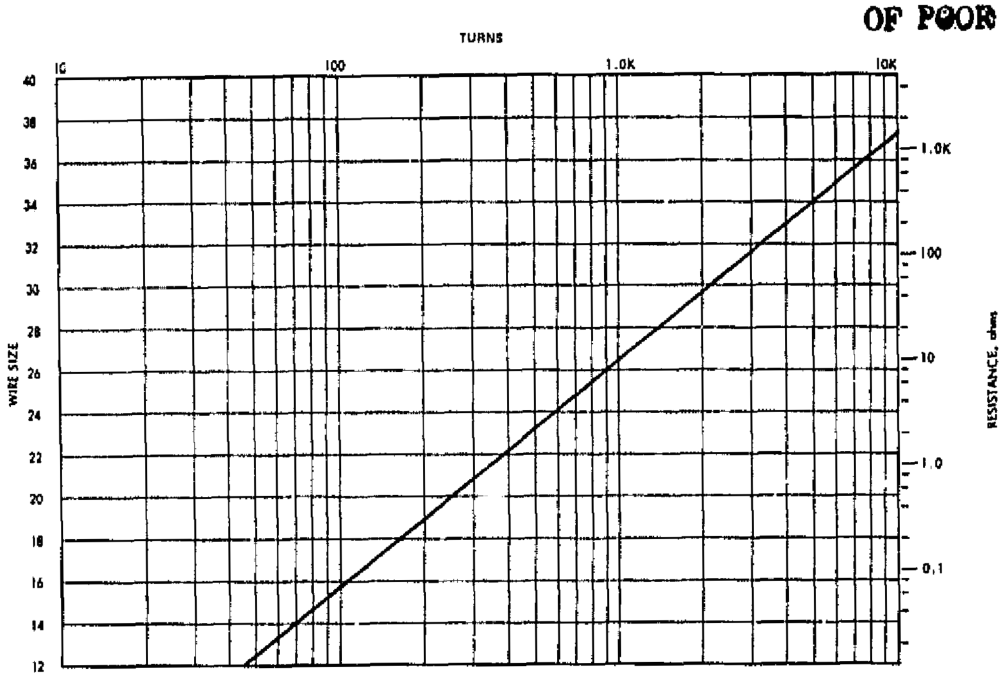
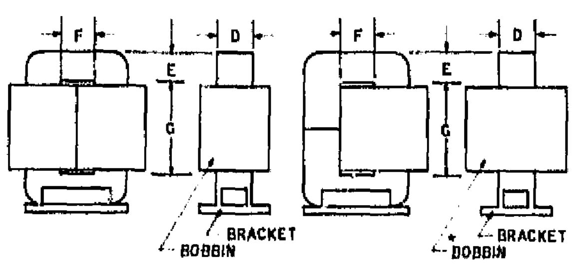


Fig. 4.B-6. Wiregraph for "C" core AL-8

Table 4.B-7. "C" core AL-9

"C" CORE	AL-8			
	ENGLISH		METRIC	
Wa/Ac			2.37	
Wa x Ac	0.074	in^4	3.09	cm^4
Wa	0.446	in^2	2.870	cm^2
Ac (effective)	0.167	in^2	1.077	cm^2
lm	4.198	in	10.66	cm
CORE WT	0.197	lb	89.2	grams
COPPER WT	0.198	lb	89.0	grams
* MLT FULLWOUND	3.02	in	7.69	cm
$G/\sqrt{Ac}$			2.90	
Wa (effective) /Wa			0.898	
A_T	12.15	in^2	78.39	cm^2
D	0.500	in	1.27	cm
E	0.375	in	0.952	cm
F	0.375	in	0.952	cm
G	1.197	in	3.015	cm
BOBBIN	DORCO ELECTRONICS *		1-L-9	
LENGTH	1.142	in	2.90	cm
BUILD	0.350	in	0.889	cm
* Wa (effective)	0.399	in^2	2.578	cm^2
BRACKET	HALLMARK METALS *		08-102-08	



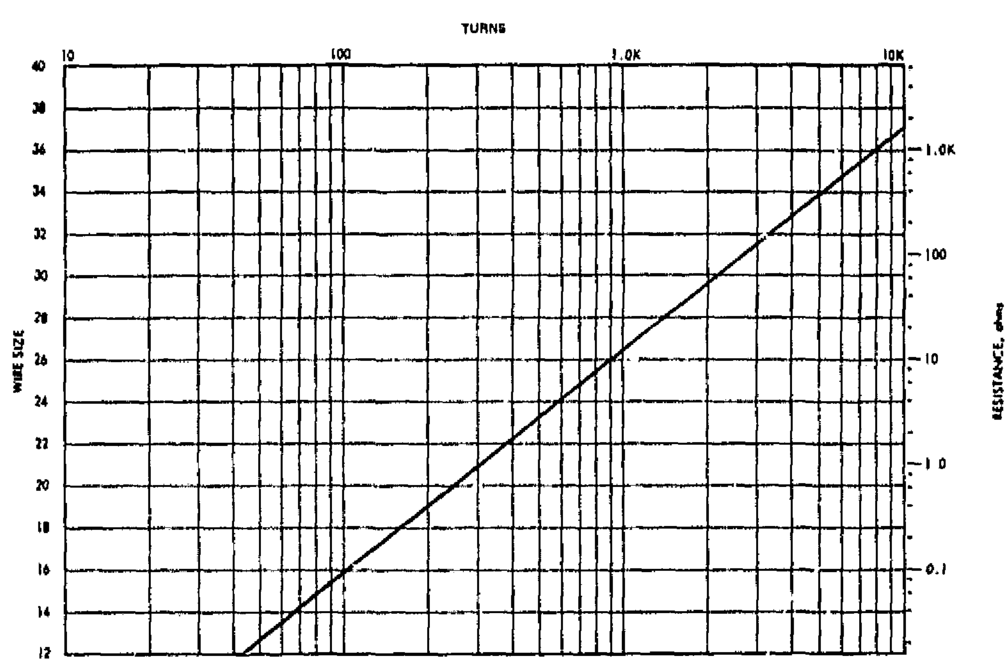
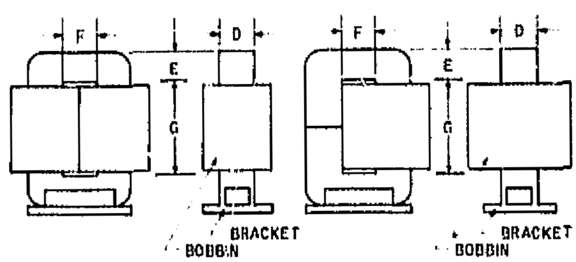


Fig. 4.B-7. Wiregraph for "C" core AL-9

Table 4.B-8. "C" core AL-10

"C" CORE	AL-10			
	ENGLISH		METRIC	
Wa/Ac			1.90	
Wa x Ac	0.092	in^4	3.86	cm^4
Wa	0.445	in^2	2.870	cm^2
Ac (effective)	0.208	in^2	1.342	cm^2
lm	4.198	in	10.66	cm
CORE WT	0.243	lb	110	grams
COPPER WT	0.213	lb	90.4	grams
* MLT FULLWOUND	3.27	in	8.33	cm
$G/\sqrt{Ac}$			2.603	
Wa (effective) /Wa			0.898	
A_T	13.01	in^2	83.9	cm^2
D	0.625	in	1.587	cm
E	0.375	in	0.952	cm
F	0.375	in	0.952	cm
G	1.187	in	3.015	cm
BOBBIN	DORCO ELECTRONICS *		1 L-10	
LENGTH	1.142	in	2.90	cm
BUILD	0.350	in	0.889	cm
* Wa (effective)	0.399	in^2	2.578	cm^2
BRACKET	HALLMARK METALS *		010-102-06	



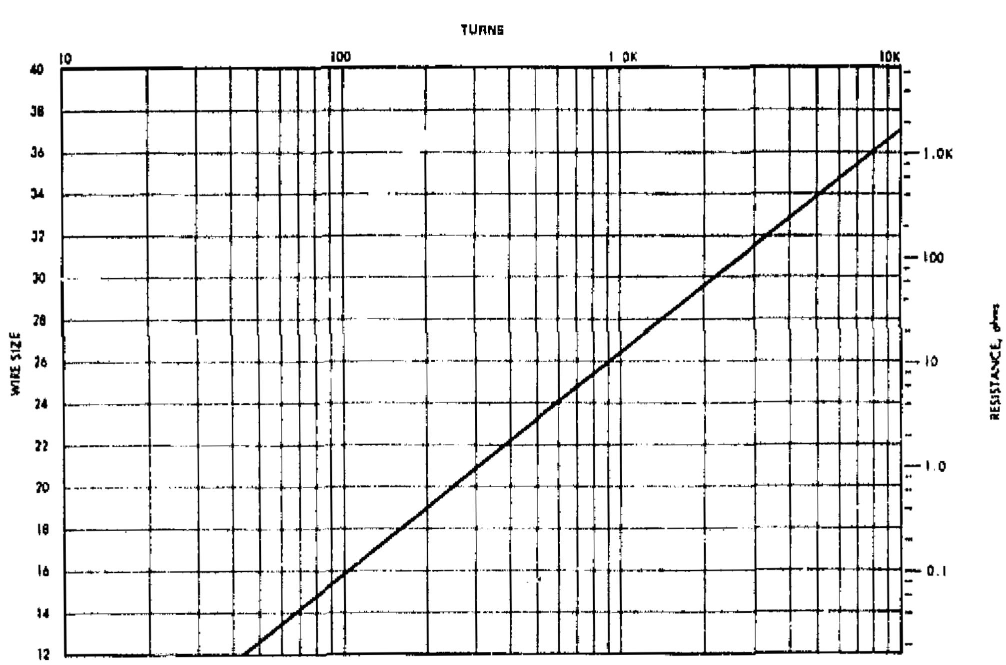
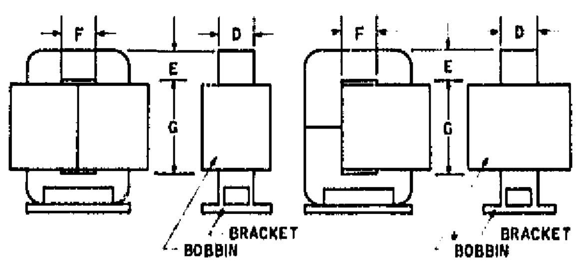


Fig. 4.B-8. Wiregraph for "C" core AL-10

Table 4.B-9. "C" core AL-12

"C" CORE	AL-12			
	ENGLISH		METRIC	
Wa/Ac			2.57	
Wa x Ac	0.108	in^4	4.57	cm^4
Wa	0.563	in^2	3.63	cm^2
Ac (effective)	0.195	in^2	1.26	cm^2
lm	4.523	in	11.5	cm
CORE WT	0.244	lb	110	grams
COPPER WT	0.295	lb	133.7	grams
* MLT FULLWOUND	3.54	in	9.00	cm
$G/\sqrt{Ac}$			2.55	
Wa (effective) /Wa			0.911	
A_T	15.61	in^2	100.7	cm^2
D	0.500	in	1.27	cm
E	0.437	in	1.11	cm
F	0.500	in	1.27	cm
G	1.125	in	2.857	cm
BOBBIN	DORCO ELECTRONICS *		1-L-12	
LENGTH	1.08	in	2.74	cm
BUILD	0.476	in	1.21	cm
* Wa (effective)	0.513	in^2	3.31	cm^2
BRACKET	HALLMARK METALS *		08-106-07	



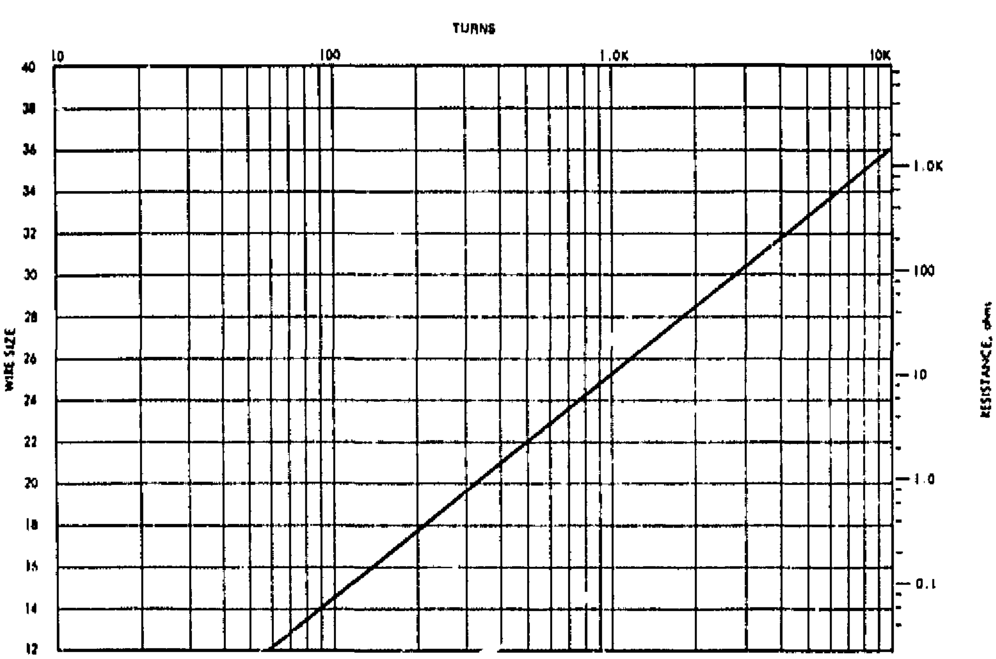


Fig. 4.B-9. Wiregraph for "C" core AL-12

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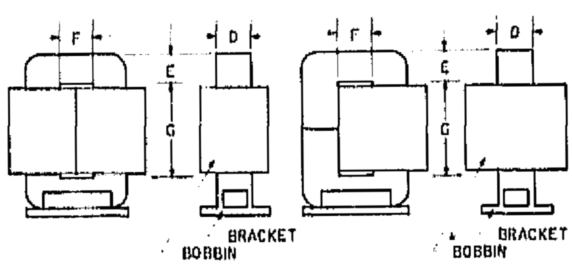
Table 4.B-10. "C" core AL-135

"C" CORE	AL 135			
	ENGLISH		METRIC	
Wa/Ac			2.89	
Wa x Ac	0.123	in^4	5.14	cm^4
Wa	0.633	in^2	4.083	cm^2
Ac (effective)	0.195	in^2	1.26	cm^2
lm	4.648	in	11.8	m
CORE WT	0.251	lb	114	grams
COPPER WT	0.312	lb	169	grams
* MLT FULLWOUND	3.74	in	9.50	cm
$G/\sqrt{Ac}$			2.55	
Wa (effective) /Wa			0.015	
A_T	17.04	in^2	110	cm^2
D	0.500	in	1.27	cm
E	0.437	in	1.11	cm
F	0.562	in	1.43	cm
G	1.125	in	2.857	cm
BOBBIN	DORCO ELECTRONICS *		1 L-135	
LENGTH	1.08	in	2.74	cm
BUILD	0.537	in	1.36	cm
* Wa (effective)	0.579	in^2	3.74	cm^2
BRACKET	HALLMARK METALS *		08-107-07	



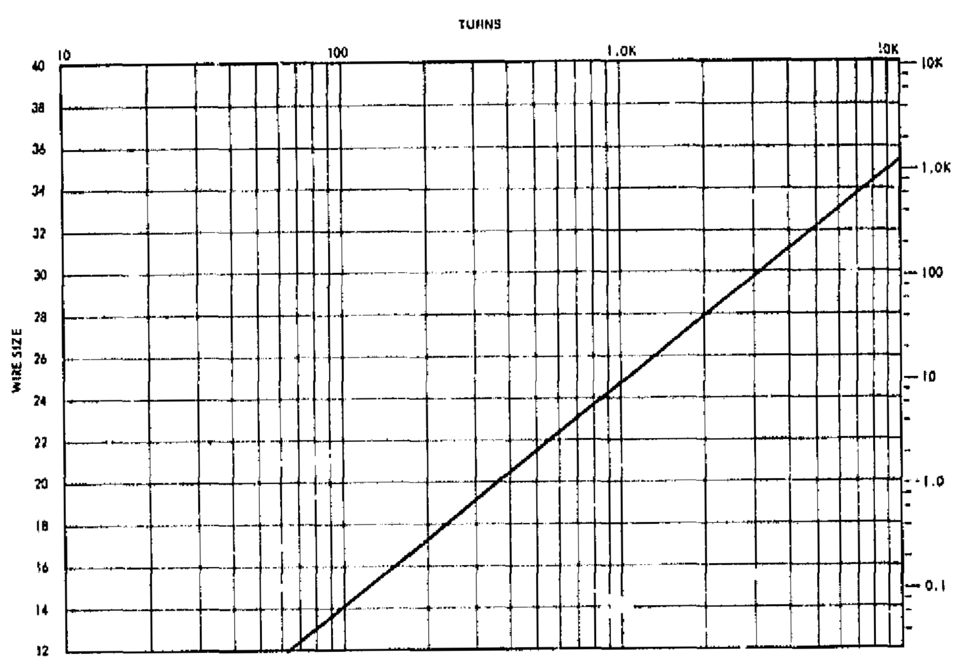
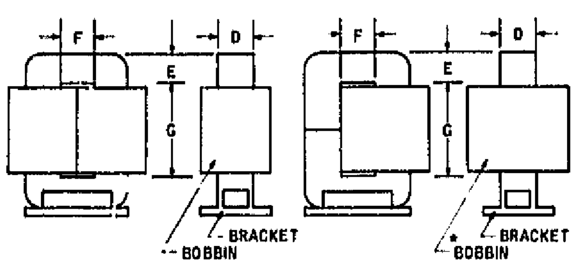


Fig. 4.B-10. Wiregraph for "C" core AL-135

Table 4.B-11. "C" core AL-78

"C" CORE	AL-78			
	ENGLISH		METRIC	
Wa/Ac			3.00	
Wa x Ac	0.146	in^4	6.07	cm^4
Wa	0.703	in^2	4.53	cm^2
Ac (effective)	0.208	in^2	1.34	cm^2
lm	5.891	in	14.96	cm
CORE WT	0.342	lb	164	grams
COPPER WT	0.331	lb	150	grams
* MLT FULLWOUND	3.21	in	8.15	cm
$G/\sqrt{Ac}$			4.93	
Wa (effective) /Wa			0.905	
A_T	16.99	in^2	109.6	cm^2
D	0.750	in	1.91	cm
E	0.313	in	0.785	cm
F	0.313	in	0.795	cm
G	2.250	in	5.715	cm
BOBBIN	DORCO ELECTRONICS #		1-L-78	
LENGTH	2.205	in	5.60	cm
BUILD	0.288	in	0.731	cm
* Wa (effective)	0.635	in^2	4.10	cm^2
BRACKET	HALLMARK METALS #		012-016-05	



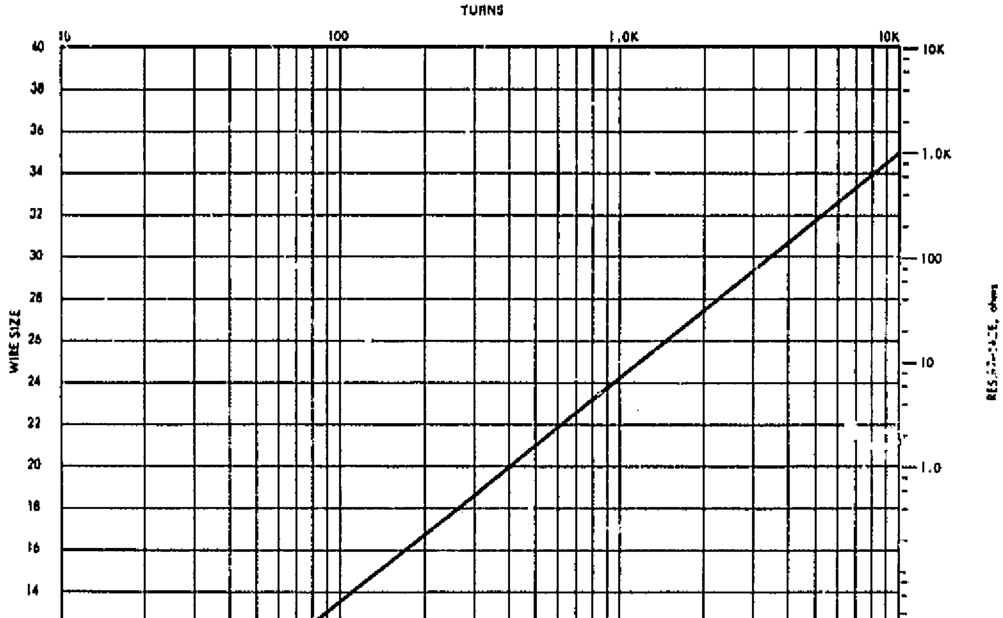
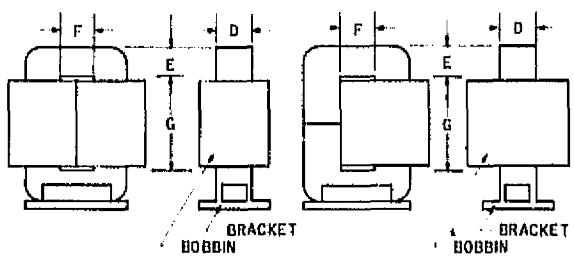


Fig. 4.B-11. Wiregraph for "C" core AL-78

Table 4.B-12. "C" core AL-18

"C" CORE	AL-18	
	ENGLISH	METRIC
Wa/Ac		6.08
Wa x Ac	0.183 in^4	7.87 cm^4
Wa	0.077 in^2	6.30 cm^2
Ac (effective)	0.104 in^2	1.257 cm^2
lm	5.648 in	14.34 cm
CORE WT	0.305 lb	138 grams
COPPER WT	0.575 lb	260 grams
* MLT FULLWOUND	2.96 in	7.51 cm
$G/\sqrt{Ac}$		3.502
Wa (effective) /Wa		0.890
A_T	21.93 in^2	141.60 cm^2
D	0.500 in	1.27 cm
E	0.437 in	1.111 cm
F	0.625 in	1.587 cm
G	1.562 in	3.927 cm
BOBBIN	DORCO ELECTRONICS * 1-L-18	
LENGTH	1.497 in	3.802 cm
BUILD	0.590 in	1.498 cm
* Wa (effective)	0.880 in^2	5.697 cm^2
BRACKET	HALLMARK METALS * 08-108-07	



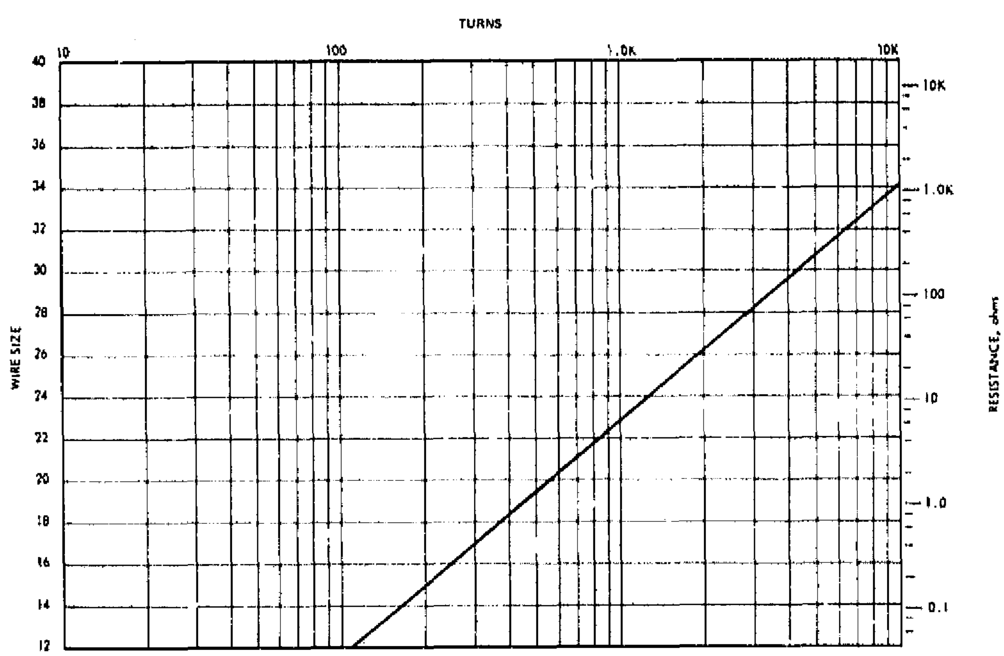


Fig. 4.B-12. Wiregraph for "C" core AL-18

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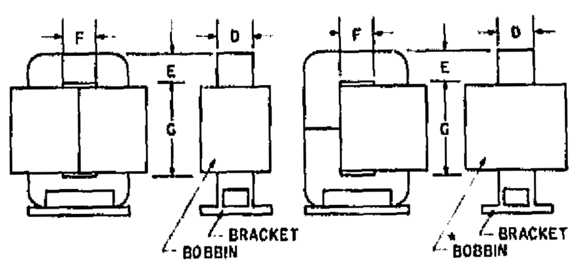
Table 4.B-13. "C" core AL-15

"C" CORE	AL-15			
	ENGLISH		METRIC	
Wa/Ac			2.50	
Wa x Ac	0.218	in^4	9.07	cm^4
Wa	0.781	in^2	5.037	cm^2
Ac (effective)	0.279	in^2	1.80	cm^2
lm	5.508	in	14.2	cm
CORE WT	0.438	lb	197	grams
COPPER WT	0.448	lb	203	grams
* MLT FULLWOUND	3.97	in	10.08	cm
$G/\sqrt{Ac}$			2.86	
Wa (effective) /Wa			0.891	
A_T	21.07	in^2	135.9	cm^2
D	0.625	in	1.587	cm
E	0.500	in	1.27	cm
F	0.500	in	1.27	cm
G	1.562	in	3.967	cm
BOBBIN	DORCO ELECTRONICS #		1-L-15	
LENGTH	1.497	in	3.80	cm
BUILD	0.465	in	1.18	cm
* Wa (effective)	0.696	in^2	4.49	cm^2
BRACKET	HALLMARK METALS #		010-108-08	



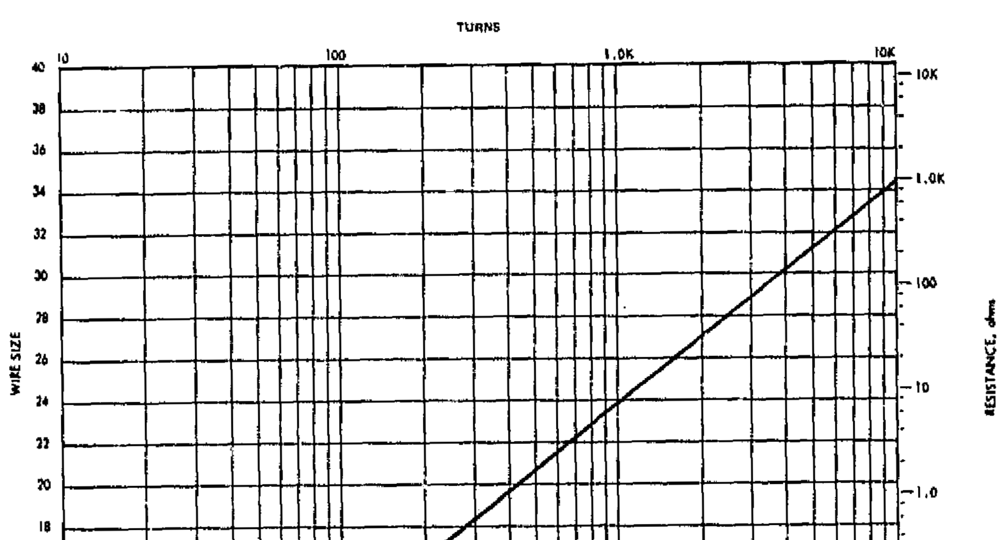
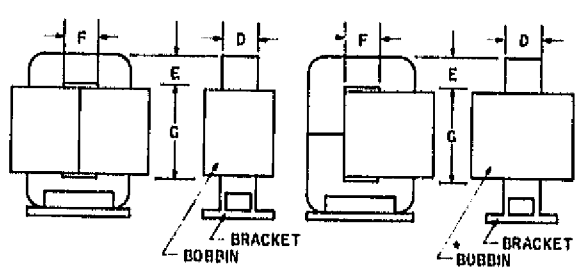


Fig. 4.B-13. Wiregraph for "C" core AL-15

Table 4.B-14. "C" core AL-16

"C" CORE	AL-16			
	ENGLISH		METRIC	
Wa/Ac			2.08	
Wa x Ac	0.26	in^4	10.8	cm^4
Wa	0.781	in^2	5.037	cm^2
Ac (effective)	0.334	in^2	2.15	cm^2
lm	5.588	in	14.2	cm
CORE WT	0.519	lb	235	grams
COPPER WT	0.476	lb	216	grams
* MLT FULLWOUND	4.22	in	10.72	cm
$G/\sqrt{Ac}$			2.70	
Wa (effective) /Wa			0.891	
A_T	22.21	in^2	143.3	cm^2
D	0.750	in	1.905	cm
E	0.500	in	1.27	cm
F	0.500	in	1.27	cm
G	1.562	in	3.967	cm
BOBBIN	DORCO ELECTRONICS * 1-1-16			
LENGTH	1.497	in	3.80	cm
BUILD	0.465	in	1.18	cm
* Wa (effective)	0.696	in^2	4.49	cm^2
BRACKET	HALLMARK METALS * 012-108-08			



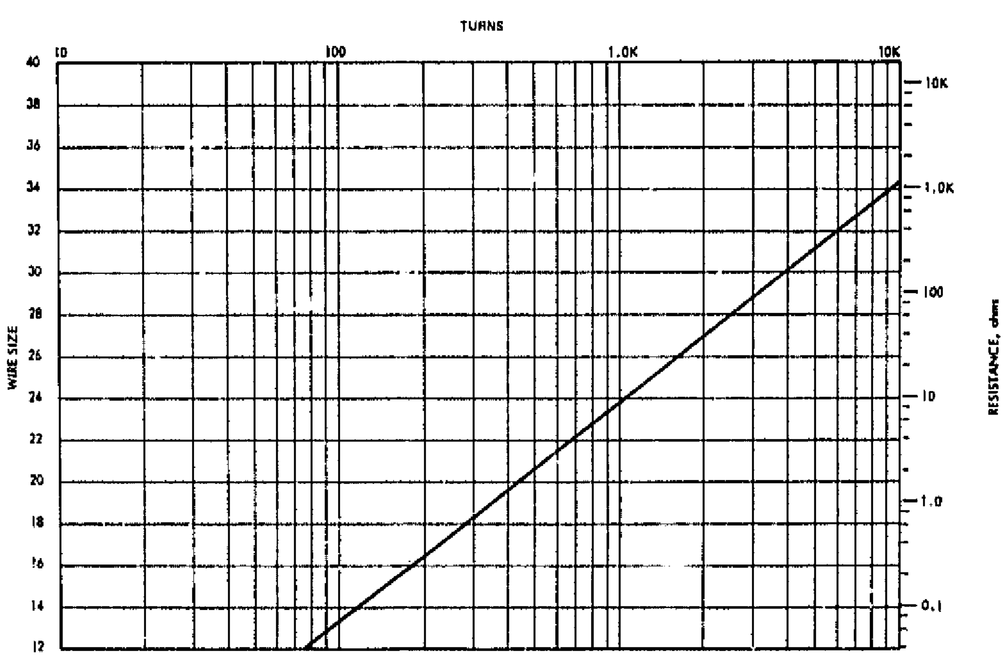
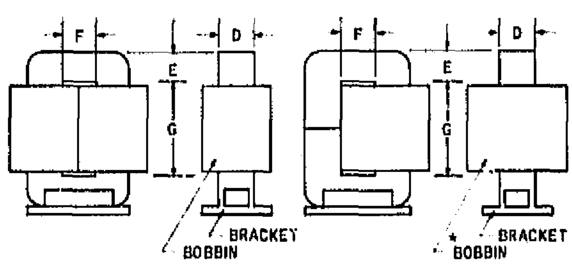


Fig. 4.B-14. Wiregraph for "C" core AL-16

Table 4.B-15. "C" core AL-17

"C" CORE	AL-17			
	ENGLISH		METRIC	
Wa/Ac			1.66	
Wa x Ac	0.35	in^4	14.4	cm^4
Wa	0.781	in^2	6.037	cm^2
Ac (effective)	0.445	in^2	2.870	cm^2
lm	5.688	in	14.2	cm
CORE WT	0.693	lb	314	grams
COPPER WT	0.533	lb	241	grams
* MLT FULLWOUND	4.72	in	11.99	cm
$G/\sqrt{Ac}$			2.342	
Wa (effective) /Wa			0.801	
A_T	24.5	in^2	158	cm^2
D	1.000	in	2.54	cm
E	0.500	in	1.27	cm
F	0.500	in	1.27	cm
G	1.562	in	3.967	cm
BOBBIN	DORCO ELECTRONICS • I-L-17			
LENGTH	1.497	in	3.80	cm
BUILD	0.466	in	1.18	cm
* Wa (effective)	0.696	in^2	4.49	cm^2
BRACKET	HALLMARK METALS • 10-108 08			



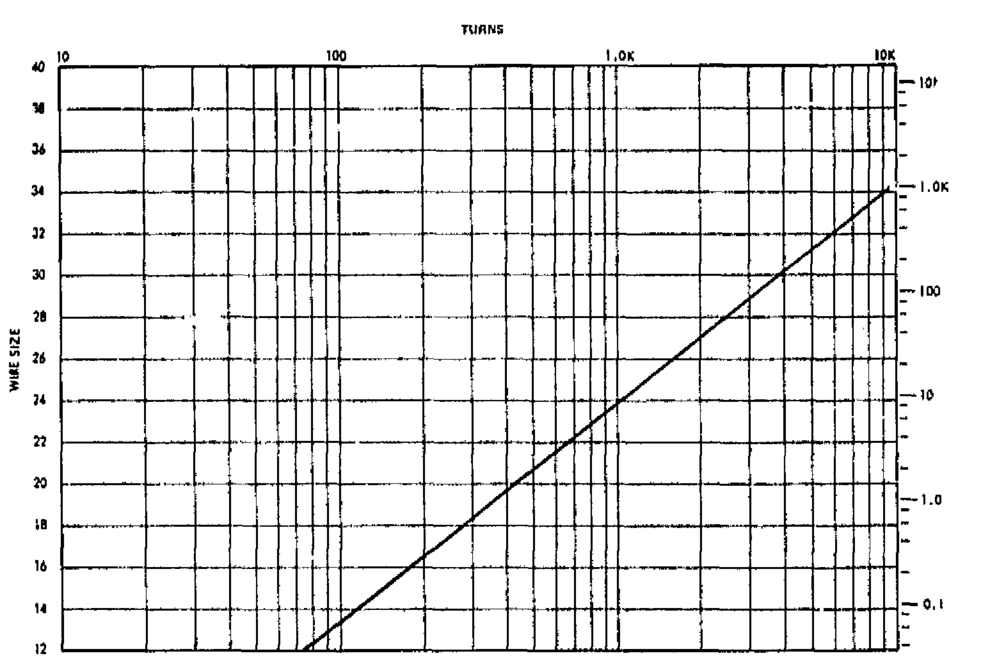


Fig. 4.B-15. Wiregraph for "C" core AL-17

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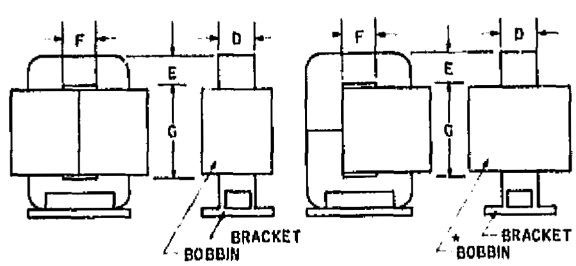
Table 4.B-16. "C" core AL-19

"C" CORE	AL-19			
	ENGLISH		METRIC	
Wa/Ac			1.96	
Wa x Ac	0.435	in^4	18.1	cm^4
Wa	0.877	in^2	6.30	cm^2
Ac (effective)	0.445	in^2	2.87	cm^2
lm	5.838	in	14.8	cm
CORE WT	0.724	lb	328	grams
COPPER WT	0.731	lb	332	grams
* MLT FULLWOUND	5.11	in	12.98	cm
$G/\sqrt{Ac}$			2.34	
Wa (effective) /Wa			0.903	
A_T	28.2	in^2	182	cm^2
D	1.000	in	2.54	cm
E	0.500	in	1.27	cm
F	0.626	in	1.587	cm
G	1.562	in	3.967	cm
BOBBIN	DORCO ELECTRONICS • L-L-19			
LENGTH	1.497	in	3.80	cm
BUILD	0.590	in	1.498	cm
* Wa (effective)	0.883	in^2	5.69	cm^2
BRACKET	HALLMARK METALS • 10-110-08			



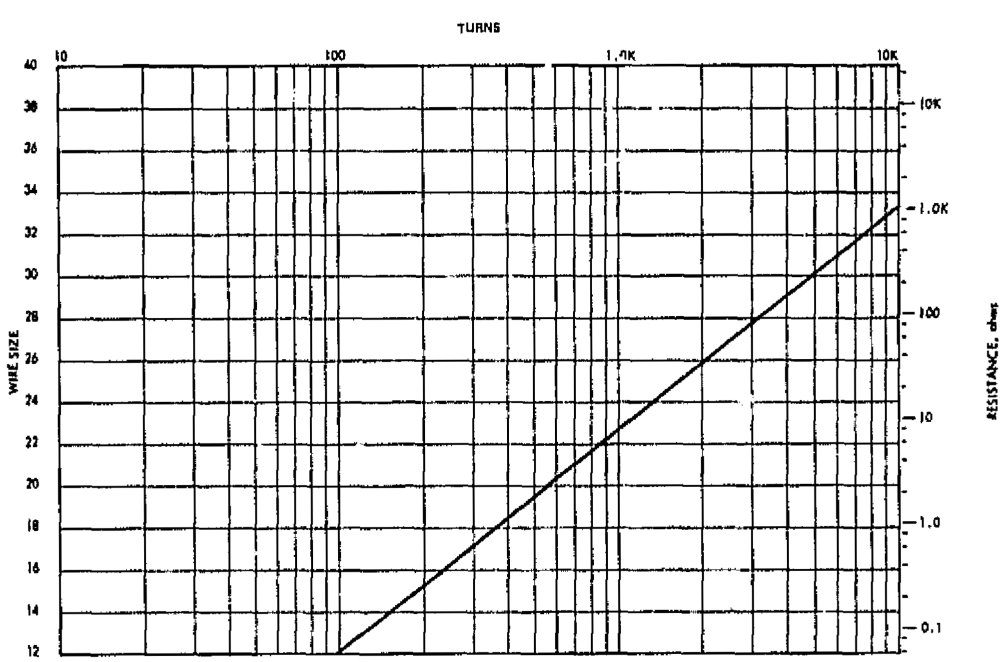
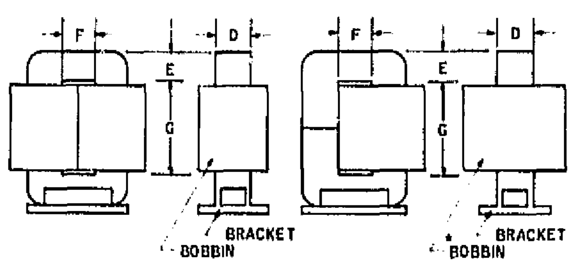


Fig. 4.B-16. Wiregraph for "C" core AL-19

Table 4.B-17. "C" core AL-20

"C" CORE	AL-20			
	ENGLISH		METRIC	
Wa/Ac			1.68	
Wa x Ac	0.543	in^4	22.6	cm^4
Wa	0.977	in^2	6.30	cm^2
Ac (effective)	0.556	in^2	3.58	cm^2
lm	6.228	in	16.8	cm
CORE WT	0.965	lb	437	grams
COPPER WT	0.767	lb	348	grams
* MLT FULLWOUND	5.36	in	13.62	cm
$G/\sqrt{Ac}$			2.09	
Wa (effective)/Wa			0.903	
A_T	31.7	in^2	205	cm^2
D	1.000	in	2.54	cm
E	0.625	in	1.587	cm
F	0.625	in	1.587	cm
G	1.562	in	3.967	cm
BOBBIN	DORCO ELECTRONICS • 1-L-20			
LENGTH	1.497	in	3.80	cm
BUILD	0.590	in	1.498	cm
* Wa (effective)	0.883	in^2	5.69	cm^2
BRACKET	HALLMARK METALS • 10-114-010			



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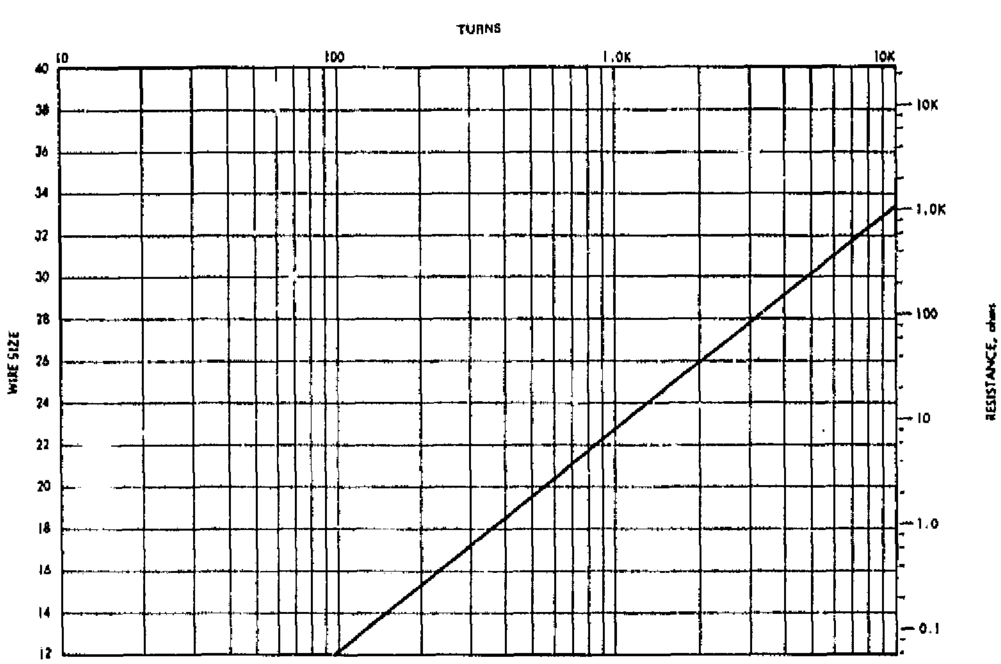
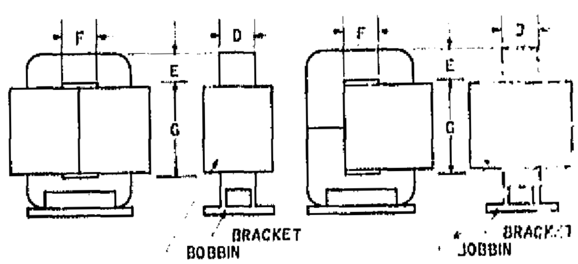


Fig. 4.B-17. Wiregraph for "C" core AL-20

Table 4.B-18. "C" core AL-22

"C" CORE	AL-22			
	ENGLISH		METRIC	
Wa/Ac			1.04	
Wa x Ac	0.692	in^4	28.0	cm^4
Wa	1.21	in^2	7.804	cm^2
Ac (effective)	0.556	in^2	3.58	cm^2
lm	6.978	in	17.2	cm
CORE WT	1.08	lb	480	grams
COPPER WT	0.961	lb	436	grams
* MLT FULLWOUND	5.36	in	13.62	cm
$G/\sqrt{Ac}$			2.598	
Wa (effective) /Wa			0.912	
A_T	35.3	in^2	228	cm^2
D	1.000	in	2.54	cm
E	0.625	in	1.587	cm
F	0.625	in	1.587	cm
G	1.937	in	4.92	cm
BOBBIN	DORCO ELECTRONICS *		14-22	
LENGTH	1.872	in	4.75	cm
BUILD	0.590	in	1.498	cm
* Wa (effective)	1.10	in^2	7.12	cm^2
BRACKET	HALLMARK METALS *		10 114-010	



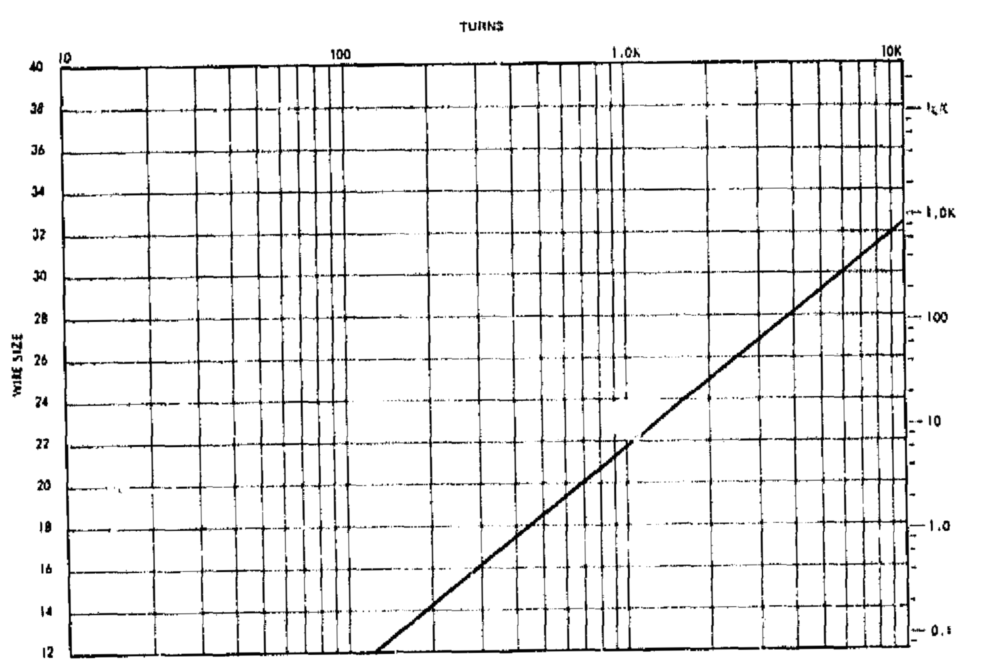
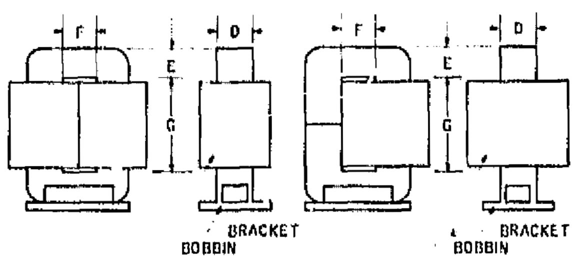


Fig. 4.B-18. Wiregraph for "C" core AL-22

Table 4.B-19. "C" core AL-23

"C" CORE	AL 23			
	ENGLISH		METRIC	
Wa/Ac			1.55	
Wa x Ac	0.841	in^4	34.96	cm^4
Wa	1.21	in^2	7.804	cm^2
Ac (effective)	0.695	in^2	4.48	cm^2
lm	6.978	in	17.2	cm
CORE WT	1.352	lb	612	grams
COPPER WT	1.056	lb	479	grams
* MLT FULLWOUND	5.86	in	14.89	cm
$G/\sqrt{Ac}$			2.37	
Wa (effective) /Wa			0.912	
A_T	38.1	in^2	246	cm^2
D	1.250	in	3.175	cm
E	0.625	in	1.587	cm
F	0.625	in	1.587	cm
G	1.937	in	4.92	cm
BOBBIN	GORCO ELECTRONICS •		TL-23	
LENGTH	1.872	in	4.75	cm
BUILD	0.690	in	1.408	cm
* Wa (effective)	1.10	in^2	7.12	cm^2
BRACKET	HALLMARK METALS •		14-114-010	



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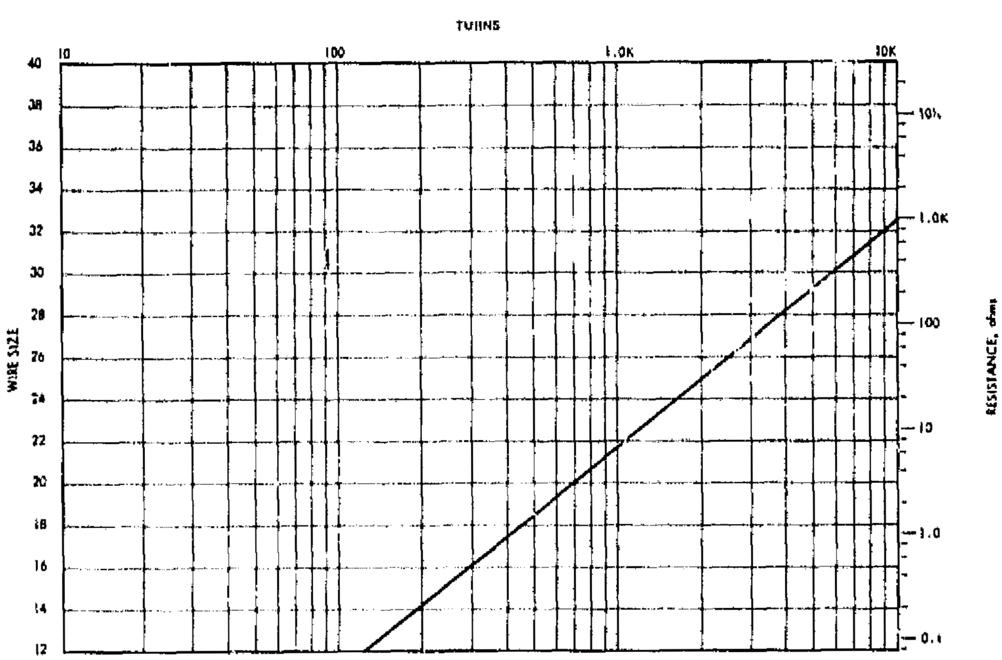
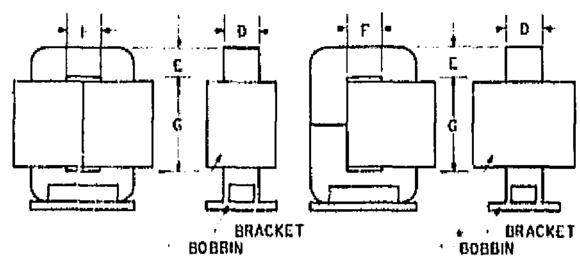


Fig. 4.B-19. Wiregraph for "C" core AL-23

Table 4.B-20. "C" core AL-24

"C" CORE	AL-24			
	ENGLISH		METRIC	
Wa/Ac			2.77	
Wa x Ac	0.962	in^4	40.0	cm^4
Wa	1.73	in^2	11.10	cm^2
Ac (effective)	0.550	in^2	3.58	cm^2
lm	7.871	in	20.0	cm
CORE WT	1.220	lb	552	grams
COPPER WT	1.501	lb	68	grams
* MLT FULLWOUND	5.76	in	1[illegible].02	cm
$G/\sqrt{Ac}$			[illegible].10	
Wa (effective) /Wa			0.920	
A_T	43.6	in^2	281.6	cm^2
D	1.000	in	2.54	cm
E	0.625	in	1.587	cm
F	0.750	in	1.905	cm
G	2.313	in	5.875	cm
BOBBIN	DORCO ELECTRONICS # 1-L-24			
LENGTH	2.248	in	5.709	cm
BUILD	0.715	in	1.810	cm
* Wa (effective)	1.607	in^2	10.37	cm^2
BRACKET	HALLMARK METALS # 10-200-010			



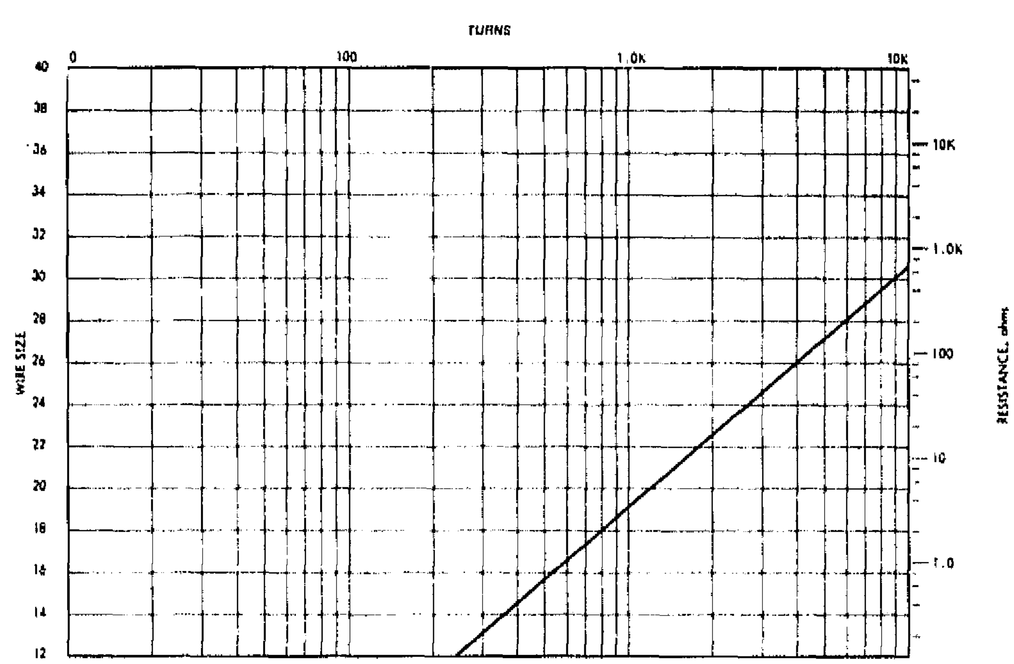


Fig. 4.B-20. Wiregraph for "C" core AL-24

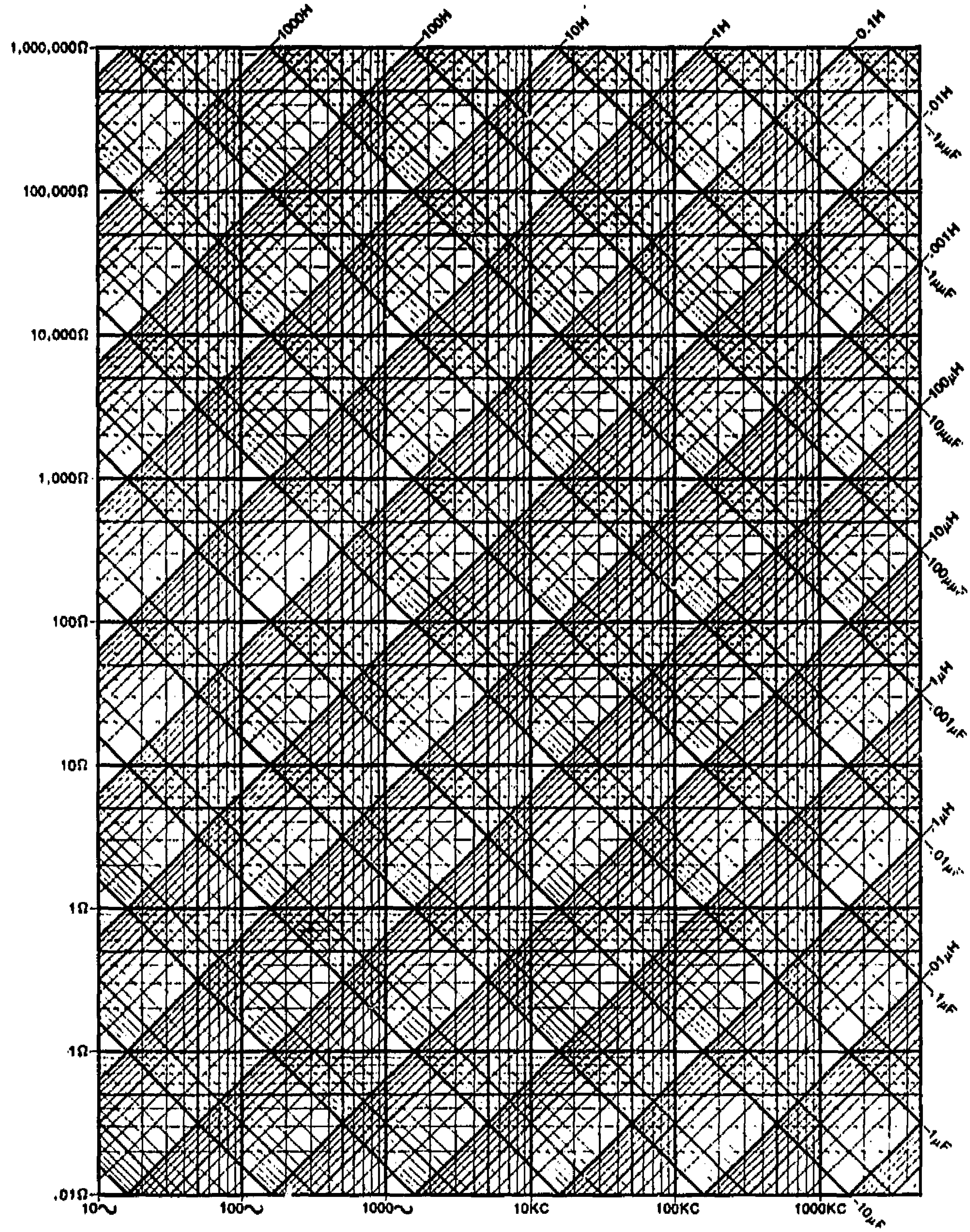


Fig. 4.B-21. Graph for inductance, capacitance, and reactance

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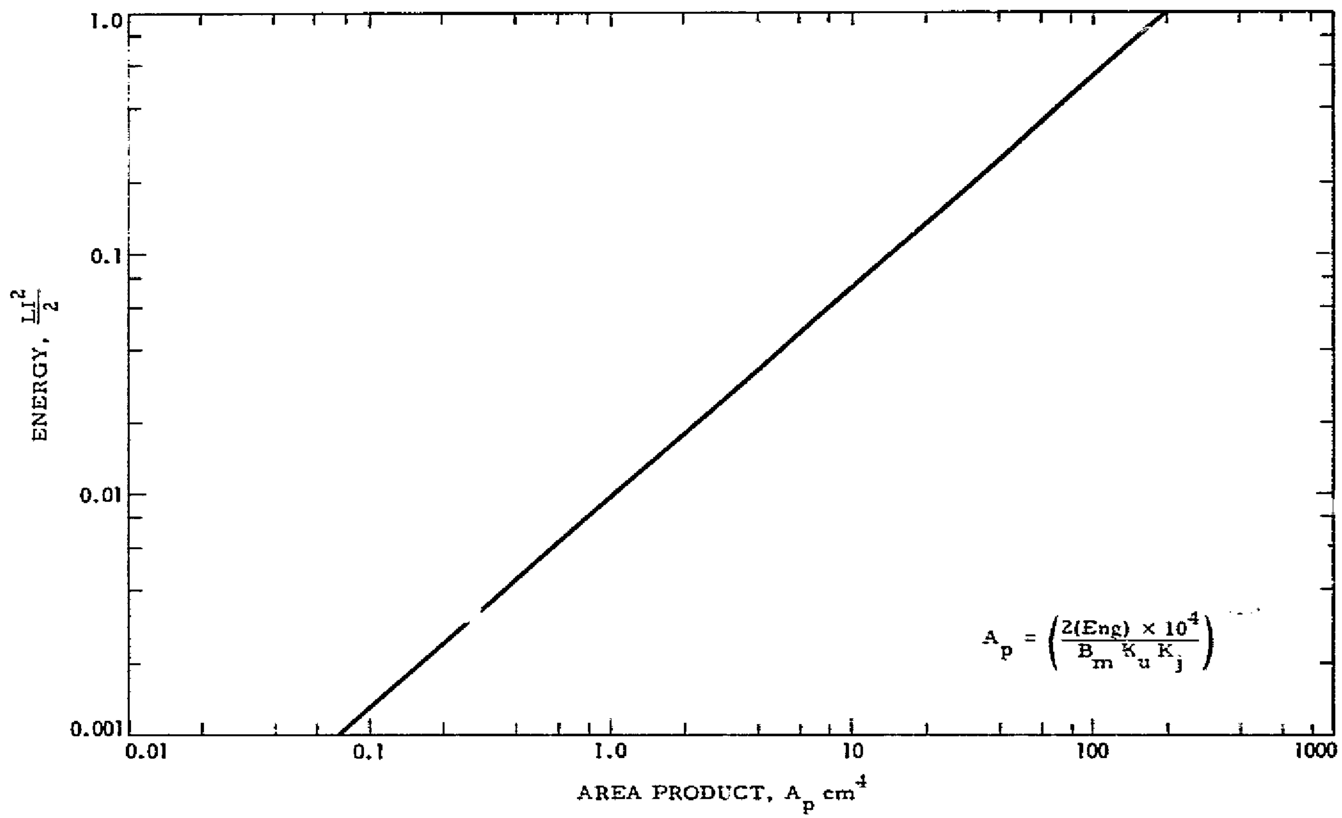

Fig. 4.B-22. Area product vs energy $\frac{LI^2}{2}$

B_m = 1.2 (tesla)
K_u = 0.4
K_j = 395

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CHAPTER V

TOROIDAL POWDER CORE SELECTION WITH dc CURRENT

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A. INTRODUCTION

Inductors which carry direct current are used frequently in a wide variety of ground, air, and space applications. Selection of the best magnetic core for an inductor frequently involves a trial-and-error type of calculation.

The design of an inductor also frequently involves consideration of the effect of its magnetic field on other devices near where it is placed. This is especially true in the design of high-current inductors for converters and switching regulators used in spacecraft, which may also employ sensitive magnetic field detectors. For this type of design problem it is frequently imperative that a toroidal core be used. The magnetic flux in a moly-permalloy toroid (core) can be contained inside the core more readily than in a lamination or C type core, as the winding covers the core along the whole magnetic path length.

The author has developed a simplified method of designing optimum dc carrying inductors with moly-permalloy powder cores. This method allows the correct core permeability to be determined without relying on trial and error.

B. RELATIONSHIP OF A_p TO INDUCTOR'S ENERGY HANDLING CAPABILITY

According to the newly developed approach, the energy-handling capability of a core is related to its area product A_p:

$$A_p = \left(\frac{2(Eng) \times 10^4}{B_m K_u K_j}\right)^{1.14} \qquad [cm^4] \quad (5\text{-}1)$$

where:

K_j = current density coefficient (see Chapter 2)

K_u = window utilization factor (see Chapter 6)

B_m = flux density, tesla

Eng = energy, watt seconds

From the above, it can be seen that factors such as flux density, window utilization factor K_u (which defines the maximum space that may be occupied by the copper in the window), and the constant K_j (which is related to temperature rise) all have an influence on the inductor area product. The constant K_j is a new parameter that gives the designer control of the copper losses. Derivation is set forth in detail in Chapter 2. The energy-handling capability of a core is derived from

$$Eng = \frac{LI^2}{2} \qquad \text{[watt second]} \quad (5\text{-}2)$$

III. FUNDAMENTAL CONSIDERATIONS

The design of a linear reactor depends upon four related factors:

1. Desired inductance
2. Direct current
3. Alternating current ΔI
4. Power loss and temperature rise

With these requirements established, the designer must determine the maximum values for B_{dc} and for B_{ac} which will not produce magnetic saturation, and must make tradeoffs which will yield the highest inductance for a given volume. The core permeability chosen dictates the maximum dc flux density which can be tolerated for a given design. Permeability values for different powder cores are shown in Table 5-1.

Table 5-1. Different powder core permeabilities

Permeability	Amp turn/cm with dc bias $L < 80\%$
14	253
36	140
60	56
125	28
147	23
160	20
173	19
200	16
300	11
555	4

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If an inductance is to be constant with increasing direct current, there must be a negligible drop in inductance over the operating current range. The maximum H, then, is an indication of a core's capability. In terms of ampere-turns and mean magnetic path length l_m,

$$H = \frac{NI}{l_m} \qquad \text{[amp turn/cm]} \quad (5\text{-}3)$$

$$NI = 0.8\, H l_m \qquad \text{[amp turn]} \quad (5\text{-}4)$$

Inductance decreases with increasing flux density and magnetizing force for various materials of different values of permeability μ_Δ. The selection of the correct permeability for a given design is made using equation 5-4 after solving for the area product A_p:*

$$\mu_\Delta = \frac{B_m l_m \times 10^4}{0.4\pi\, W_a J K_u} \quad (5\text{-}5)$$

It should be remembered that maximum flux density depends upon $B_{dc} + B_{ac}$ in the manner shown in Fig. 5-1.

$$B_m = B_{dc} + B_{ac} \qquad \text{[tesla]} \quad (5\text{-}6)$$

$$B_{dc} = \frac{0.4\pi\, NI_{dc} \times 10^{-4}}{\frac{l_m}{\mu_\Delta}} \qquad \text{[tesla]} \quad (5\text{-}7)$$

*Derivation is set forth in detail in Appendix 5.A at the end of this Chapter.

$$B_{ac} = \frac{0.4\pi \, N \frac{\Delta I}{2} \times 10^{-4}}{\frac{l_m}{\mu_\Delta}} \qquad [\text{tesla}] \quad (5\text{-}8)$$

Combining Eqs. (5-7) and(5-8),

$$B_m = \frac{0.4\pi \, N I_{dc} \times 10^{-4}}{\frac{l_m}{\mu_\Delta}} + \frac{0.4\pi \, N \frac{\Delta I}{2} \times 10^{-4}}{\frac{l_m}{\mu_\Delta}} \qquad [\text{tesla}] \quad (5\text{-}9)$$

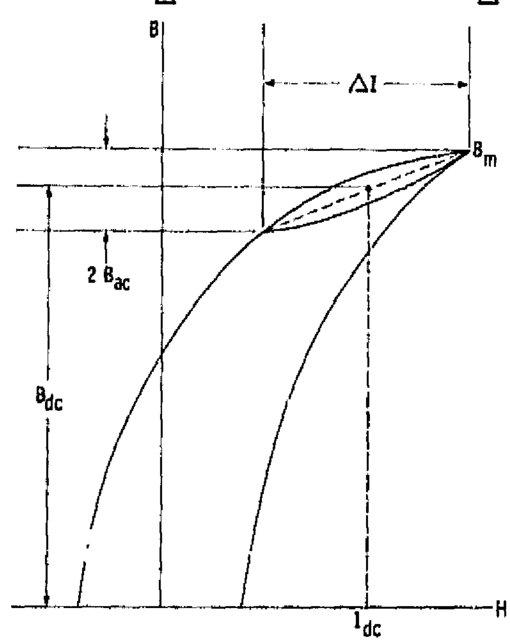

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Fig. 5-1. Flux density versus $I_{dc} + \Delta I$

Moly-permalloy powder cores operating with a dc bias of 0.3 tesla have only about 80% of their original inductance, with very rapid falloff at higher densities as shown in Fig. 5-2.

The flux density for the initial design for moly-permalloy powder cores should be limited to 0.2 tesla maximum for B_{dc} plus B_{ac}.

The losses in a moly-permalloy inductor due to ac flux density are very low compared to the steady state dc copper loss. It is then assumed that the majority of the losses are copper:

$$P_{cu} \gg P_{fe} \qquad (5\text{-}10)$$

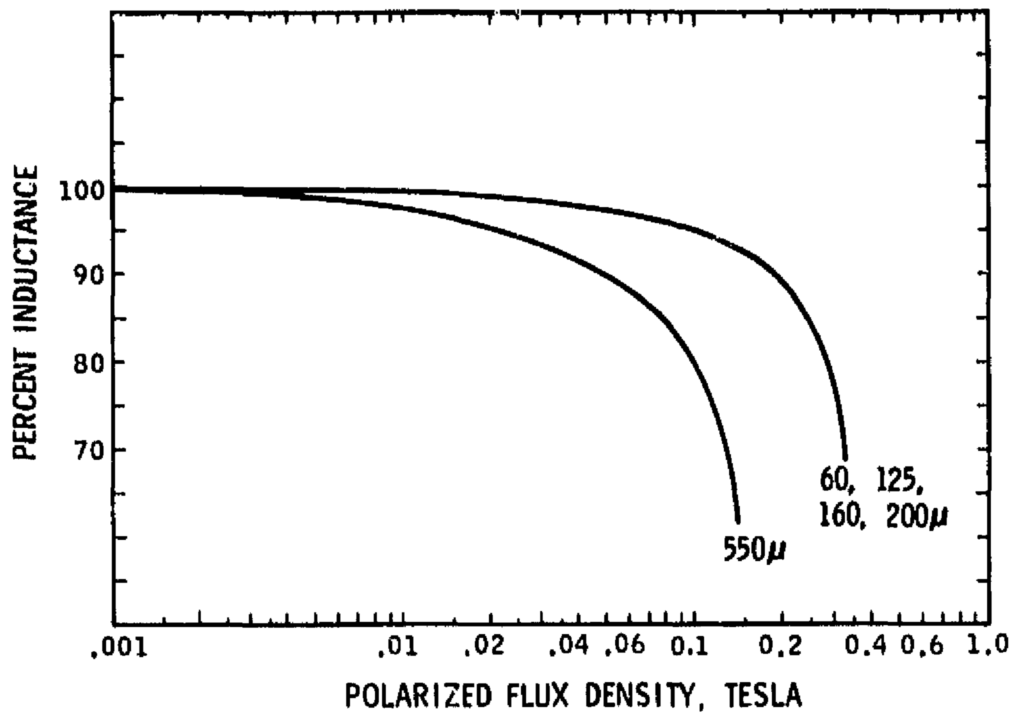

Fig. 5-2. Inductance versus dc bias

D. A SPECIFIED DESIGN PROBLEM AS AN EXAMPLE

For a typical design example, assume the following:

(1) Inductance 0.0015 henry

(2) dc current 2 amperes

(3) 25°C rise

The procedure would be as shown below.

Step No. 1. Calculate the energy-handling capability from equation 5-2:

$$\text{Energy} = \frac{LI^2}{2} \qquad [\text{watt second}]$$

$$\text{Energy} = \frac{(0.0015)\,(2)^2}{2}$$

$$\text{Energy} = 0.003 \qquad [\text{watt second}]$$

Step No. 2. Calculate the area product A_p from equation 5-1:

$$A_p = \left(\frac{2(\text{Energy}) \times 10^4}{B_m K_u K_j}\right)^{1.14} \quad [cm^4]$$

$$B_m = 0.2 \quad [\text{tesla}]$$

$$K_u = 0.4$$

$$K_j = 403$$

$$A_p = \left(\frac{2(0.003) \times 10^4}{(0.2)\ (0.4)\ (403)}\right)^{1.14} \quad [cm^4]$$

or

$$A_p = 2.03 \quad [cm^4]$$

After the A_p has been determined, the geometry of the inductor can be evaluated as described in Chapter 2 for weight, for surface area, and for volume, and appropriate changes made, if required.

Step No. 3. Select a powder core from Table 2-2 with a value of A_p closest to the one calculated:

55071 with an A_p = 1.966 $[cm^4]$

For more information, see Table 5.B-6.

Step No. 4. Calculate the current density J from equation 5.A-19:

$$J = K_j A_p^{-0.12} \quad [A/cm]$$

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The value for K_j is found in Table 2-1:

$$J = (403)(1.966)^{-0.12}$$

$$J = 372 \qquad [A/cm]$$

Step No. 5. Calculate the permeability of the core required from equation 5.A-24:

$$\mu_{\Delta} = \frac{B_m l_m \times 10^4}{0.4\pi W_a J K_u}$$

(see Table 5.B-6.)

$$\mu_{\Delta} = \frac{(0.2)(8.15) \times 10^4}{(1.25)(2.93)(372)(0.4)}$$

$$\mu_{\Delta} = 38$$

From the manufacturer's catalog, the core that has the same size but has a permeability closer to the one calculated is the core 55550, with a permeability of 26. This particular core has 28 millihenry per 1000 turns.

Step No. 6. Calculate the number of turns required for 1.5 millihenry.

$$N = 1000 \sqrt{\frac{L}{L_{1000}}}$$

L = inductance

L_{1000} = inductance at 1000 turns

$$N = 1000 \sqrt{\frac{1.5}{28}}$$

$$N = 231$$

Step No. 7. Calculate the bare wire size $A_{w(B)}$:

$$A_{w(B)} = I/J \qquad [cm^2]$$

$$A_{w(B)} = 2.0/372$$

$$A_{w(B)} = 0.00537 \qquad [cm^2]$$

Step No. 8. Select the wire area A_w in Table 6-1 for equivalent (AWG) wire size, column A:

AWG No. 20 = 0.005188

Step No. 9. Calculate the resistance of the winding, using Table 6-1, column C, and Table 2-2, column 4, for the MLT:

$$R = MLT \times N \times (\text{column C}) \times \zeta \times 10^{-6} \qquad [\Omega]$$

$$R = (4.77)(231)(332)(1.098) \times 10^{-6}$$

$$R = 0.402 \qquad [\Omega]$$

Step No. 10. Calculate the copper loss:

$$P_{cu} = I^2 R \qquad [\text{watts}]$$

$$P_{cu} = (2)^2 (0.402)$$

$$P_{cu} = 1.608 \qquad [\text{watts}]$$

From chapter 7, the surface area A_t required to dissipate waste heat (expressed as watts loss per unit area) is:

$$A_t = \frac{P_\Sigma}{\psi}$$

$$P_\Sigma = P_{cu}$$

$$\psi = 0.03 \text{ W/cm}^2 \text{ at } 25^\circ\text{C rise}$$

Referring to Table 2-2, column 2, for the 55071 size core, the surface area A_t is 44.7 cm^2:

$$\psi = \frac{P_\Sigma}{A_t}$$

$$\psi = \frac{1.608}{44.7}$$

$$\psi = 0.036 \qquad \left[\text{W/cm}^2\right]$$

which will produce the required temperature rise.

(In a test sample made to prove out this example, the measured inductance was found to be 0.0015 hy with a resistance of 0.36 ohms at 25°C and 0.388 ohms at 45°C.)

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Blinchikoff, H., Toroidal Inductor Design Electro-Technology. November 1964, Page 42-50.

APPENDIX 5.A

TOROID POWDER CORE SELECTION WITH dc CURRENT

After calculating the inductance and dc current, select the proper permeability and size of powder core with a given $LI^2/2$. The energy-handling capability of an inductor can be determined by its A_p product, of which W_a is the available core window area in cm^2 and A_c is the core effective cross sectional area in cm^2. The $W_a A_c$ or area product A_p relationship is obtained by solving $E = L dI/dt$ as follows:*

$$E = L \frac{dI}{dt} = N \frac{d\phi}{dt} \qquad (5.A-1)$$

$$L = N \frac{d\phi}{dI} \qquad (5.A-2)$$

$$\phi = B_m A'_c \qquad (5.A-3)$$

$$B_m = \mu_\Delta \mu_o H = \frac{\mu_\Delta \mu_o NI}{l'_m} \qquad (5.A-4)$$

$$\phi = \frac{\mu_\Delta \mu_o NI A'_c}{l'_m} \qquad (5.A-5)$$

$$\frac{d\phi}{dI} = \frac{\mu_\Delta \mu_o N A'_c}{l'_m} \qquad (5.A-6)$$

$$L = N \frac{d\phi}{dI} = \frac{\mu_\Delta \mu_o N^2 A'_c}{l'_m} \qquad (5.A-7)$$

$$\text{Energy} = \frac{LI^2}{2} = \frac{\mu_r \mu_o N^2 A'_c I^2}{l'_m{}^2} \qquad (5.A-8)$$

---

*Primes indicate measurements in the mks system.

If B_m is specified,

$$I = \frac{B_m l'_m}{\mu_\Delta \mu_o N} \qquad (5.A\text{-}9)$$

$$Eng = \frac{\mu_\Delta \mu_o N^2 A'_c}{{l'_m}^2} \left(\frac{B_m l'_m}{\mu_\Delta \mu_o N} \right)^2 \qquad (5.A\text{-}10)$$

Reducing to

$$Eng = \frac{B_m l'_m A'_c}{2 \mu_\Delta \mu_o} \quad \left[\text{watt seconds} \right] \qquad (5.A\text{-}11)$$

$$I = \frac{K_u W'_a J'}{N} = \frac{B_m l'_m}{\mu_\Delta \mu_o N} \qquad (5.A\text{-}12)$$

Solving for $\mu_\Delta \mu_o$,

$$\mu_\Delta \mu_o = \frac{B_m l'_m}{K_u W'_a J'} \qquad (5.A\text{-}13)$$

Substituting into the energy equation,

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$$Eng = \frac{B_m^2 l'_m A'_c}{2} \cdot \frac{K_u W'_a J'}{B_m l'_m} = \frac{W'_a A'_c B_m J' K_u}{2} \qquad (5.A\text{-}14)$$

let

$$l'_m = l_m \times 10^{-2}$$

$$W'_a = W_a \times 10^{-4}$$

$$A'_c = A_c \times 10^{-4}$$

$$J' = J \times 10^{4}$$

Substituting into the energy equation,

$$\text{Eng} = \frac{W_a A_c B_m J K_u}{2} \times 10^{-4} \qquad (5.A\text{-}15)$$

Solving for $W_a A_c$,

$$W_a A_c = \frac{2 \text{ Eng} \times 10^4}{K_u B_m J} \qquad (5.A\text{-}16)$$

and since the area product is

$$A_p = W_a A_c \qquad (5.A\text{-}17)$$

then

$$A_p = \frac{2 \text{ (Energy)} \times 10^4}{K_u B_m J} \qquad (5.A\text{-}18)$$

Combining the equation from Table 2-1,

$$J = K_j A_p^{-0.12} \qquad (5.A\text{-}19)$$

yielding

$$A_p = \frac{2\,(\text{Energy}) \times 10^4}{K_u\, B_m \left(K_j\, A_p^{-0.12}\right)} \qquad (5.A\text{-}20)$$

$$A_p^{0.88} = \frac{2\,(\text{Energy}) \times 10^4}{K_u\, B_m\, K_j} \qquad (5.A\text{-}21)$$

$$A_p = \left(\frac{2\,(\text{Energy}) \times 10^4}{K_u\, B_m\, K_j}\right)^{1.14} \qquad [\text{cm}^4] \quad (5.A\text{-}22)$$

After the core size has been determined, the next step is to pick the right permeability for that core size. This is done by solving for μ_Δ in equation 5.A-13.

$$\mu_\Delta = \frac{B_m\, l_m \times 10^{-2}}{\mu_o\, W_a\, J\, K_u} \qquad (5.A\text{-}23)$$

for $\mu_o = 4\pi \times 10^{-7}$

$$\mu_\Delta = \frac{B_m\, l_m \times 10^4}{0.4\pi\, W_a\, J\, K_u} \qquad (5.A\text{-}24)$$

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APPENDIX 5.B

MAGNETIC AND DIMENSIONAL SPECIFICATIONS FOR 13 COMMONLY USED MOLY-PERMALLOY CORES

The following remarks apply to each of Tables 5.B-1 to 5.B-13, the data in which was compiled from manufacturers' data.

(1) Total weight is core weight plus wire weight assuming AWG 20

(2) Maximum OD of wound core with residual hole = 1/2 ID

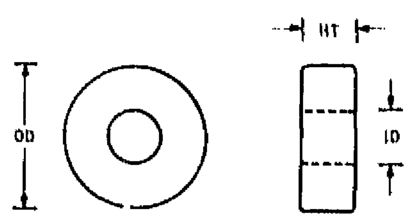
(3) MLT (mean length/turn) full wound toroid

(4) Effective window area $W_{a(eff)} = 3\pi r^2/4$

Graphs (Figs. 5.B-1 to 5.B-13) relate to the 13 different core sizes. The graphs show resistance, number of turns, inductance and wire size for a window utilization factor of 0.40, and are based on a permeability of 60. To convert for other permeability values, the appropriate inductance multiplication factors listed should be used. The information appearing in the tables and on the figures will enable the engineer to arrive at a close approximation for breadboarding purposes.

Table 5. B-1. Dimensional specifications for Magnetic Inc 55051-A2, Arnold Engineering A-051027-2

	ENGLISH	METRIC
Wa Ac		3.39
Wa x Ac	0.00104 in^4	0.0432 cm^4
OD	0.510 in	1.346 cm
ID	0.275 in	0.699 cm
HT	0.217 in	0.551 cm
Wa WINDOW AREA	0.075 x 10^6 CIR-MIL	0.381 cm^2
Wa EFFECTIVE	0.0445 in^2	0.288 cm^2
Ac CROSS SECTION	0.0175 in^2	0.113 cm^2
lm PATH LENGTH	1.229 in	3.12 cm
CORE WEIGHT	0.0066 lb	3.0 grams
TOTAL WEIGHT	0.0106 lb	5.175 grams
WOUND OD MIN	0.581 in	1.475 cm
MLT	0.850 in	2.160 cm
A_t SURFACE AREA	1.018 in^2	6.568 cm^2
PERMEABILITY		60
μ 125		2,08 x L @ μ 60
μ 160		2,67 x L @ μ 60
μ 200		3,33 x L @ μ 60
μ 550		9,17 x L @ μ 60



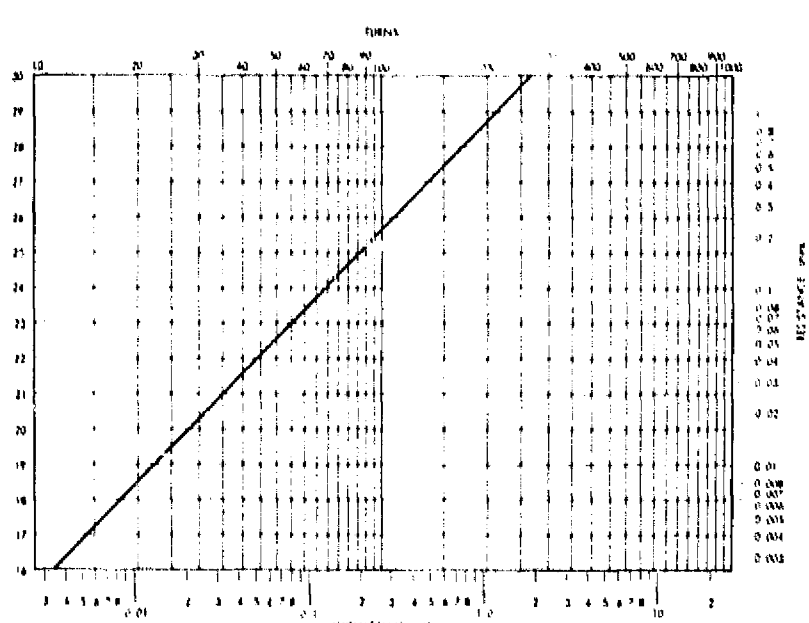


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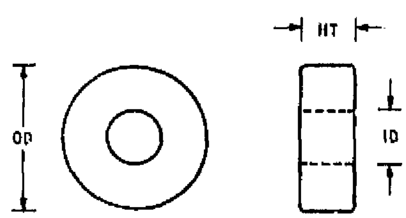
Fig. 5. B-1. Wire and inductance graph for Core 55051-A2

Table 5. B-2. Dimensional specifications for Magnetic Inc 55121-A2, Arnold Engineering A-266036-2

	ENGLISH	METRIC
Wa/Ac		3, 63
Wa x Ac	0, 0[illegible]336 in^4	0, 139 cm^4
OD	0, 680 in	1. 740 cm
ID	0. [illegible]75 in	0. 953 cm
HT	0, 280 in	0, 711 cm
Wa = WINDOW AREA	0, 141 x 10^6 CIR-MIL	0, 713 cm^2
Wa = EFFECTIVE	0, 0828 in^2	0, 535 cm^2
Ac = CROSS SECTION	0, 0304 in^2	0, 196 cm^2
lm = PATH LENGTH	1, 62 in	4, 11 cm
CORE WEIGHT	0, 014[illegible] lb	6, 50 grams
TOTAL WEIGHT	0, 0257 lb	11, 70 grams
WOUND OD MIN	0, 751 in	1, 925 cm
MLT	1, 075 in	2, 74 cm
A_t = SURFACE AREA	1, 742 in^2	11, 24 cm^2
PERMEABILITY		60
μ 125		2,08 x L @ μ 60
μ 160		2,67 x L @ μ 60
μ 200		3,33 x L @ μ 60
μ 550		9,17 x L @ μ 60



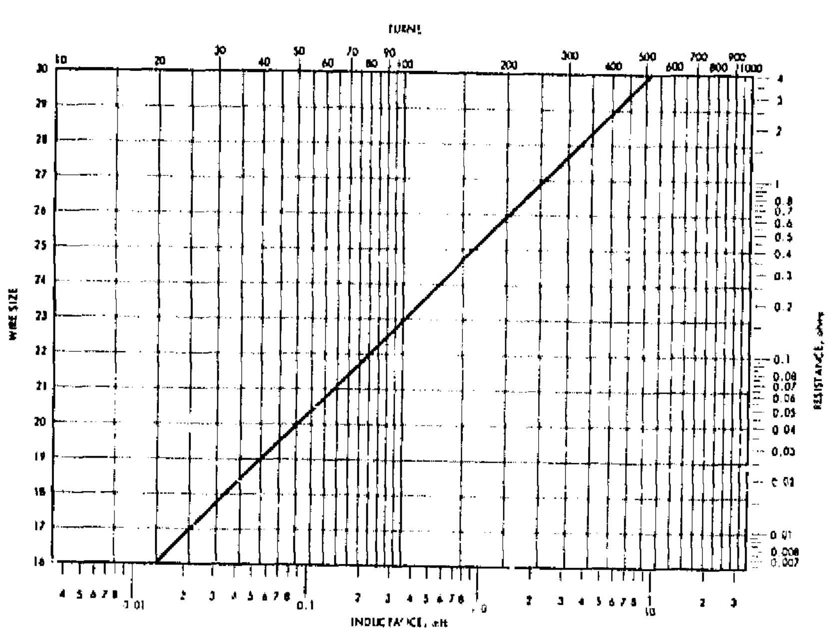
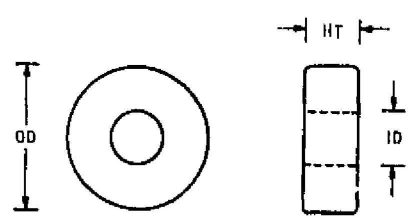


Fig. 5. B-2. Wire and inductance graph for Core 55121-A2

Table 5.B-3. Dimensional specifications for Magnetic Inc 55848-A2, Arnold Engineering A-848032-2

	ENGLISH	METRIC
Wa/Ac		4.91
Wa x Ac	0.00636 in^4	0.264 cm^4
OD	0.830 in	2.11 cm
ID	0.475 in	1.21 cm
HT	0.280 in	0.711 cm
Wa = WINDOW AREA	0.23 x 10^6 CIR-MIL	1.14 cm^2
Wa = EFFECTIVE	0.13290 in^2	0.858 cm^2
Ac = CROSS SECTION	0.036 in^2	0.232 cm^2
lm = PATH LENGTH	2.01 in	5.09 cm
CORE WEIGHT	0.021 lb	9.6 grams
TOTAL WEIGHT	0.041 lb	18.6 grams
WOUND OD MIN	0.926 in	2.35 cm
MLT	1.166 in	2.97 cm
A_t = SURFACE AREA	2.431 in^2	15.69 cm^2
PERMEABILITY		60
μ 125		2.08 x L = μ 60
μ 160		2.67 x L = μ 60
μ 200		3.33 x L = μ 60
μ 550		9.17 x L = μ 60



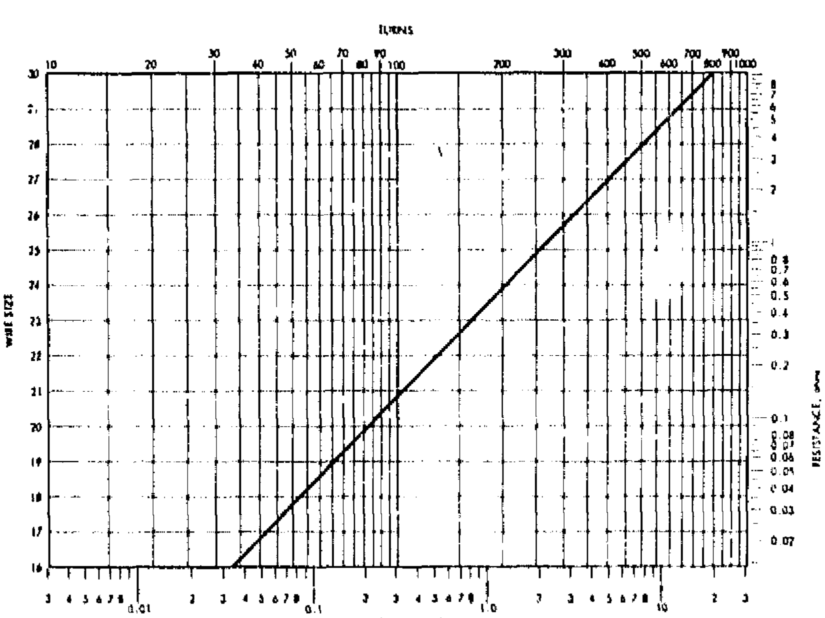
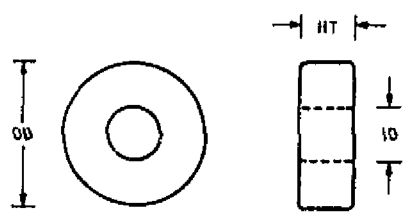


Fig. 5.B-3. Wire and inductance graph for Core 55848-A2

Table 5. B-4. Dimensional specifications for Magnetic Inc 55594-A2, Arnold Engineering A-059043-2

	ENGLISH	METRIC
Wa/Ac		4.30
Wa x Ac	0.0713 in^4	0.460 cm^4
OD	0.930 in	2.36 cm
ID	0.527 in	1.339 cm
HT	0.330 in	0.838 cm
Wa = WINDOW AREA	0.28 x 10^6 CIR-MIL	1.407 cm^2
Wa = EFFECTIVE	0.164 in^2	1.056 cm^2
Ac = CROSS SECTION	0.0507 in^2	0.327 cm^2
lm = PATH LENGTH	2.23 in	5.67 cm
CORE WEIGHT	0.033 lb	15.0 grams
TOTAL WEIGHT	0.0716 lb	32.5 grams
WOUND OD MIN	1.035 in	2.63 cm
MLT	1.356 in	3.45 cm
A_t = SURFACE AREA	3.103 in^2	20.019 cm^2
PERMEABILITY		60
μ 125		2.08 x L @ μ 60
μ 160		2.67 x L @ μ 60
μ 200		3.33 x L @ μ 60
μ 550		9.17 x L @ μ 60



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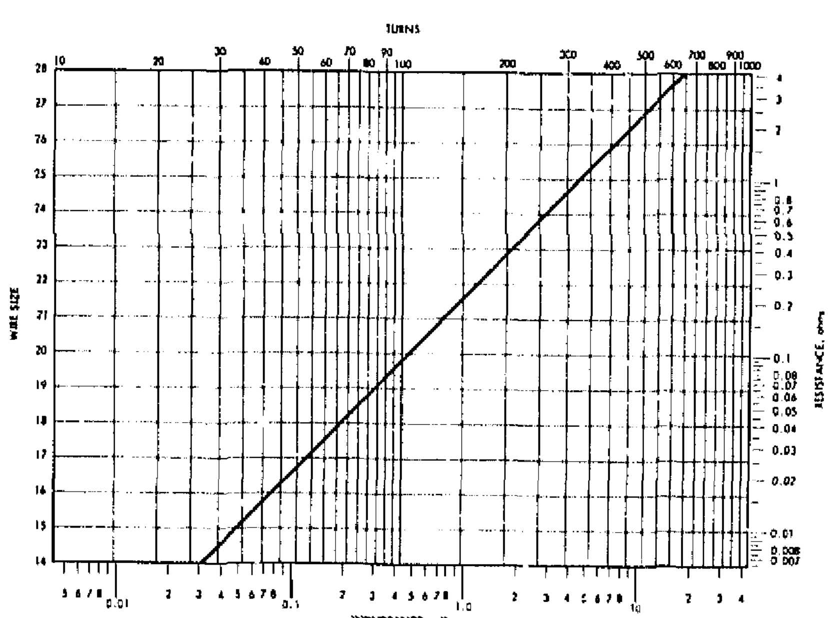
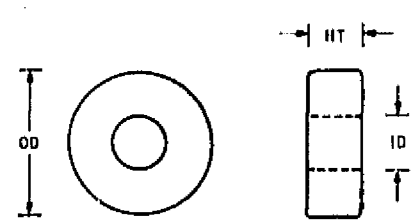


Fig. 5. B-4. Wire and inductance graph for Core 55059-A2

Table 5. B-5. Dimensional specifications for Magnetic Inc 55059-A2, Arnold Engineering A-894075-2

	ENGLISH	METRIC
Wa/Ac		2, 44
Wa x Ac	0, 0239 in^4	0, 997 cm^4
OD	1, 090 in	2, 77 cm
ID	0, 555 in	1, 41 cm
HT	0, 472 in	1, 20 cm
Wa = WINDOW AREA	0, 31 x 10^6 CIR-MIL	1, 561 cm^2
Wa - EFFECTIVE	0, 1814 in^2	1, 17 cm^2
Ac = CROSS SECTION	0, 099 in^2	0, 639 cm^2
lm = PATH LENGTH	2, 50 in	6, 35 cm
CORE WEIGHT	0, 077 lb	35 grams
TOTAL WEIGHT	0, 132 lb	59, 7 grams
WOUND OD MIN	1, 191 in	3, 03 cm
MLT	1, 81 in	4, 61 cm
A_t = SURFACE AREA	4, 38 in^2	28 32 cm^2
PERMEABILITY		60
μ 125		2,08 x L = μ 60
μ 160		2,67 x L = μ 60
μ 200		3,33 x L = μ 60
μ 550		9,17 x L = μ 60



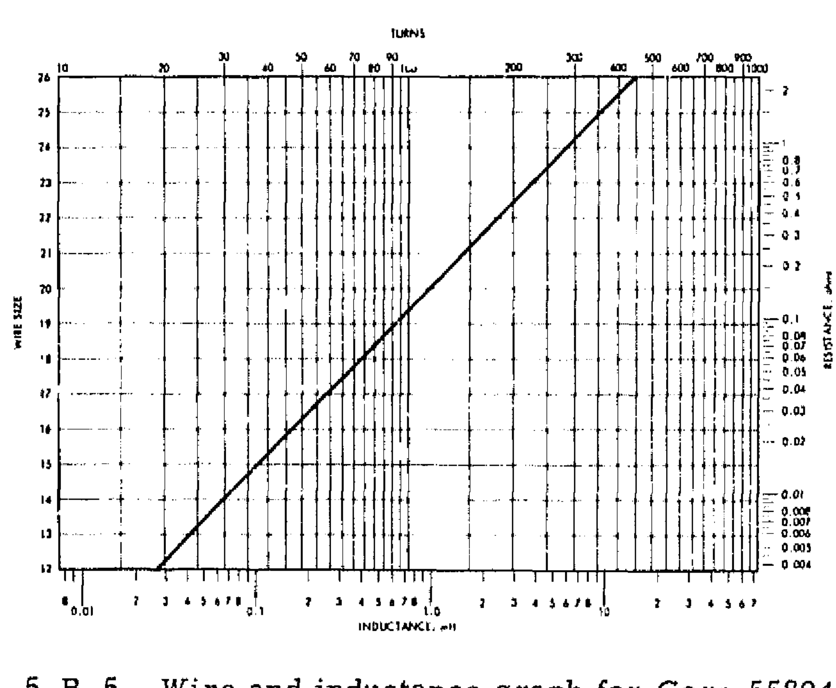
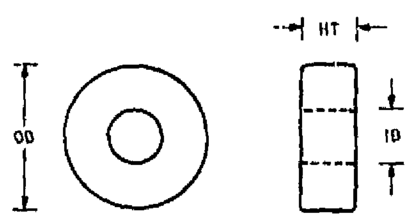


Fig. 5. B-5. Wire and inductance graph for Core 55894-A2

Table 5.B-6. Dimensional specifications for Magnetic Inc 55071-A2, Arnold Engineering A-291061-2

	ENGLISH	METRIC
Wa/Ac		4.39
Wa x Ac	0,0468 in^4	1.95 cm^4
OD	1,332 in	3.38 cm
ID	0,760 in	1.93 cm
HT	0,457 in	1.16 cm
Wa = WINDOW AREA	0,58 x 10^6 CIR-MIL	2.93 cm^2
Wa = EFFECTIVE	0,340 in^2	2,1941 cm^2
Ac = CROSS SECTION	0,1032 in^2	0,666 cm^2
lm - PATH LENGTH	3,21 in	8,15 cm
CORE WEIGHT	0,101 lb	46 grams
TOTAL WEIGHT	0,198 lb	90 grams
WOUND OD MIN	1,486 in	3,77 cm
MLT	1,89 in	4,80 cm
A_t = SURFACE AREA	4,389 in^2	40,68 cm^2
PERMEABILITY		60
μ 125		2,08 x L @ μ 60
μ 160		2,67 x L @ μ 60
μ 200		3,33 x L @ μ 60
μ 550		9,17 x L @ μ 60



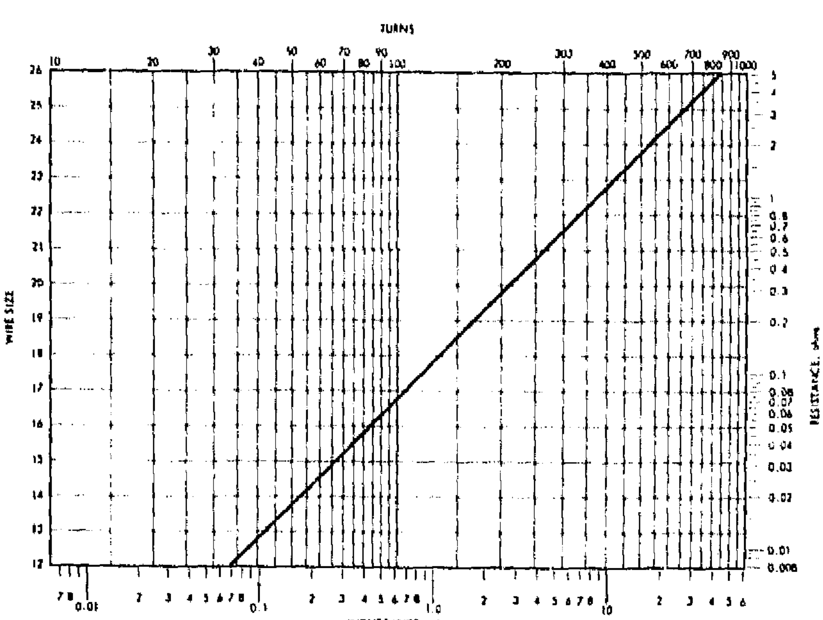
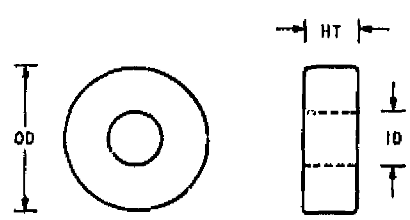


Fig. 5.B-6. Wire and inductance graph for Core 55071-A2

Table 5, B-7. Dimensional specifications for Magnetic Inc 55586-A2, Arnold Engineering A-345038-2

	ENGLISH		METRIC	
Wa/Ac			8. 73	
Wa x Ac	0, 044	in^4	1. 832	cm^4
OD	1, 382	in	3, 51	cm
ID	0, 888	in	2, 26	cm
HT	0, 387	in	0, 983	cm
Wa = WINDOW AREA	0, 79 x 10^6 CIR-MIL		4, 00	cm^2
Wa = EFFECTIVE	0, 4644	in^2	3, 009	cm^2
Ac = CROSS SECTION	0, 0710	in^2	0, 458	cm^2
lm = PATH LENGTH	3, 53	in	8, 95	cm
CORE WEIGHT	0, 075	lb	34	grams
TOTAL WEIGHT	0, 193	lb	87, 4	grams
WOUND OD MIN	1, 58	in	4, 02	cm
MLT	1, 70	in	4, 32	cm
A_t = SURFACE AREA	6. 85	in^2	44. 24	cm^2
PERMEABILITY			60	
μ 125			2,08 x L @ μ 60	
μ 160			2,67 x L @ μ 60	
μ 200			3,33 x L @ μ 60	
μ 550			9,17 x L @ μ 60	



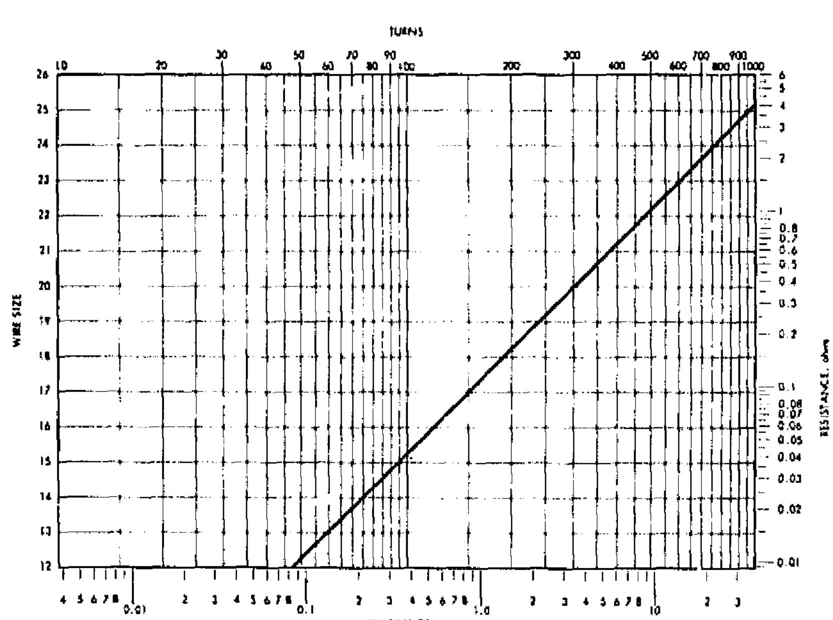
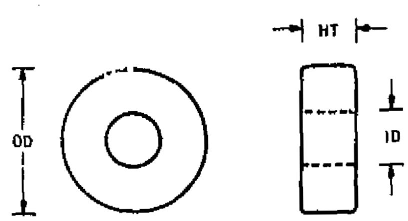


Fig. 5, B-7. Wire and inductance graph for Core 55586-A2

Table 5.B-8. Dimensional specifications for Magnetic 55076-A2, Arnold Engineering A-076056-2

	ENGLISH	METRIC
Wa/Ac		5.43
Wa x Ac	0.0580 in^4	2.44 cm^4
OD	1.44 in	3.66 cm
ID	0.848 in	2.15 cm
HT	0.444 in	1.128 cm
Wa = WINDOW AREA	0.72×10^6 CIR-MIL	3.64 cm^2
Wa = EFFECTIVE	0.424 in^2	2.723 cm^2
Ac = CROSS SECTION	0.1039 in^2	0.670 cm^2
lm = PATH LENGTH	3.54 in	8.98 cm
CORE WEIGHT	0.112 lb	51 grams
TOTAL WEIGHT	0.239 lb	108.4 grams
WOUND OD MIN	1.62 in	4.11 cm
MLT	1.91 in	4.88 cm
A_t = SURFACE AREA	7.271 in^2	46.91 cm^2
PERMEABILITY		60
μ 125		2.08 x L @ μ 60
μ 160		2.67 x L @ μ 60
μ 200		3.33 x L @ μ 60
μ 550		9.17 x L @ μ 60



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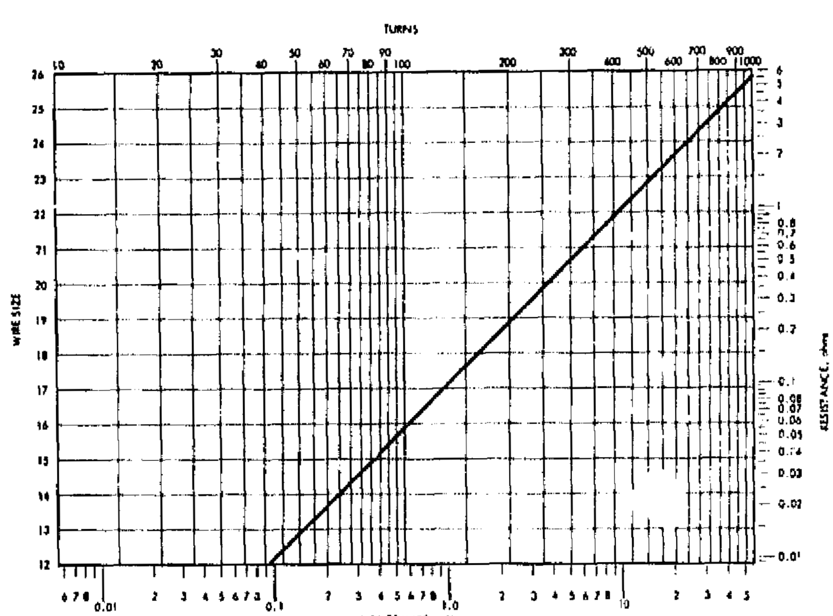
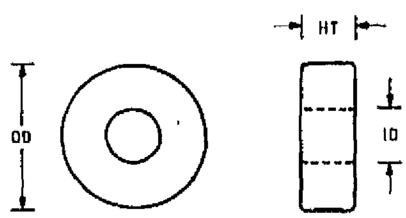


Fig. 5.B-8. Wire and inductance graph for Core 55076-A2

Table 5.B-9. Dimensional specifications for Magnetic Inc 55083-A2, Arnold Engineering A-083081-2

	ENGLISH		METRIC	
Wa/Ac			4,02	
Wa x Ac	0,108	in^4	4,53	cm^4
OD	1,602	in	4,07	cm
ID	0,918	in	2,33	cm
HT	0,605	in	1,54	cm
Wa = WINDOW AREA	$0,84 \times 10^6$	CIR-MIL	4,27	cm^2
Wa = EFFECTIVE	0,496	in^2	3,198	cm^2
Ac = CROSS SECTION	0,164	in^2	1,06	cm^2
lm = PATH LENGTH	3,88	in	9,84	cm
CORE WEIGHT	0,198	lb	90	grams
TOTAL WEIGHT	0,388	lb	176	grams
WOUND OD MIN	1,79	in	4,54	cm
MLT	2,36	in	6,07	cm
A_t = SURFACE AREA	9,46	in^2	61,05	cm^2
PERMEABILITY			60	
μ 125			2.08 x L = μ 60	
μ 160			2.67 x L = μ 60	
μ 200			3.33 x L = μ 60	
μ 550			9.17 x L = μ 60	



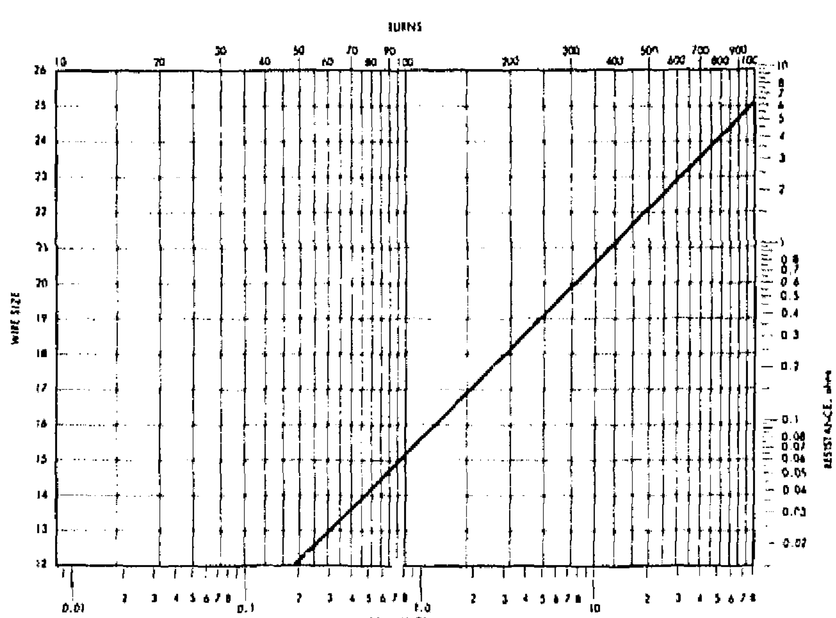
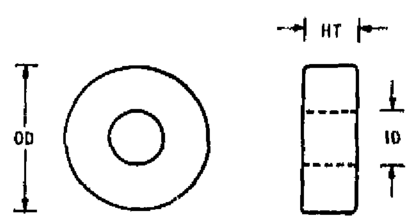


Fig. 5.B-9. Wire and inductance graph for Core 55083-A2

Table 5. B-10. Dimensional specifications for Magnetic Inc 55439-A2, Arnold Engineering A-759135-2

	ENGLISH		METRIC	
Wa/Ac			2.19	
Wa x Ac	0.200	in^4	8.33	cm^4
OD	1.875	in	4.76	cm
ID	0.918	in	2.33	cm
HT	0.745	in	1.89	cm
Wa WINDOW AREA	0.84 x 10^6	CIR-MIL	4.27	cm^2
Wa = EFFECTIVE	0.496	in^2	3.198	cm^2
Ac = CROSS SECTION	0.302	in^2	1.95	cm^2
lm = PATH LENGTH	4.23	in	10.74	cm
CORE WEIGHT	0.396	lb	180	grams
TOTAL WEIGHT	0.641	lb	291	grams
WOUND OD MIN	2.04	in	5.17	cm
MLT	3.00	in	7.62	cm
A_t SURFACE AREA	12.30	in^2	79.37	cm^2
PERMEABILITY			60	
μ 125			2.08 x L @ μ 60	
μ 160			2.67 x L @ μ 60	
μ 200			3.33 x L @ μ 60	
μ 550			9.17 x L @ μ 60	



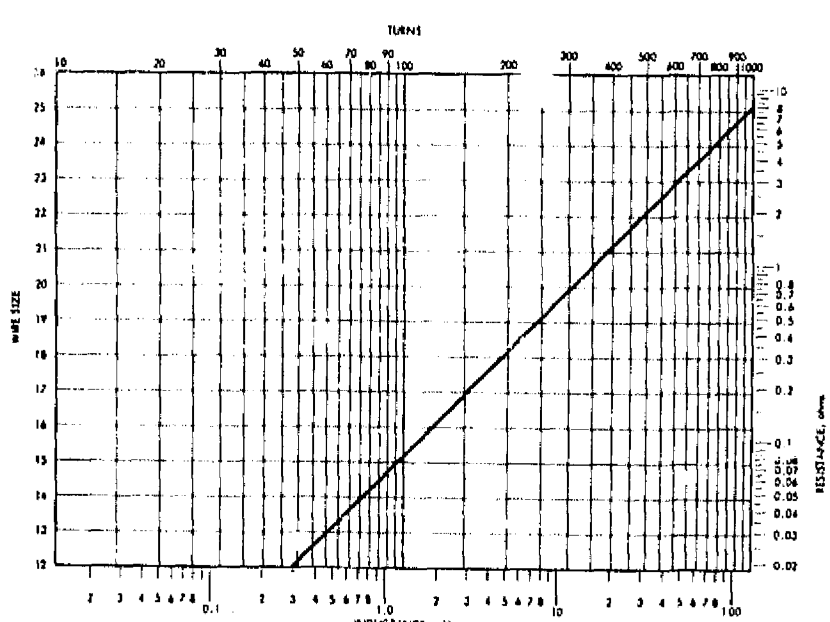
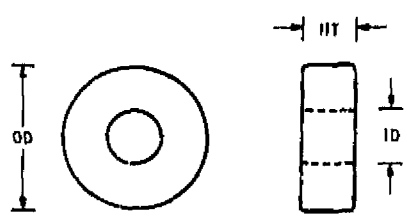


Fig. 5. B-10. Wire and inductance graph for Core 55439-A2

Table 5.B-11. Dimensional specifications for Magnetic Inc 55110-A2, Arnold Engineering A-488075-2

	ENGLISH	METRIC
Wa/Ac		6, 58
Wa x Ac	0, 328 in^4	11, 65 cm^4
OD	2, 285 in	5, 8 cm
ID	1, 368 In	3, 47 cm
HT	0, 585 in	1, 486 cm
Wa = WINDOW AREA	1, 87 x 10^6 CIR-MIL	9, 48 cm^2
Wa = EFFECTIVE	1, 1023 in^2	7, 093 cm^2
A_c = CROSS SECTION	0, 223 in^2	1, 44 cm^2
lm = PATH LENGTH	5, 63 in	14, 30 cm
CORE WEIGHT	0, 385 lb	175 grams
TOTAL WEIGHT	0, 864 lb	392 grams
WOUND OD MIN	2, 57 in	6, 53 cm
MLT	2, 75 in	7, 00 cm
A_t = SURFACE AREA	17, 42 in^2	112, 4 cm^2
PERMEABILITY		60
μ 125		2,08 x L = μ 60
μ 160		2,67 x L = μ 60
μ 200		3,33 x L = μ 60
μ 550		9,17 x L = μ 60



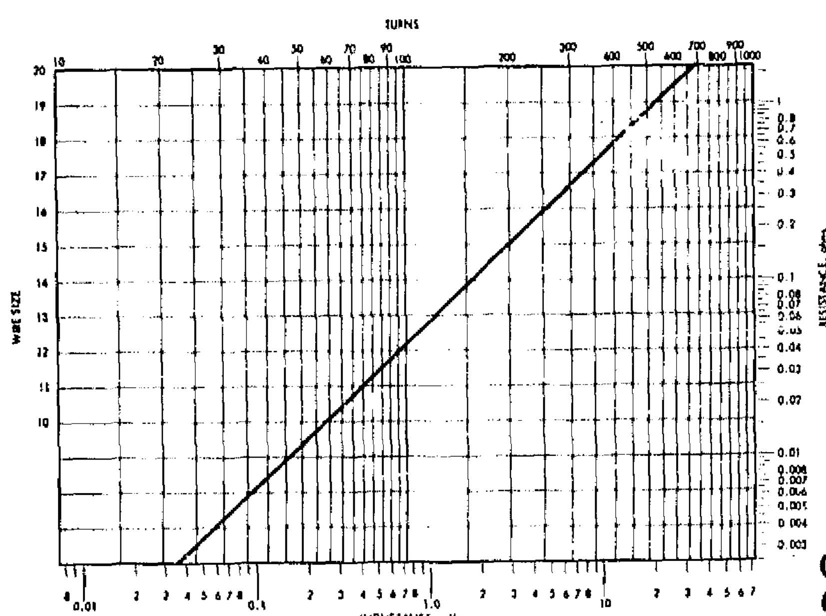


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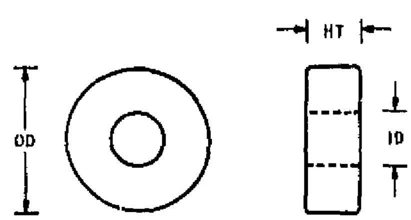
Fig. 5.B-11. Wire and inductance graph for Core 55110-A2

Table 5.B-12. Dimensional specifications for Magnetic Inc 55716-A2, Arnold Engineering A-106073-2

	ENGLISH		METRIC	
Wa/Ac			6,06	
Wa x Ac	0,224	in^4	9,32	cm^4
OD	2,035	in	5,17	cm
ID	1,218	in	3,09	cm
HT	0,565	in	1,435	cm
Wa WINDOW AREA	1.48×10^6 CIR-MIL		7,52	cm^2
Wa - EFFECTIVE	0,874	in^2	5,62	cm^2
Ac - CROSS SECTION	0,192	in^2	1,24	cm^2
lm PATH LENGTH	5,02	in	12,73	cm
CORE WEIGHT	0,298	lb	135	grams
TOTAL WEIGHT	0,652	lb	296	grams
WOUND OD MIN	2,29	in	5,82	cm
MLT	2,55	in	6,50	cm
A_t - SURFACE AREA	14,15	in^2	91,32	cm^2
PERMEABILITY			60	
μ 125			2,08 x L @ μ 60	
μ 160			2,67 x L @ μ 60	
μ 200			3,33 x L @ μ 60	
μ 550			9,17 x L @ μ 60	



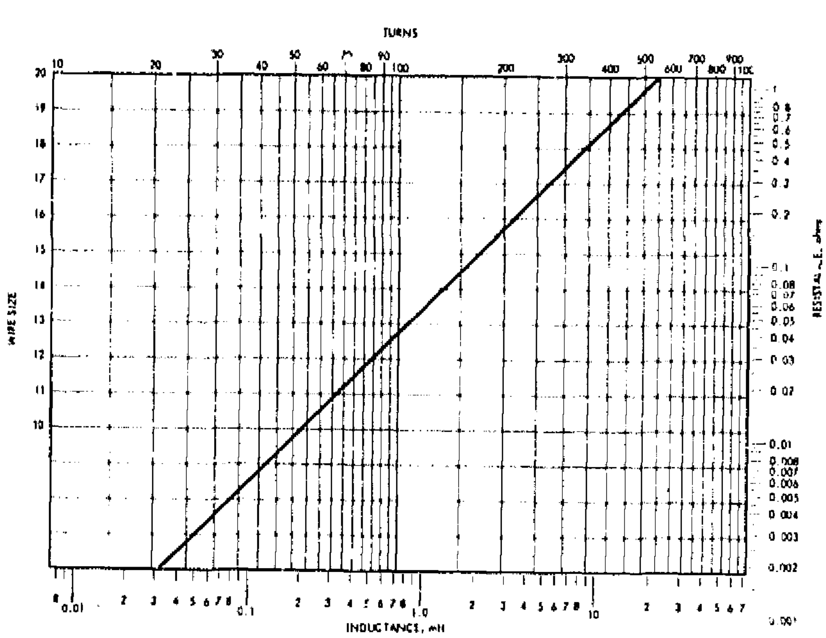
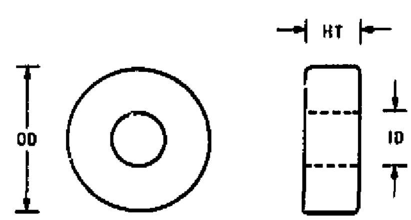


Fig. 5.B-12. Wire and inductance graph for Core 55716-A2

Table 5.B-13. Dimensional specifications for Magnetic Inc 55090-A2, Arnold Engineering A-090086-2

	ENGLISH	METRIC
Wa/Ac		4.63
Wa x Ac	0.194 in^4	8.06 cm^4
OD	1.875 in	4.76 cm
ID	1.098 in	2.79 cm
HT	0.635 in	1.61 cm
Wa = WINDOW AREA	1.21×10^6 CIR-MIL	6.11 cm^2
Wa = EFFECTIVE	0.710 in^2	4.58 cm^2
Ac = CROSS SECTION	0.205 in^2	1.32 cm^2
lm = PATH LENGTH	4.58 in	11.62 cm
CORE WEIGHT	0.286 lb	130 grams
TOTAL WEIGHT	0.588 lb	267 grams
WOUND OD MIN	2.10 in	5.34 cm
MLT	2.62 in	6.66 cm
A_t = SURFACE AREA	12.64 in^2	81.58 cm^2
PERMEABILITY		60
μ 125		2.08 x L @ μ 60
μ 160		2.67 x L @ μ 60
μ 200		3.33 x L @ μ 60
μ 550		9.17 x L @ μ 60



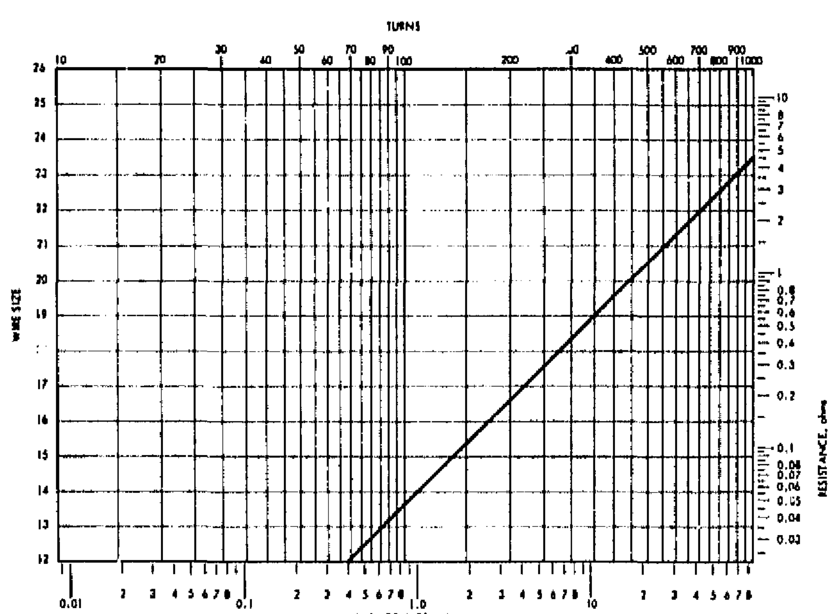


Fig. 5.B-13. Wire and inductance graph for Core 55090-A2

CHAPTER VI

WINDOW UTILIZATION FACTOR K_u

A. INTRODUCTION

The window utilization factor is the amount of copper that appears in the window area of the transformer or inductor. The window utilization factor is influenced by 4 different factors: (1) wire insulation, (2) wire lay (fill factor), (3) bobbin area (or, when using a toroid, the clearance hole for passage of the shuttle), and (4) insulation required for multilayer windings or between windings. In the design of high-current or low-current transformers, the ratio of conductor area over total wire area can vary from 0.941 to 0.673, depending on the wire size. The wire lay or fill factor can vary from 0.7 to 0.55, depending on the winding technique. The amount and the type of insulation are dependent on the voltage.

B. WINDOW UTILIZATION FACTOR

The fraction K_u of the available core window space which will be occupied by the winding (copper) is calculated from areas S_1, S_2, S_3, and S_4:

$$K_u = S_1 \times S_2 \times S_3 \times S_4 \qquad (6\text{-}1)$$

where

$$S_1 = \frac{\text{conductor area}}{\text{wire area}}$$

$$S_2 = \frac{\text{wound area}}{\text{usable window area}}$$

$$S_3 = \frac{\text{usable window area}}{\text{window area}}$$

$$S_4 = \frac{\text{usable window area}}{\text{usable window area + insulation area}}$$

in which

conductor area = copper area

wire area = copper area + insulation area

wound area = number of turns x wire area of one turn

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usable window area = available window area minus residual area which results from the particular winding technique used

window area = available window area

insulation area = area usable for winding insulation

S_1 is dependent upon wire size. Columns A and D of Table 6-1 may be used for calculating some typical values such as for AWG 10, AWG 20, AWG 30 and AWG 40.

Thus:

$$\text{AWG 10} = \frac{52.61\ \text{cm}^2}{55.90\ \text{cm}^2} = 0.941;$$

$$\text{AWG 20} = \frac{5.188\ \text{cm}^2}{6.065\ \text{cm}^2} = 0.855;$$

$$\text{AWG 30} = \frac{0.5067\ \text{cm}^2}{0.6785\ \text{cm}^2} = 0.747;\ \text{and}$$

$$\text{AWG 40} = \frac{0.048\text{[illegible]}3\ \text{cm}^2}{0.0723\ \text{cm}^2} = 0.673 \quad .$$

When designing low-current transformers, it is advisable to reevaluate S_1 because of the increased amount of insulation.

S_2 is the fill factor for the usable window area. It can be shown that for circular cross-section wire wound on a flat form the ratio of wire area to the area required for the turns can never be greater than 0.91. In practice, the actual maximum value is dependent upon the tightness of winding, variations in insulation thickness, and wire lay. Consequently, the fill factor is always less than the theoretical maximum.

As a typical working value for copper wire with a heavy synthetic film insulation, a ratio of 0.60 may be safely used.

The term S_3 defines how much of the available window space may actually be used for the winding. The winding area available to the designer depends on the bobbin configuration. A single bobbin design offers an effective area W_a between 0.835 to 0.929 while a two bobbin configuration offers an effective area W_a between 0.687 to 0.872. A good value to use for both configurations is 0.75.

When designing with a pot core, S_3 has to be reduced because the effective W_a varies between 0.55 and 0.71.

The term S_4 defines how much of the usable window space is actually being used for insulation. If the transformer has multiple secondaries having significant amounts of insulation S_4 should be reduced by 10% for each additional secondary winding because of the added space occupied by insulation and partly due to poorer space factor.

A typical value for the copper fraction in the window area is about 0.40. For example, for AWG 20 wire, $S_1 \times S_2 \times S_3 \times S_4 = 0.855 \times 0.60 \times 0.75 \times 1.0 = 0.385$, which is very close to 0.4.

This may be stated somewhat differently as:

$$0.4 = \underbrace{\frac{A_w \text{ Bare}}{A_w \text{ Total}}}_{(S_1)} \times \underbrace{\text{Fill Factor}}_{(S_2)} \times \underbrace{\frac{W_{a(eff)}}{W_a}}_{(S_3)} \times \underbrace{\text{Insulation Factor}}_{(S_4)}$$

C. CONVERSION DATA FOR WIRE SIZES FROM #10 to #44

Columns A and B in Table 6-1 give the bare area in the commonly used circular mils notation and in the metric equivalent for each wire size. Column C gives the equivalent resistance in microhms/centimeter ($\mu\Omega$/cm or $10^{-6}\Omega$/cm.) in wire length for each wire size. Columns D to L relate to coated wires showing the effect of insulation on size and the number of turns and the total weight in grams/centimeter.

The total resistance for a given winding may be claculated by multiplying the MLT (mean length/turn) of the winding in centimeters, by the microhms cm for the appropriate wire size (Column C), and the total number of turns. Thus

$$R = (MLT) \times (N) \times (\text{Column C}) \times \zeta \times 10^{-6} \qquad \text{[ohms]}$$

For resistance correction factor ζ (Zeta) for higher and lower temperature, see Figure 6-1.

Table 6-1. Wire table

Awg Wire Size	Bare Area		Resistance	Heavy Synthetics								
	$cm^2 10^{-3}$ (footnote b)	CIR-MIL[a]	$\frac{10^{-6}\,\Omega}{cm \text{ at } 20°C}$	Area $cm^2 10^{-3}$	Area CIR-MIL[a]	Diameter cm	Diameter Inch[a]	Turns-Per cm	Turns-Per Inch[a]	Turns-Per cm^2	Turns-Per $Inch^2$	Weight gm/cm
10	52.61	10384	32.70	55.9	11046	0.267	0.1051	3.87	9.5	10.73	69.20	0.468
11	41.68	8226	41.37	44.5	8798	0.238	0.0938	4.36	10.7	13.48	89.95	0.1750
12	33.08	6529	52.09	35.64	7022	0.213	0.0838	4.85	11.9	16.81	108.4	0.2977
13	26.26	5184	65.64	28.36	5610	0.190	0.0749	5.47	13.4	21.15	136.4	0.2367
14	20.82	4109	82.80	22.95	4556	0.171	0.0675	6.04	14.8	26.14	168.6	0.1879
15	16.51	3260	104.3	18.37	3624	0.153	0.0602	6.77	16.6	32.66	210.6	0.1492
16	13.07	2581	131.8	14.73	2905	0.137	0.0539	7.32	18.6	40.73	262.7	0.1184
17	10.39	2052	165.8	11.68	2323	0.122	0.0482	8.18	20.8	51.[illegible]	331.2	0.0941
18	8.228	1624	209.5	9.326	1857	0.109	0.0431	9.13	23.2	64.33	41[illegible].9	0.07472
19	6.531	1289	263.9	7.539	1490	0.0980	0.0386	10.19	25.9	79.85	515.0	0.05940
20	5.188	1024	332.3	6.065	1197	0.0879	0.0346	11.37	28.9	98.93	638.1	0.04726
21	4.116	812.3	418.9	4.837	954.8	0.0785	0.0309	12.75	32.4	124.0	799.8	0.03757
22	3.243	640.1	531.4	3.857	761.7	0.0701	0.0276	14.25	36.2	155.5	1003	0.02965
23	2.588	510.8	666.0	3.135	620.0	0.0632	0.0249	15.82	40.2	191.3	1234	0.02372
24	2.047	404.0	842.1	2.514	497.3	0.0566	0.0223	17.63	44.8	238.6	1539	0.01884
25	1.623	320.4	1062.9	2.002	396.0	0.0505	0.0199	19.80	50.3	299.7	1933	0.01498
26	1.280	252.8	1345.0	1.603	316.8	0.0452	0.0178	22.12	56.2	374.2	2414	0.01185
27	1.021	201.6	1687.6	1.313	259.2	0.0409	0.0161	24.44	62.1	456.9	2947	0.00945
28	0.8046	158.8	2142.7	1.0515	207.3	0.0366	0.0144	27.32	69.4	570.6	3680	0.00747
29	0.6470	127.7	2664.3	0.8548	169.0	0.0330	0.0130	30.27	76.9	701.9	4527	0.00602
30	0.5067	100.0	3402.2	0.6785	134.5	0.0294	0.0116	33.93	86.2	884.3	5703	0.00472
31	0.4013	79.21	4294.6	0.5596	110.2	0.0267	0.0105	37.48	95.2	1072	6914	0.00372
32	0.3242	64.00	5314.9	0.4559	90.25	0.0241	0.0095	41.45	105.3	1316	8488	0.00305
33	0.2554	50.41	6748.6	0.3662	72.25	0.0216	0.0085	46.33	117.7	1638	10565	0.00241
34	0.2011	39.69	8572.8	0.2863	56.25	0.0191	0.0075	52.48	133.3	2095	13512	0.00189
35	0.1589	31.36	10849	0.2268	44.89	0.0170	0.0067	58.77	149.3	2645	17060	0.00150
36	0.1266	25.00	13608	0.1813	36.00	0.0152	0.0060	65.62	166.7	3309	21343	0.00119
37	0.1026	20.25	16801	0.1538	30.25	0.0140	0.0055	71.57	181.8	3901	25161	0.000977
38	0.08107	16.00	21266	0.1207	24.01	0.0124	0.0049	80.35	204.1	4971	32062	0.000773
39	0.06207	12.25	27775	0.0932	18.49	0.0109	0.0043	91.57	232.6	6437	41518	0.000593
40	0.04869	9.61	35400	0.0723	14.44	0.0096	0.0038	103.6	263.2	8298	53522	0.000464
41	0.03972	7.84	43405	0.0584	11.56	0.00863	0.0034	115.7	294.1	10273	66260	0.000379
42	0.03166	6.25	54429	0.04558	9.00	0.00762	0.0030	131.2	333.3	13163	84901	0.000299
43	0.02452	4.84	70308	0.03683	7.29	0.00685	0.0027	145.8	370.4	162[illegible]1	105076	0.000233
44	0.0202	4.00	85072	0.03165	6.25	0.00635	0.0025	157.4	400.0	18957	122272	0.000195
	A	B	C	D	E	F	G	H	I	J	K	L

[a] This data from REA Magnetic Wire Datalator (Ref. 1).

[b] This notation means the entry in the column must be multiplied by 10^{-3}

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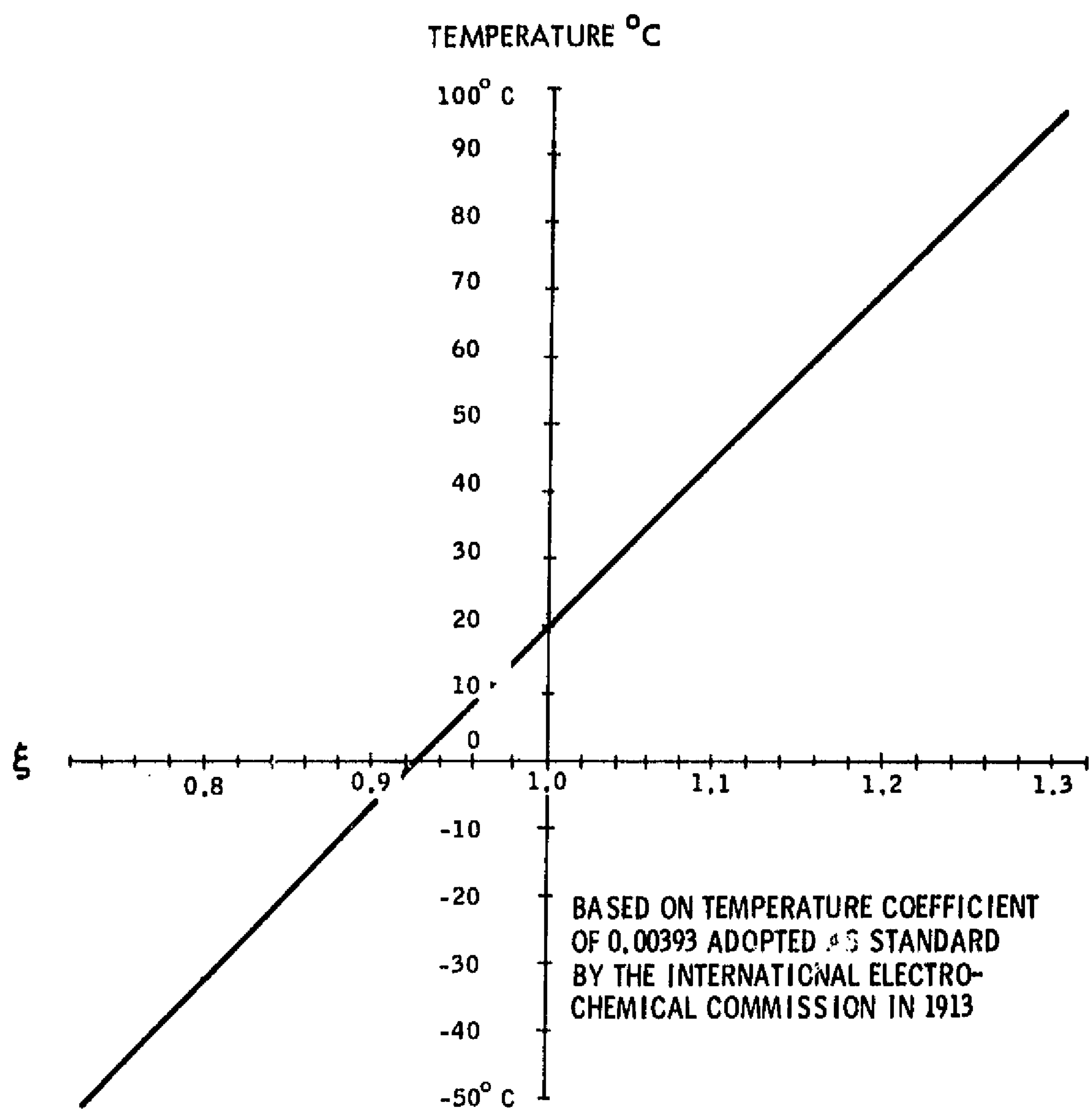

Fig. 6-1. Resistance Correction Factor ζ, (Zeta) for wire resistance at temperatures between -50° and 100°C

The weight of the copper in a given winding may be calculated by multiplying the MLT by the grams/cm (Column L) and by the total number of turns. Thus

$$W_t = (MLT) \times (N) \times (\text{Column L}) \qquad [\text{grams}]$$

Turns per square inch and turns per square cm are based on 60% wire fill factor. Mean length/turn for a given winding may be calculated with the aid of Fig. 6-2. Figure 6-3 shows a transformer being constructed using layer insulation. When a transformer is being built in this way, Table 6-2 and 6-3 will help the designer find the correct insulation thickness and margin for the appropriate wire size.

D. TEMPERATURE CORRECTION FACTORS

The resistance values given in Table 6-1 are based upon a temperature of 20°C. For other temperatures the effect upon wire resistance can be calculated by multiplying the resistance value for the wire size shown in column C of Table 6-1 by the appropriate correction factor shown on the graph. Thus, Corrected Resistance = $\mu\Omega$/cm (at 20°C) x ζ.

E. WINDOW UTILIZATION FACTOR FOR A TOROID

The toroidal magnetic component has found wide use in industry and aerospace because of its high frequency capability. The high frequency capability of the toroid is due to its high ratio of window area over core cross section and its ability to accommodate different strip thickness in its boxed configuration. Tape strip thickness is an important consideration in selecting cores. Eddy-current losses in the core can be reduced at higher frequencies by use of thinner strip stock. The high ratio of window area over core cross section insures the minimum of iron and large winding area to minimize the flux density and core loss.

The magnetic flux in the tape wound toroid can be contained inside the core more readily than in lamination or C type core as the winding covers the core along the whole magnetics path length which gives lower electromagnetic interference.

The toroid does not give a smooth A_p relationship as lamination, C core, powder cores and pot cores with respect to volume, weight, surface area and current density as can be seen in Chapter 2. This is because the actual core is always embedded in a case having a wall thickness which has no fixed relation to the actual core and becomes relatively large the smaller the actual core

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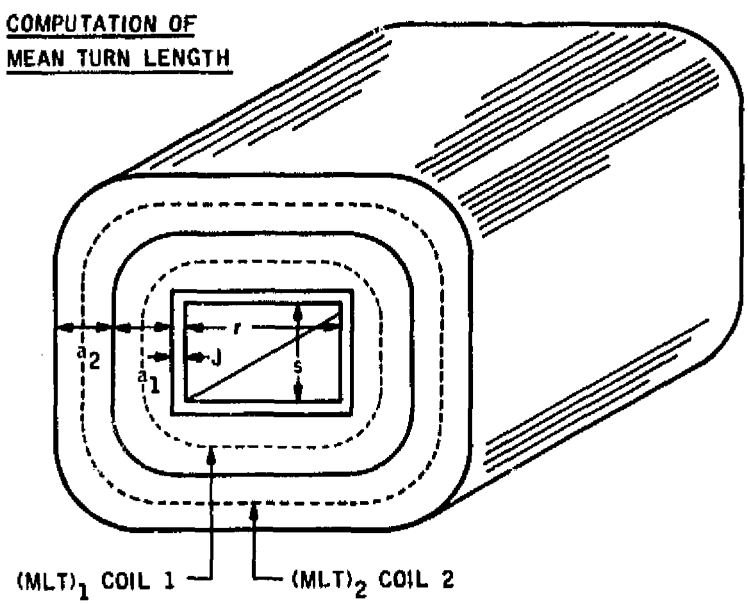

$(MLT)_1 = 2(r+2J) + 2(s+2J) + \pi a_1$

$(MLT)_2 = 2(r+2J) + 2(s+2J) + \pi(2a_1+a_2)$

OR

$(MLT)_2 = (MLT)_1 + \quad (a_1+a_2+2c)$

OR

$(MLT)_n = 2(r+2J) + 2(s+2J) + \pi \left[2(a_1+a_2+\ldots+a_{n-1}) + a_n\right]$

WHERE:

a_1 = BUILD OF WINDING #1
a_2 = BUILD OF WINDING #2
a_n = BUILD OF WINDING #n
c = THICKNESS OF INSULATION BETWEEN a_1 & a_2

Fig. 6-2. Computation of mean turn length

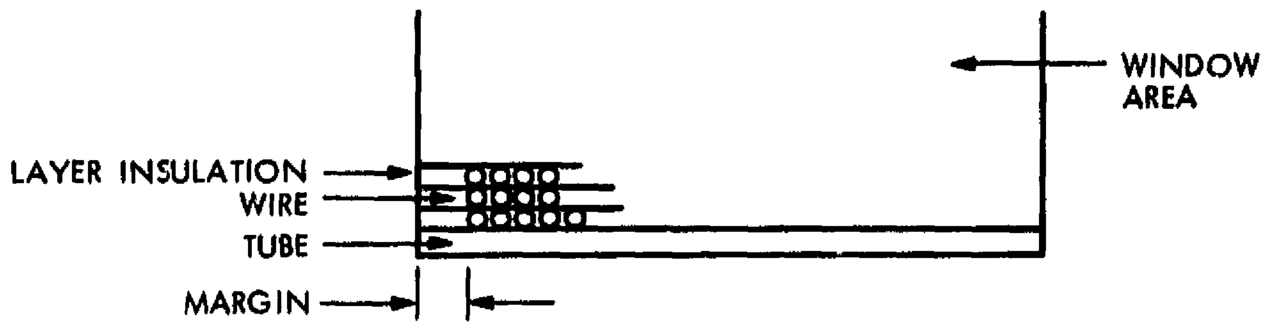

Fig. 6-3. Layer insulated coil

Table 6-2. Layer insulation vs AWG

AWG	Insulation thickness	
	cm	inch
10-16	0.0254	0.01
17-19	0.0178	0.007
20-21	0.0127	0.005
22-23	0.0076	0.003
24-27	0.0051	0.002
28-33	0.00381	0.0015
34-41	0.00254	0.001
42-46	0.00127	0.0005

Table 6-3. Margin vs AWG

AWG	Margin	
	cm	inch
10-15	0.635	0.250
16-18	0.475	0.187
19-21	0.396	0.156
22-31	0.318	0.125
32-37	0.236	0.093
38 →	0.157	0.062

cross section is. The available window area inside the case, therefore, is not a fixed percentage of the window area of the uncased core.

Design Manual TWC-300 of MAGNETICS, Inc. indicates that random wound cores can be produced with fill factors as high as 0.7, but that progressive sector wound cores can be produced with fill factors of only up to 0.55. As a typical working value for copper wire with a heavy synthetic film insulation, a ratio of 0.60 may be used safely. Figure 6-4 is based upon a fill factor ratio of 0.60 for wire sizes 14 through 42 with 0.5 I.D. remaining.

The term usable window cm^2/window cm^2 (S_3) defines how much of the available window space may actually be used for the winding. Figure 6-5 is based on the assumption that the inside diameter (ID) of the wound core is one-half that of the bare core, i.e., $S_3 = 0.75$ (to allow free passage of the shuttle).

Insulation factor (S4) in Figure 6-4 is 1.0; this does not take into account any insulation. The window utilization factor (K_u) is highly influenced by the insulation factor (S_4) because of the rapid build-up of insulation in a toroid as shown in Figure 6-6.

It can be seen in Figure 6-6 the insulation build up is greater on the inside than on the outside. For an example in Figure 6-6 if 1.27 cm wide tape was to be used with an overlap of 0.32 cm on the O.D. the overlap thickness

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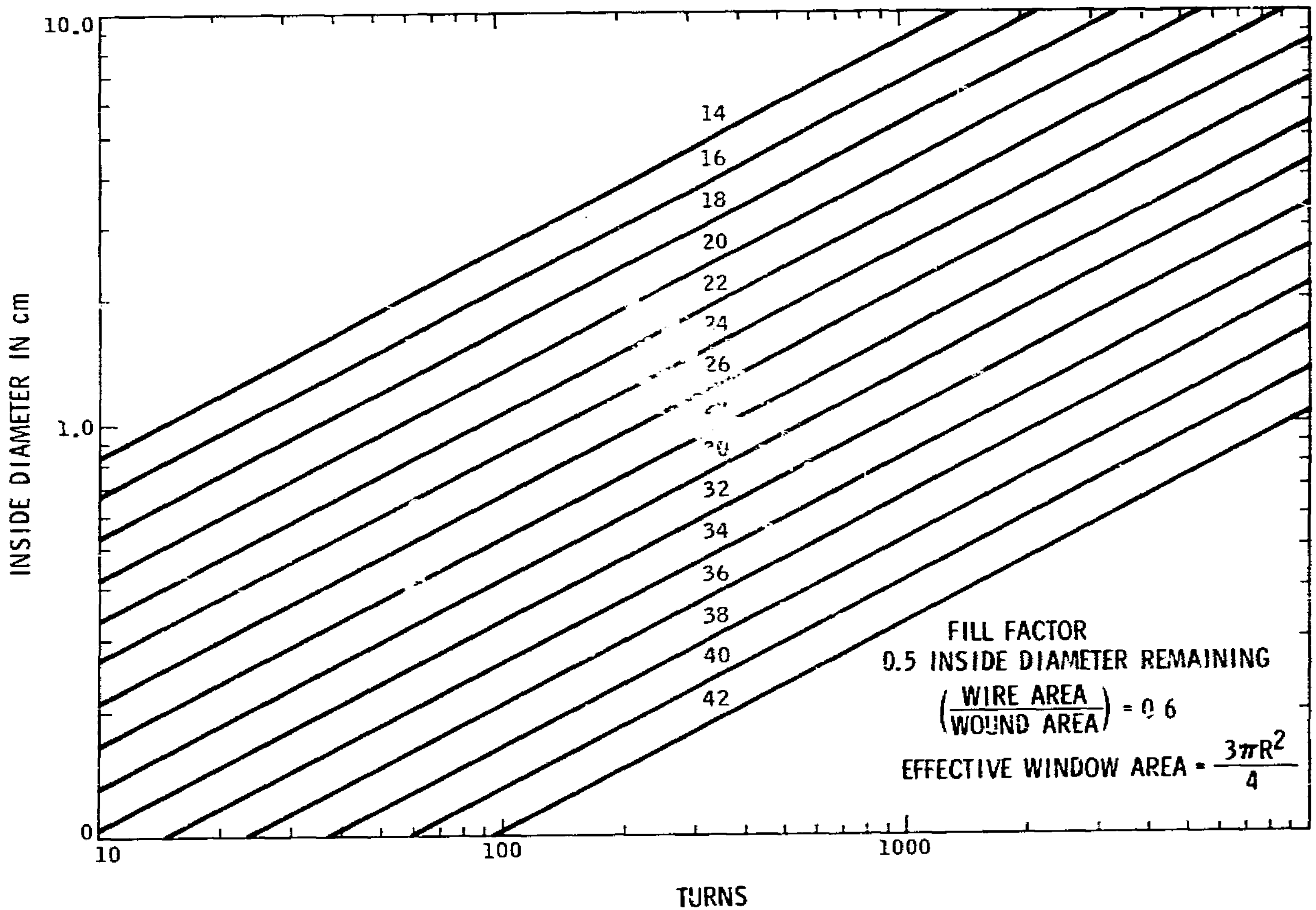

Figure 6-4. Toroid inside diameter versus turns

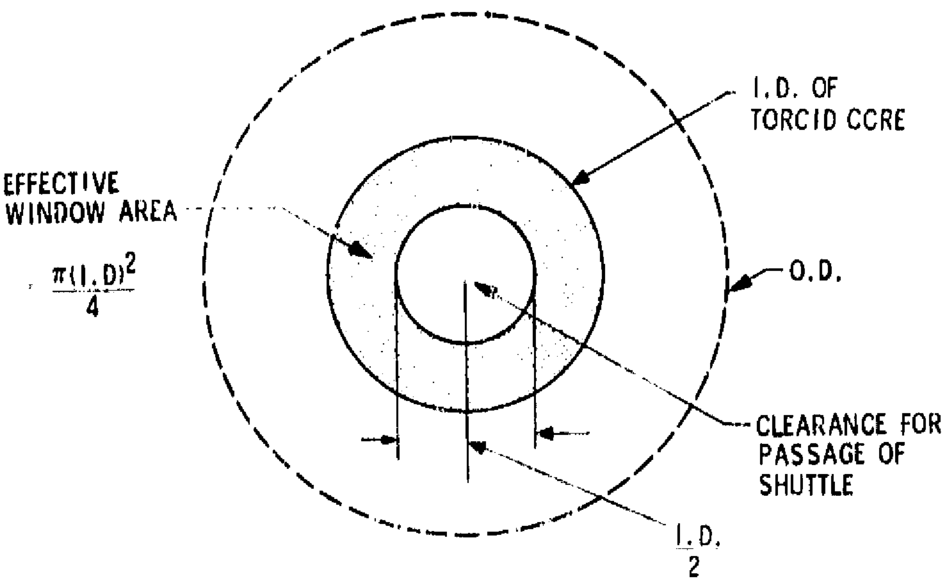

Figure 6-5. Effective winding area of a toroid

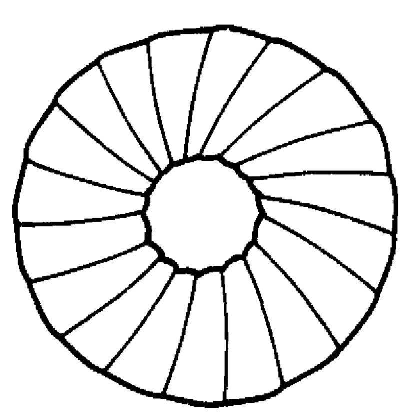

Figure 6-6. Wrap toroid

would be four times the thickness of the tape. It will be noted that the amount of overlap will depend greatly on the size of the toroid. As the toroid window gets smaller the over-lap increases. There is a way to minimize the build on a wrapped toroid and that is to use periphery insulation as shown in Figure 6-7. The use of periphery insulation minimizes the inside diameter overlay as shown in Figure 6-8.

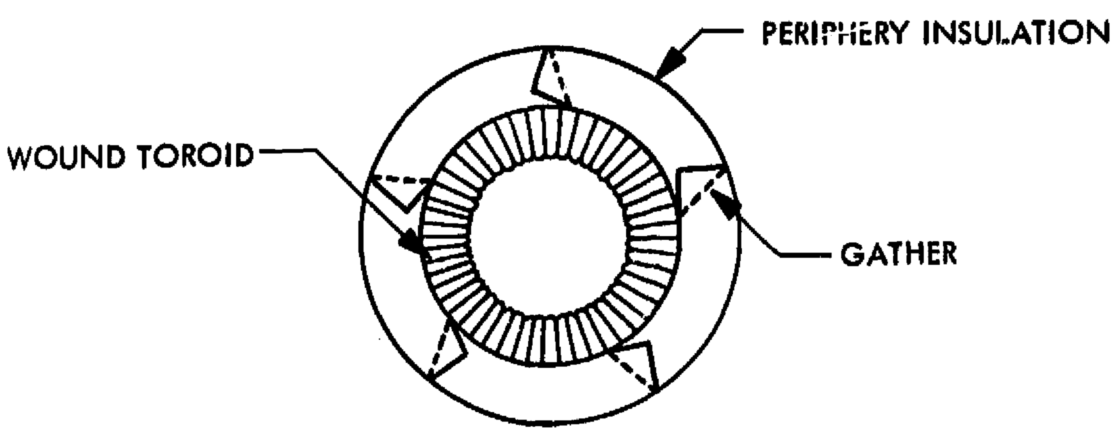

Figure 6-7. Periphery insulation

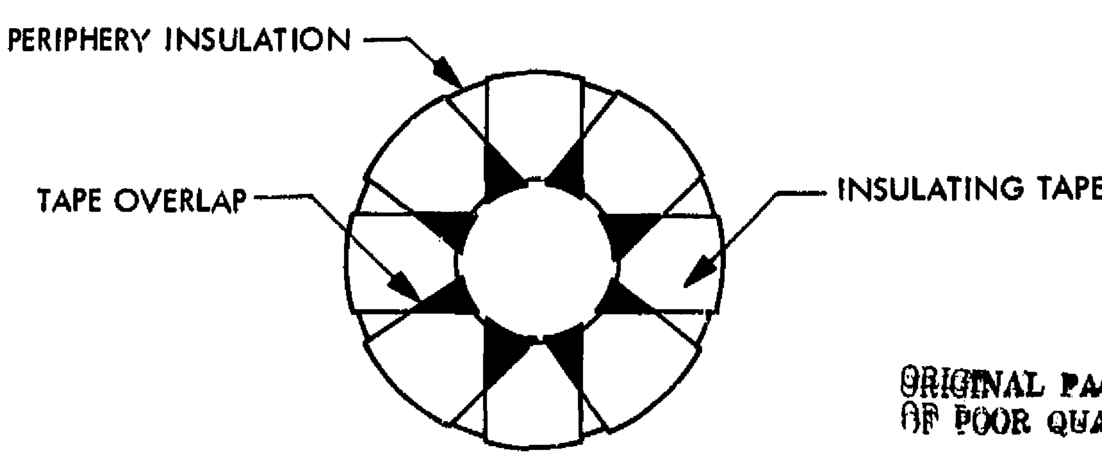

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Figure 6-8. Minimizing toroidal inside build

When a design requires a multitude of windings, all of which have to be insulated, then the insulation factor (S_4) becomes very important in the window utilization factor (K_u). For example, a low current toroidal transformer with insulation has a significant influence on the window utilization factor as shown below:

$$S_1 = \#40 \text{ AWG} \qquad K_u \quad S_1 \times S_2 \times S_3 \times S_4$$

$$K_u = 0.673 \times 0.60 \times 0.75 \times 0.80$$

$$K_u = 0.242.$$

Table 6-4 was generated as an aid for the engineer; it is a listing of 29 A.I.E.E. preferred tape-wound toroidal cores with metric dimension. The power handling capability is listed in the last column under A_p area product.

Table 6-4. A.I.E.E. preferred tape-wound toroidal cores

Mag Inc	Arnold	(1) $A_c(cm^2)$	$W_a\ (cm^2)$	(2) ID(cm)	OD(cm)	Ht(cm)	$l_m(cm)$	(3) Core Wt (grams)	$A_p(cm^4)$
52056	8T8043	0,043	0,915	1.079	1.778	0,559	4,49	1,67	0,0393
52000	8T5340	0.086	0,915	1,079	2,095	0,559	4,99	3,73	0,0787
52076	8T5958	0,193	1,478	1.372	2,756	0,711	6,48	10,9	0,285
52007	8T5651	0,257	1,478	1,372	2,756	0,876	6,19	14,5	0,380
52002	8T5515	0,086	1,674	1,450	2,476	0,559	6,98	4,62	0,144
52061	8T5502	0,171	2,274	1,702	2,743	0,876	7,48	10,4	0,389
52106	8T5504	0,193	2,274	1,702	3,051	0,711	8,98	12,6	0.437
52011	8T4168	0,086	4,242	2,324	3,391	0.559	8,98	6,71	0 365
52004	8T7699	0,171	4,242	2,324	3,391	0,876	9,43	13,4	0,725
52029	8T4635	0,257	4,242	2,324	3,701	0,876	9,97	21,2	1,090
52032	8T5800	0,343	4,242	2,324	4,026	0,876	9,97	29,8	1,465
52026	8T5233	0,514	4,242	2,324	4,026	1,194	9,97	44,7	2,180
52038	8T6847	0,686	4,242	2,324	4,026	1,537	11,96	59,6	2,910
52030	8T5387	0,343	6,816	2,946	4,674	0,889	11,96	35,8	2,379
52035	8T7441	0,686	6,816	2,946	4,674	1,549	11,96	65,6	4,676
52425	8T5772	0,771	6,816	2,946	5,308	1,219	12,96	87,2	5,255
52001	8T5320	1,371	9,648	3,505	6,629	1,575	15,95	191	13,23
52028	8T4179	0,257	11,55	3,835	5,372	0,876	14,46	32,4	2,958
52017	8T4178	0,686	18,19	4,813	6,617	1,575	17,95	107	12,48
52103	8T6110	1,371	17,91	4,775	7,925	1,587	19,94	238	24,55
52022	8T8027	2,742	17,91	4,775	7,925	2,845	19,94	477	49,11
52031	8T4180	0,686	28,22	5,994	7,899	1,575	21,93	131	19,36
52128	8T6100	1,371	28,22	5,994	9,195	1,613	23,93	286	38,68
52042	8T5468	2,742	28,22	5,994	9,195	2,883	23,93	572	77,38
52100	8T5690	5,142	28,22	5,994	9,881	4,216	24,93	1117	145,0
52081	8T5737	5,142	48,69	7,874	11,811	4,242	30,91	1386	250,3
52427	8T9259	7,198	48,37	7,848	13,106	4,305	32,90	3065	348,1
52112	8T5611	6,855	75,52	9,741	13,754	5,601	36,89	2205	517,7
52426	8T9260	10,968	74,14	9,716	15,680	5,601	39,88	3814	813,2

(1) Cross-sectional area calculated for 2 mil (0,002 in.) material

(2) Dimensions listed are sizes of aluminum boxed cores (not coated)

(3) 0,002 mil thickness and high nickel material

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CHAPTER VII

TRANSFORMER - INDUCTOR

EFFICIENCY, REGULATION, AND TEMPERATURE RISE

A. INTRODUCTION

Transformer efficiency, regulation, and temperature rise are all interrelated. Not all of the input power to the transformer is delivered to the load. The difference between the input power and output power is converted into heat. This power loss can be broken down into two components: core loss and copper loss. The core loss is a fixed loss, and the copper loss is a variable loss which is related to the current demand of the load. Copper loss goes up by the square of the current and is termed quadratic loss. Maximum efficiency is achieved when the fixed loss is equal to the quadratic at rated load. Transformer regulation is the copper loss P_{cu} divided by the output power P_o.

B. TRANSFORMER EFFICIENCY

The efficiency of a transformer is a good way to measure the effectiveness of the design. Efficiency is defined as the ratio of the output power P_o to the input power P_{in}. The difference between the P_o and the P_{in} is due to losses. The total power loss in a transformer is determined by the fixed losses in the core and the quadratic losses in the windings or copper. Thus

$$P_\Sigma = P_{fe} + P_{cu} \tag{7-1}$$

where P_{fe} represents the core loss and P_{cu} represents the copper loss.

Maximum efficiency is achieved when the fixed loss is made equal to the quadratic loss as shown by equation 7-11. Transformer loss versus output load current is shown in Figure 7-1.

The copper loss increases as the square of the output power multiplied by a constant K which is thus:

$$P_{cu} = KP_o^2 \tag{7-2}$$

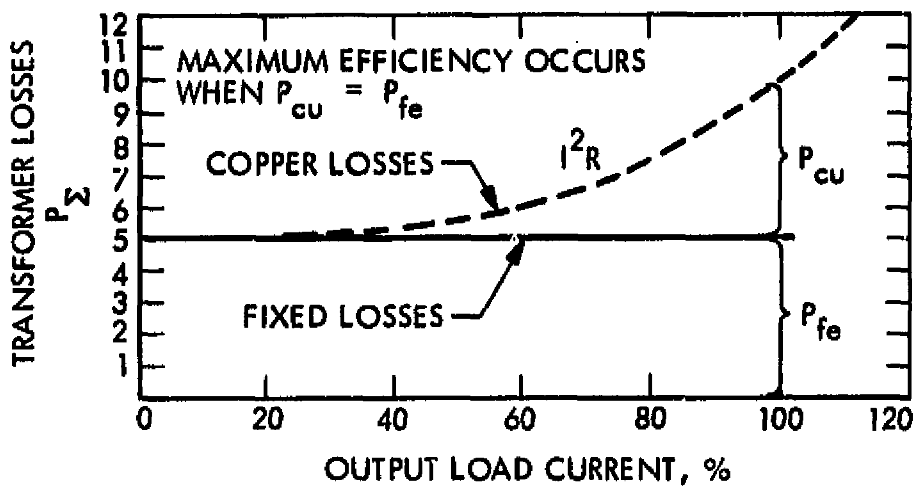

Fig. 7-1. Transformer loss versus output load current

which may be rewritten as

$$P_{\Sigma} = P_{fe} + KP_o^2 \tag{7-3}$$

Since

$$P_{in} = P_o + P_{\Sigma} \tag{7-4}$$

and the efficiency is

$$\eta = \frac{P_o}{P_o + P_{\Sigma}} \tag{7-5}$$

then

$$\eta = \frac{P_o}{P_o + P_{fe} + KP_o^2} = \frac{P_o}{P_{fe} + P_o + KP_o^2} \tag{7-6}$$

and, differentiating with respect to P_o:

$$\frac{d\eta}{dP_o} = -P_o \left[P_{fe} + P_o + KP_o^2\right]^{-2} (1 + 2\ KP_o) \tag{7-7}$$

$$+ \left[P_{fe} + P_o + KP_o^2\right] = O \text{ for max } \eta \tag{7-8}$$

$$-P_o\ (1 + 2\ KP_o) + \left(P_{fe} + P_o + KP_o^2\right) = O \tag{7-9}$$

$$-P_o - 2KP_o^2 + P_{fe} + P_o + KP_o^2 = O \tag{7-10}$$

$$\therefore\ P_{fe} = KP_o^2 = P_{cu} \tag{7-11}$$

C. RELATIONSHIP OF A_p TO CONTROL OF TEMPERATURE RISE

1. Temperature Rise

Not all of the P_{in} input power to the transformer is delivered to the load as the P_o. Some of the input power is converted to heat by hysteresis and eddy currents induced in the core material, and by the resistance of the windings. The first is a fixed loss arising from core excitation and is termed "core loss." The second is a variable loss in the windings which is related to the current demand of the load and thus varies as I^2R. This is termed the quadratic or copper loss.

The heat generated produces a temperature rise which must be controlled to prevent damage to or failure of the windings by breakdown of the wire insulation at elevated temperatures. This heat is dissipated from the exposed surfaces of the transformer by a combination of radiation and convection. The dissipation is therefore dependent upon the total exposed surface area of the core and windings.

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Ideally, maximum efficiency is achieved when the fixed and quadratic losses are equal. Thus:

$$P_\Sigma = P_{fe} + P_{cu} \tag{7-12}$$

and

$$P_{cu} = \frac{P_\Sigma}{2} \tag{7-13}$$

When the copper loss in the primary winding is equal to the copper loss in the secondary, the current density in the primary is the same as the current density in the secondary:

$$\frac{P_p}{R_p} = \frac{P_s}{R_s} \tag{7-14}$$

and

$$\frac{P_\Sigma}{R_t} = \frac{2P_p}{R_p/2} = \frac{4P_p}{R_p} = (2I_p)^2 \tag{7-15}$$

Then

$$J_p = \frac{I_p}{W_a/2} = \frac{2I_p}{W_a} = J_s = J \tag{7-16}$$

2. Calculation of Temperature Rise

Temperature rise in a transformer winding cannot be predicted with complete precision, despite the fact that many different techniques are described in the literature for its calculation. One reasonably accurate method for open core and winding construction is based upon the assumption that core and winding losses may be lumped together as:

$$P_{\Sigma} = P_{fe} + P_{cu} \qquad (7\text{-}17)$$

and the assumption is made that thermal energy is dissipated uniformly throughout the surface area of the core and winding assembly.

Transfer of heat by thermal radiation occurs when a body is raised to a temperature above its surroundings and emits radiant energy in the form of waves. In accordance with the Stefan-Boltzmann law,* this may be expressed as:

$$W_r = K_r \epsilon \, (T2^4 - T1^4) \qquad (7\text{-}18)$$

in which

W_r = watts per square centimeter of surface

K_r = 5.70×10^{-12} W cm^{-2} (^{o}K)$^{-4}$

ϵ = emissivity factor

T2 = hot body temperature in degrees kelvin

T1 = ambient or surrounding temperature in degrees kelvin

Transfer of heat by convection occurs when a body is hotter than the surrounding medium, which usually is air. The layer of air in contact with the hot body which is heated by conduction expands, and rises, taking the absorbed heat with it. The next layer, being colder, replaces the risen layer, and in turn on being heated also rises. This continues as long as the air or medium surrounding the body is at a lower temperature. The transfer of heat by convection is stated mathematically as:

$$W_c = K_c F \theta^{\eta} \sqrt{P} \qquad (7\text{-}19)$$

*Reference 2, Chapter 3.

in which:

W_c = watts loss per square centimeter

K_c = 2.17×10^{-4}

F = air friction factor (unity for a vertical surface)

θ = temperature rise, degrees C

p = relative barometric pressure (unity at sea level)

η = exponential value ranging from 1.0 to 1.25, depending on the shape and position of the surface being cooled.

The total heat dissipated from a plane vertical surface is expressed by the sum of equations 7-18 and 7-19:

$$W = 5.70 \times 10^{-12} \epsilon (T2^4 - T1^4) + 1.4 \times 10^{-3} F\theta^{1.25} \sqrt{p} \qquad (7\text{-}20)$$

3. Temperature Rise Versus Surface Area Dissipation

The temperature rise which may be expected for various levels of power loss is shown in the monograph of Figure 7-2 below. It is based on equation 7-20 relying on data obtained from Reference 2* for heat transfer effected by a combination of 55% radiation and 45% convection, from surfaces having an emissivity of 0.95, in an ambient temperature of 25°C, at sea level. Power loss (heat dissipation) is expressed in watts/cm^2 of total surface area. Heat dissipation by convection from the upper side of a horizontal flat surface is on the order of 15 to 20% more than from vertical surfaces. Heat dissipation from the underside of a horizontal flat surface depends upon surface area and conductivity.

*See References in Chapter 3.

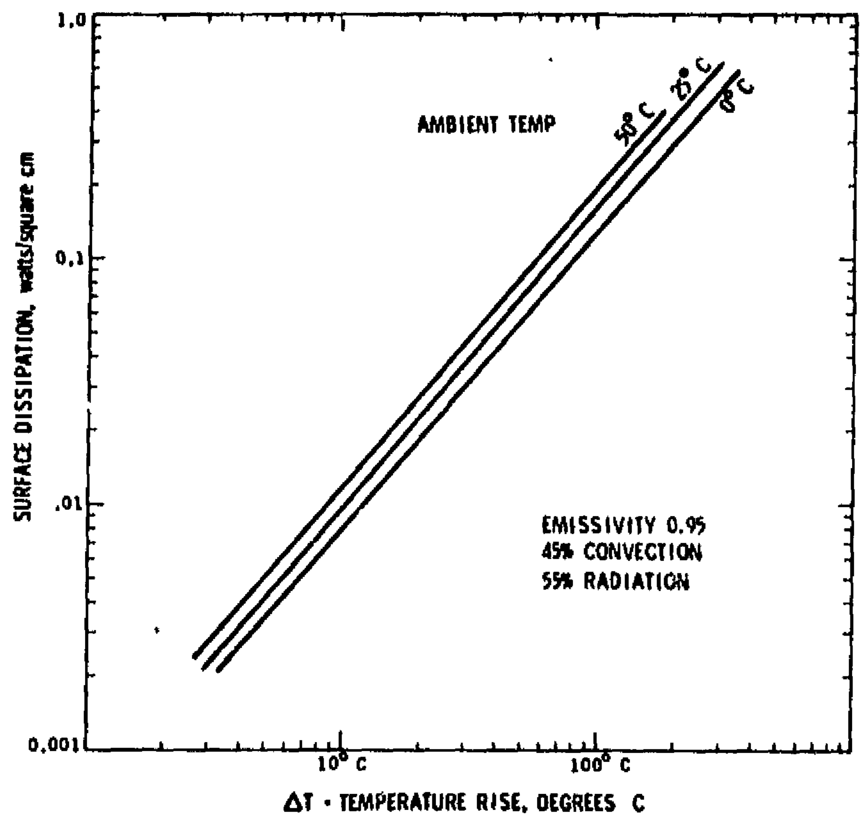

Fig. 7-2. Temperature rise versus surface dissipation

4. Surface Area Required for Heat Dissipation

The effective surface area A_t required to dissipate heat (expressed as watts dissipated per unit area) is:

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$$A_t = \frac{P_\Sigma}{\Psi} \qquad (7\text{-}21)$$

in which Ψ is the power density or the average power dissipated per unit area from the surface of the transformer and P_Σ is the total power lost or dissipated.

Surface area A_t of a transformer can be related to the area product A_p of a transformer. The straightline logarithmic relationship shown in Figure 7-3 below has been plotted from the data shown in Table 2-5, Chapter 2.

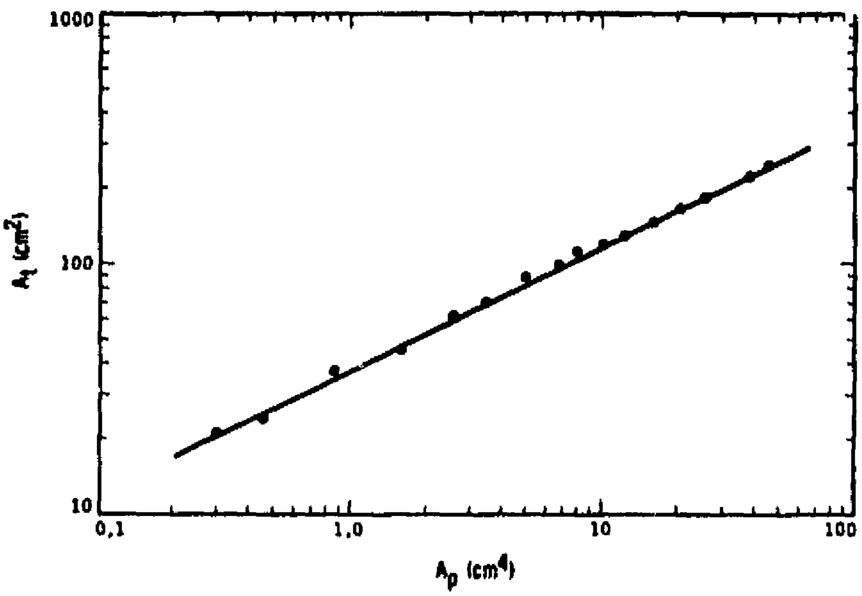

Fig. 7-3. Surface area versus area product A_p

From this, the following relationship evolves:

$$A_t = K_s(A_p)^{0.5} = \frac{P_\Sigma}{\Psi} \quad (7\text{-}22)$$

and (from Fig. 7-2)

$$\Psi = 0.03\ \text{W/cm}^2 \text{ at } 25°\text{C rise} \quad (7\text{-}23)$$

$$\Psi = 0.07\ \text{W/cm}^2 \text{ at } 50°\text{C rise} \quad (7\text{-}24)$$

Figure 7-4 utilizes the efficiency rating in watts dissipated in terms of two different, but commonly allowable temperature rises for the transformer over ambient temperature. The data presented are used as bases for determining the needed transformer surface area A_t (in cm^2).

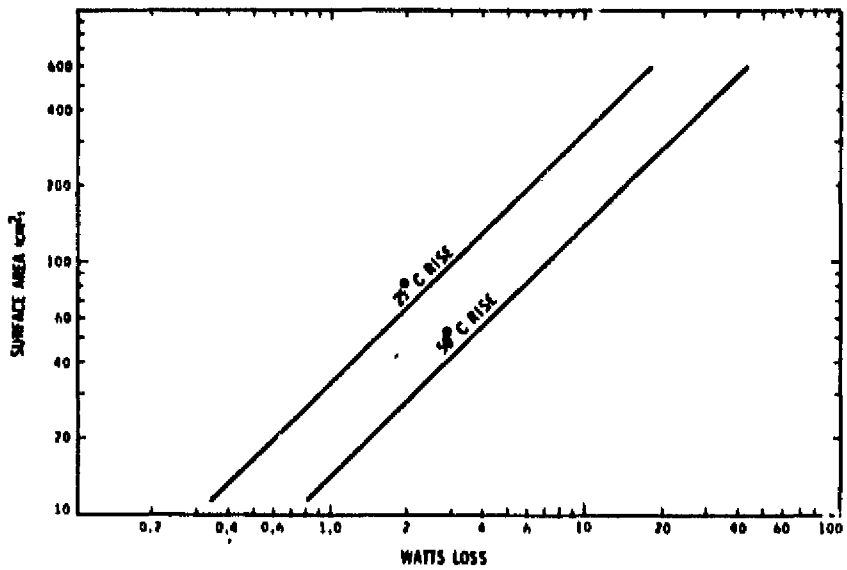

Fig. 7-4. Surface area versus total watt loss for a 25°C and 50°C rise

D. REGULATION AS A FUNCTION OF EFFICIENCY

The minimum size of a transformer is usually determined either by a temperature rise limit, or by allowable voltage regulation, assuming that size and weight are to be minimized.

Figure 7-5 shows the circuit diagram of a transformer with one secondary. Note that α = regulation (%).

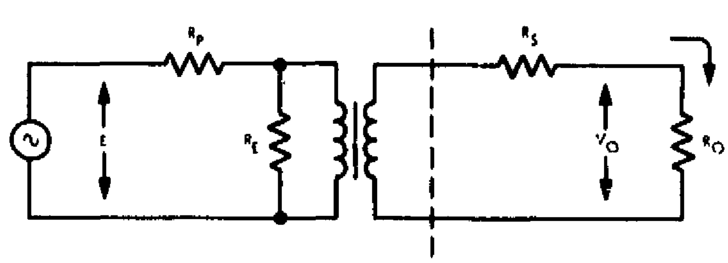

Fig. 7-5. Transformer circuit diagram

The analytical equivalent is shown in Figure 7-6.

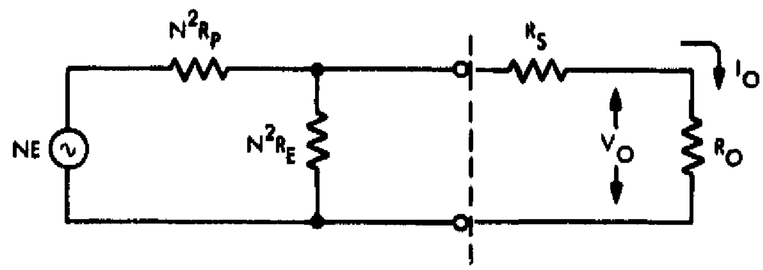

Fig. 7-6. Transformer analytical equivalent

This assumes that distributed capacitance in the secondary can be neglected because the frequency and secondary voltage are not excessive high. Also the winding geometry is designed to limit the leakage inductance to a level low enough to be neglected under most operating conditions.

Transformer voltage regulation can now be expressed as:

$$\alpha = \frac{V_o(N.L.) - V_o(F.L.)}{V_o(N.L.)} \times 100 \qquad (7\text{-}25)$$

in which $V_o(N.L.)$ is the no load voltage and $V_o(F.L.)$ is the full load voltage.

The output voltage computed using Figure 7-5 is:

$$V_o = \frac{R_o}{R_o + R_s} \frac{\left(N^2R_p\right) \parallel \left(N^2R_E\right) \parallel (R_o + R_s)}{N^2R_p} NE \qquad (7\text{-}26)$$

For the usual condition of

$$N^2R_E \gg N^2R_p \parallel (R_o + R_s),$$

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V_o simplifies to

$$V_o = V_o\,(\text{F.L.}) = \frac{R_o}{R_o + \left(N^2 R_p + R_s\right)} NE \tag{7-27}$$

For equal window areas allocated for the primary and secondary windings, it can be shown that $N^2 R_p = R_s$.

For simplicity, let

$$R_{cu} \equiv N^2 R_p + R_s = 2R_s$$

At no load (N. L.) R_o approaches infinity, therefore:

$$V_o\,(\text{N.L.}) = NE \tag{7-28}$$

$$\alpha = \frac{NE - \dfrac{R_o}{R_o + R_{cu}} NE}{NE} \times 100 \tag{7-29}$$

$$= \left(1 - \frac{R_o}{R_o + R_{cu}}\right) \times 100 \tag{7-30}$$

$$= \frac{R_{cu}}{R_o + R_{cu}} \times 100 \tag{7-31}$$

This shows that regulation is independent of the transformer turns ratio.

For regulation as a function of copper loss, multiply equation 7-31 by I_o^2:

$$\alpha = \frac{I_o^2 R_{cu}}{I_o^2 (R_o + R_{cu})} \times 100 \tag{7-32}$$

then

$$\alpha = \frac{P_{cu}}{P_o + P_{cu}} \times 100 \qquad (7\text{-}33)$$

$$P_{in} = P_{cu} + P_{fe} + P_o \qquad (7\text{-}34)$$

For regulation as a function of efficiency,

$$\frac{P_o}{P_{in}} = \frac{P_o}{P_{cu} + P_{fe} + P_o} = \eta \qquad (7\text{-}35)$$

By definition

$$P_{cu} = P_{fe} \qquad (7\text{-}36)$$

Solving for $P_{cu} + P_{fe}$

$$\frac{P_o(1 - \eta)}{\eta} = P_o\left(\frac{1}{\eta} - 1\right) = P_{cu} + P_{fe} = 2\,P_{cu} \qquad (7\text{-}37)$$

$$\frac{\alpha}{100} = \frac{1}{1 + \frac{P_o}{P_{cu}}} = \frac{1}{1 + \frac{2}{1/\eta - 1}} = \frac{1 - \eta}{1 + \eta} \qquad (7\text{-}38)$$

$$\alpha = \frac{1 - \eta}{1 + \eta} \times 100 \qquad (7\text{-}39)$$

For efficiency as a function of regulation, multiply both sides of the equation by $(1 + \eta)$:

$$\alpha + \eta\,\alpha = 100 - \eta\,100 \qquad (7\text{-}40)$$

Solve for η

$$\eta 100 + \eta \alpha = 100 - \alpha \tag{7-41}$$

$$\eta (100 + \alpha) = 100 - \alpha \tag{7-42}$$

$$\eta = \frac{100 - \alpha}{100 + \alpha} \tag{7-43}$$

E. DESIGNING FOR A GIVEN REGULATION

1. Transformers

Although most transformers are designed for a given temperature rise, they can also be designed for a given regulation.* The regulation and power-handling ability of a core is related to two constants:

$$VA = K_g K_e \alpha \tag{7-44}$$

$$\alpha = \text{Regulation (\%)}$$

The constant K_g is determined by the core geometry which may be related by the following equation:

$$K_g = \frac{W_a A_c^2 K_u}{MLT} \tag{7-45}$$

The constant K_e is determined by the magnetic and electric operating conditions which may be related by the following equation:

$$K_e = 0.145 K^2 f^2 B_m^2 \times 10^{-4} \tag{7-46}$$

The derivation of the relationship for K_g and K_e is given at the end of this chapter.

*Reference

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The area product A_p can be related to the core geometry K_g in the following equation:

$$A_p = K_p K_g^{0.8} \tag{7-47}$$

The derivation is given in detail at the end of this chapter.

Rewriting equation 7-44,

$$K_g = \frac{VA}{K_e \alpha} \tag{7-48}$$

$$A_p = K_p \left(\frac{VA}{K_e \alpha}\right)^{0.8} \tag{7-49}$$

Figure 7-7 shows how area product A_p varies as a function of regulation, in percent.

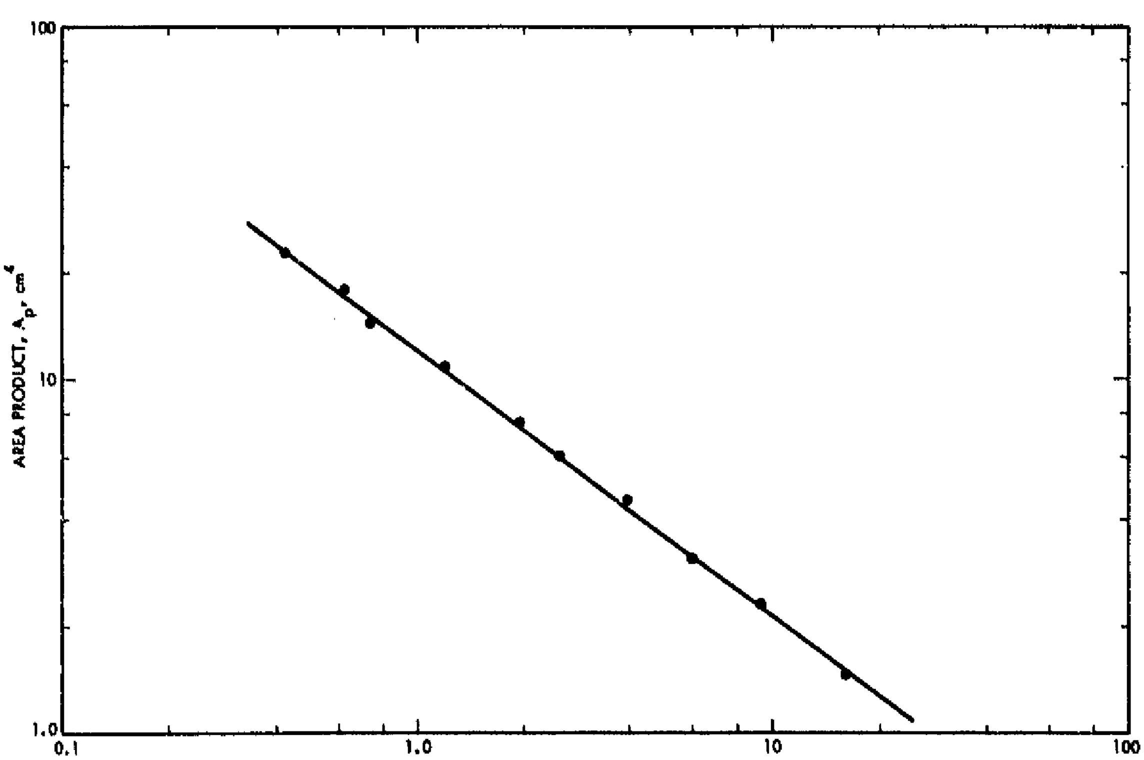

Fig. 7-7. Area product versus regulation

Figure 7-8 shows how weight W_t varies as a function of regulation, in percent.

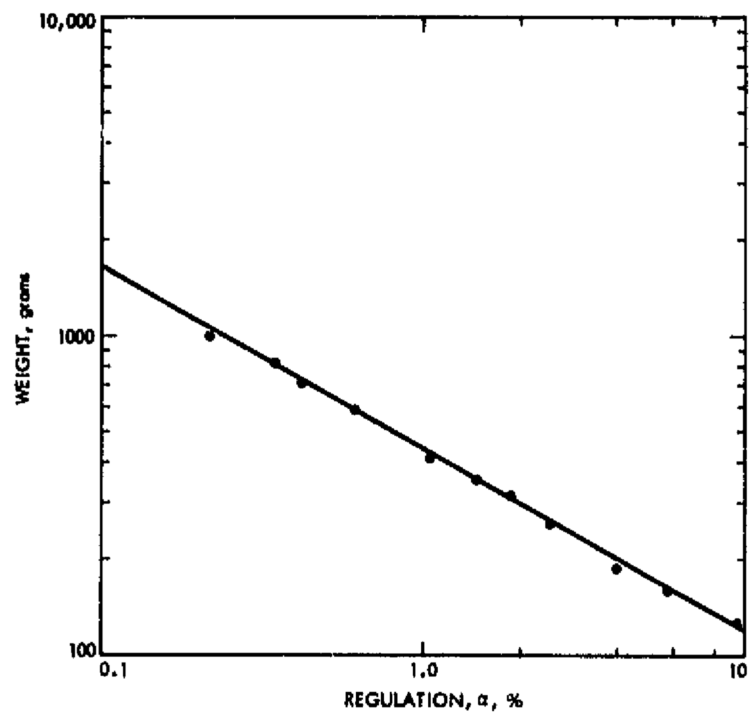

Fig. 7-8. Weight versus regulation

2. Inductors

Inductors, like transformers, are designed for a given temperature rise. They can also be designed for a given regulation. The regulation and energy handling ability of a core is related to two constants:

$$(\text{Energy})^2 = K_g K_e \alpha \tag{7-50}$$

α = Regulation (%)

The constant K_g is determined by the core geometry:

$$K_g = \frac{W_a A_c^2 K_u}{MLT} \qquad (7\text{-}51)$$

The constant K_e is determined by the magnetic and electric operating conditions:

$$K_e = 0.145 \, P_o \, B_{dc}^2 \times 10^{-4} \qquad (7\text{-}52)$$

The derivation of the specific functions for K_g and K_e is given at the end of this chapter.

3. Transformer Design Example I

For a typical design example, assume an isolation transformer with the following specifications:

(1) 115 volts

(2) 1.0 amperes

(3) Sine wave

(4) Frequency 60 Hz

(5) Regulation α 2%

The procedure would then be as follows:

Step No. 1. Calculate the output power:

$$P_o = VA$$

$$P_o = (115)(1.0)$$

$$P_o = 115 \qquad [\text{watts}]$$

Step No. 2. Calculate the electrical conditions from equation 7-46:

$$K_e = 0.145\, K^2 f^2 B_m^2 \times 10^{-4}$$

$K = 4.44$

$f = 60$ [Hz]

$B = 1.2$ [tesla]

$K_e = 0.145(4.44)^2(60)^2(1.2)^2 \times 10^{-4}$

$K_e = 1.53$

Step No. 3. Calculate the core geometry from equation 7-44:

$$K_g = \frac{VA}{K_e \alpha}$$

$$K_g = \frac{115}{(1.53)(2.0)}$$

$$K_g = 37.6$$

Step No. 4. Select a lamination from Table 7.B-2 with a value K_g closest to the one calculated:

EI - 150 with a $K_g = 35.3$

Step No. 5. Calculate the number of primary turns using Faraday's law, equation 3.A-1:

$$N = \frac{E_p \times 10^4}{4.44\, A_c B_m f}$$

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The iron cross section A_c is found in Table 7.B-2:

$$A_c = 13.1 \qquad \left[cm^2\right]$$

$$N = \frac{115 \times 10^4}{4.44(13.1)(1.2)(60)}$$

$$N = 275 \text{ turns}$$

Step No. 6. Calculate the effective window area $W_{a(eff)}$:

$$W_{a(eff)} = W_a S_3$$

A typical value for S_3 is 0.75, as shown in Chapter 6.

Select the window area W_a from Table 7.B-2 for EI 150:

$$W_{a(eff)} = (10.9(0.75)$$

$$W_{a(eff)} = 8.175 \qquad \left[cm^2\right]$$

Step No. 7. Calculate the primary winding area:

Primary winding area = Secondary winding area

$$\text{Primary winding area} = \frac{W_{a(eff)}}{2}$$

$$\text{Primary winding area} = \frac{8.175}{2}$$

$$\text{Primary winding area} = 4.09 \qquad \left[cm^2\right]$$

Step No. 8. Calculate the wire area A_w with insulation, using a fill factor S_2 of 0.6:

$$A_w = \frac{W_{a(pri)}}{N} \times S_2$$

$$A_w = \frac{(4.09)(0.6)}{275}$$

$$A_w = 0.00892 \qquad [cm^2]$$

Step No. 9. Select the wire area A_w with insulation in Table 6-1 for equivalent (AWG) wire size column D:

$$\text{AWG No. } 18 = 0.009326 \qquad [cm^2]$$

The rule is that when the calculated wire size does not fall close to those listed in the table, the next smaller size should be selected.

Step No. 10. Calculate the resistance of the winding using Table 6-1, column C, and Table 7.B-2 for the MLT:

$$R = MLT \times N \times (\text{column C}) \times 10^{-6}$$

$$R_p = (21.2)(275)(209.5) \times 10^{-6}$$

$$R_p = 1.22 \qquad [\Omega]$$

$$R_p = R_s$$

$$R_t = 2\,R_p$$

$$R_t = 2\,(1.22)$$

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$$R_t = 2.44 \qquad [\Omega]$$

Step No. 11. Calculate the copper loss P_{cu} and the regulation:

$$P_{cu} = I^2 R_t$$

$$P_{cu} = (1)^2 (2.44)$$

$$P_{cu} = 2.44 \quad \text{[watts]}$$

$$\alpha = \frac{P_{cu}}{P_o} \times 100$$

$$\alpha = \frac{2.44}{115} \times 100$$

$$\alpha = 2.12 \quad [\%]$$

4. Transformer Design Example II

For a typical design example, assume a filament transformer using a C core:

(1) 120 volt input

(2) 400 Hz

(3) Sine wave

(4) 6.3 volt output

(5) 5.0 ampere output

(6) Regulation α 1.0%

The procedure would then be as follows:

Step No. 1. Calculate the output power:

$$P_o = VA$$

$$P_o = (6.3)(5)$$

$$P_o = 31.5 \quad \text{[watts]}$$

Step No. 2. Calculate the electrical conditions from equation 7-46:

$$K_e = 0.145\ K^2 f^2 B_m^2 \times 10^{-4}$$

$$K = 4.44$$

$$f = 400 \quad [\text{Hz}]$$

$$B = 1.2 \quad [\text{tesla}]$$

$$K_e = (0.145)(4.44)^2(400)^2(1.2)^2 \times 10^{-4}$$

$$K_e = 65.8$$

Step No. 3. Calculate the core geometry from equation 7-44:

$$K_g = \frac{VA}{K_e \alpha}$$

$$K_g = \frac{31.5}{(65.8)(1)}$$

$$K_g = 0.479$$

Step No. 4. Select a C core from Table 7.B-1 with a value K_g closest to the one calculated:

AL-18 with a $K_g = 0.530$

Step No. 5. Calculate the number of primary turns using Faraday's law, equation 3.A-1,

$$N = \frac{E_p \times 10^4}{4.44\ A_c\ B_m\ f}$$

The iron cross section A_c is found in Table 7.B-1:

$$A_c = 1.257 \quad [\text{cm}^2]$$

$$N_p = \frac{120 \times 10^4}{4.44(1.257)(1.2)(400)}$$

$$N_p = 448$$

Step No. 6. Calculate the effective window area $W_{a(eff)}$:

$$W_{a(eff)} = W_a S_3$$

A typical value for S_3 is 0.75 as shown in Chapter 6. Select the window area W_a from Table 7.B-1 for AL-18:

$$W_{a(eff)} = (6.3)(0.75)$$

$$W_{a(eff)} = 4.72 \qquad [cm^2]$$

Step No. 7. Calculate primary winding area:

Primary winding area = Secondary winding area

$$\text{Primary winding area} = \frac{W_{a(eff)}}{2}$$

$$\text{Primary winding area} = \frac{4.72}{2}$$

$$\text{Primary winding area} = 2.36 \qquad [cm^2]$$

Step No. 8. Calculate the wire area A_w with insulation using a fill factor S_2 of 0.6:

$$A_w = \frac{W_{a(pri)} S_2}{N}$$

$$A_w = \frac{(2.36)(0.6)}{448}$$

$$A_w = 0.00316 \qquad [cm^2]$$

<u>Step No. 9.</u> **Select the wire area A_w with insulation in Table 6-1 for equivalent (AWG) wire size, column D:**

$$\text{AWG No. 23} = 0.003135 \qquad [cm^2]$$

The rule is that when the calculated wire size does not fall close to those listed in the table, the next smaller size should be selected.

<u>Step No. 10.</u> Calculate the resistance of the primary winding, using Table 6-1, column C, and Table 7.B-1 for the MLT:

$$R_p = MLT \times N \times (\text{column C}) \times 10^{-6}$$

$$R_p = (7.51)(448)(666) \times 10^{-6}$$

$$R_p = 2.24 \qquad [\Omega]$$

<u>Step No. 11.</u> Calculate the primary copper loss P_{cu}:

$$I_p = \frac{VA}{E_p}$$

$$I_p = \frac{31.5}{120} = 0.263 \qquad [A]$$

$$P_{cu} = I_p^2 R_p$$

$$P_{cu} = (0.263)^2(2.24)$$

$$P_{cu} = 0.155 \qquad [watts]$$

Step No. 12. Calculate the secondary turns:

$$N_s = \frac{N_p}{E_p}(E_s)$$

$$N_s = \frac{448}{120}(6.3)$$

$$N_s = 24$$

Step No. 13. Calculate the secondary wire area A_w with insulation using a fill factor S_2 of 0.6:

$$A_w = \frac{W_{a(sec)} S_2}{N}$$

$$A_w = \frac{(2.36)(0.6)}{24}$$

$$A_w = 0.059 \quad [cm^2]$$

Step No. 14. Select the secondary wire area A_w with insulation in Table 6-1 for equivalent (AWG) wire size, column D:

$$\text{AWG No. } 10 = 0.0559 \quad [cm^2]$$

The rule is that when the calculated wire size does not fall close to those listed in the table, the next smaller size should be selected.

Step No. 15. Calculate the resistance of the secondary winding using Table 6-1, column C, and Table 7.B-1 for the MLT:

$$R_s = MLT \times N \times (\text{column C}) \times 10^{-6}$$

$$R_s = (7.51)(24)(32.7) \times 10^{-6}$$

$$R_s = 0.0059 \qquad [\Omega]$$

Step No. 16. Calculate the copper loss P_{cu}:

$$P_{cu} = I_s^2 R_s$$

$$P_{cu} = (5)^2 (0.0059)$$

$$P_{cu} = 0.148 \qquad [\text{watts}]$$

Step No. 17. Calculate the regulation:

$$\alpha = \frac{P_p + P_s}{P_o} \times 100$$

$$\alpha = \frac{(0.155) + (0.148)}{31.5} \times 100$$

$$\alpha = 0.962 \qquad [\%]$$

5. Inductor Design Example

For a typical design example, assume:

(1) Inductance = 0.05 henry

(2) Output power P_o = 200 watts

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(3) Output current I_o = 2.0 amperes

(4) Regulation $\alpha = 1\%$

The procedure would then be as follows:

Step No. 1. Calculate the energy involved from equation 7.B-16:

$$\text{Energy} = \frac{L\, I_o^2}{2}$$

$$\text{Energy} = \frac{0.05(2.0)^2}{2}$$

$$\text{Energy} = 0.10 \qquad [\text{watt seconds}]$$

Step No. 2. Calculate the electrical conditions from equation 7-52:

$$K_e = 0.145\, P_o\, B_{dc}^2 \times 10^{-4}$$

$$P_o = 200 \qquad [\text{watts}]$$

$$B_{dc} = 1.2 \qquad [\text{tesla}]$$

$$K_e = 0.145(200)(1.2)^2 \times 10^{-4}$$

$$K_e = 0.00418$$

Step No. 3. Calculate the core geometry from equations 7-50:

$$K_g = \frac{(\text{Energy})^2}{K_e \alpha}$$

$$K_g = \frac{(0.1)^2}{(0.00418)(1)}$$

$$K_g = 2.39$$

Step No. 4. Select a C core from Table 7.B-1 with a value K_g closest to the one calculated:

AL-20 with a $K_g = 2.32$

Also select the area product A_p for this C core from Table 2-6:

$$A_p = 22.6 \qquad [\text{cm}^4]$$

Step No. 5. Calculate the current density from area product equation 4.A-18:

$$J = \frac{2\,(\text{Energy}) \times 10^4}{B_m K_u A_p}$$

Insert values, $K_u = 0.4$,

$$J = \frac{2(0.1) \times 10^4}{(1.2)(0.4)(22.6)}$$

$$J = 184 \qquad [\text{A/cm}^2]$$

Step No. 6. Determine the bare wire size A_w:

$$A_{w(B)} = \frac{I_o}{J} = \frac{2.0}{184}$$

$$A_{w(B)} = 0.0108 \qquad \left[cm^2\right]$$

Step No. 7. Select an AWG wire size from Table 6-1, column A. The rule is that when the calculated wire size does not fall close to those listed in the table, the next smaller size should be selected.

$$AWG\ 17 = 0.01038 \qquad \left[cm^2\right]$$

Step No. 8. Calculate the effective window area $W_{a(eff)}$:

$$W_{a(eff)} = W_a S_3$$

A typical value for S_3 is 0.75, as shown in Chapter 6.

Select the window area W_a from Table 7.B-1 for an AL-20:

$$W_{a(eff)} = (6.30)(0.75)$$

$$W_{a(eff)} = 4.72 \qquad \left[cm^2\right]$$

Step No. 9. Select the wire area with insulation for a No. 17 in Table 6-1, column D:

$$A_w \text{ with insulation} = 0.01168 \qquad \left[cm^2\right]$$

Step No. 10. Calculate the number of turns using a fill factor S_2 of 0.6:

$$N = \frac{W_{a(eff)} S_2}{A_w}$$

$$N = \frac{(4.72)(0.6)}{(0.01168)}$$

$$N = 242$$

Step No. 11. Calculate the gap from the inductance equation 4-6:

$$l_g = \frac{0.4\pi N^2 A_c \times 10^{-8}}{L}$$

The iron cross section A_c is found in Table 7.B-1:

$$A_c = 3.58$$

$$l_g = \frac{1.26(242)^2(3.58) \times 10^{-8}}{(0.05)}$$

$$l_g = 0.0528 \qquad \text{[cm]}$$

Step No. 12. Calculate the amount of fringing flux from equation 4-7 (the value for G is found in Table 4.B-17):

$$F = \left(1 + \frac{l_g}{\sqrt{A_c}} \log_e \frac{2G}{l_g}\right)$$

$$F = \left(1 + \frac{(0.0528)}{\sqrt{3.58}} \log_e \frac{2(3.967)}{(0.0528)}\right)$$

$$F = 1.14$$

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After finding the fringing flux F, insert it into equation 4-8, rearrange, and solve for the correct number of turns:

$$N = \sqrt{\frac{l_g L}{0.4\pi\ A_c\ F \times 10^{-8}}}$$

$$N = \sqrt{\frac{(0.0528)(0.05)}{(1.26)(3.58)(1.14) \times 10^{-8}}}$$

$$N = 226$$

<u>Step No. 13.</u> Calculate the resistance of the winding, using wire Table 6-1, column C and Table 7.B-1 for the MLT:

$$R = MLT \times N \times (\text{column C}) \times 10^{-6}$$

$$R = (13.62)(226)(165.8) \times 10^{-6}$$

$$R = 0.51 \qquad [\Omega]$$

<u>Step No. 14.</u> Calculate the power loss in the winding:

$$P_{cu} = I_o^2 R$$

$$P_{cu} = (2)^2(0.51)$$

$$P_{cu} = 2.04 \qquad [\text{watts}]$$

Step No. 15. Calculate the regulation from equation 7.B-23:

$$\alpha = \frac{P_{cu}}{P_o} \times 100$$

$$\alpha = \frac{2.04}{200} \times 100$$

$$\alpha = 1.02 \qquad [\%]$$

Step No. 16. Calculate the flux density for B_{dc} from equation 7.B-7:

$$B_{dc} = \frac{0.4\pi \; N \, I_{dc} \times 10^{-4}}{l_g}$$

$$B_{dc} = \frac{(1.26)(226)(2.0) \times 10^{-4}}{(0.0528)}$$

$$B_{dc} = 1.08 \qquad [\text{tesla}]$$

(In a test sample made to verify this example, the measured inductance was found to be 0.047 henry and the resistance was 0.45 ohms.)

F. MAGNETIC CORE MATERIAL TRADEOFF

The relationships between area product A_p and certain parameters are associated only with such geometric properties as surface area and volume, weight, and the factors affecting temperature rise such as current density. A_p has no relevance to the magnetic core materials used, however the designer often must make tradeoffs between such goals as efficiency and size which are influenced by core material selection.

Usually in articles written about inverter and converter transformer design, recommendations with respect to choice of core material are a compromise of material characteristics such as those tabulated in Table 7-1, and graphically displayed in Figure 7-9. The characteristics shown here are those typical of commercially available core materials. As can be seen, the core material which provides the highest flux density is supermendor. It also produces the smallest component size. If size is the most important consideration, this should determine the choice of materials. On the other hand, the type 78 Supermalloy material (see the 5/78 curve in Figure 7-9), has the lowest flux density and this material would result in the largest size transformer. However, this material has the lowest coercive force and lowest core loss of any of the available materials. These factors might well be decisive in other applications.

Table 7-1. Magnetic core material characteristics

TRADE NAMES	COMPOSITION	* SATURATED FLUX DENSITY, tesla	DC COERCIVE FORCE, AMP-TURN/cm	SQUARENESS RATIO	** MATERIAL DENSITY, g/cm^3	CURIE TEMPERATURE, °C	WEIGHT FACTOR
Supermendur Permendur	49% Co 49% Fe 2% V	1.9-2.2	0.18-0.44	0.90-1.0	8.15	930	1.066
Magnesil Silectron Microsil Supersil	3% Si 97% Fe	1.5-1.8	0.5-0.75	0.85-0.75	7.63	750	1.00
Deltamax Orthonol 49 Sq Mu	50% Ni 50% Fe	1.4-1.6	0.125-0.25	0.94-1.0	8.24	500	1.079
Allegheny 4750 48 Alloy Carpenter 49	48% Ni 52% Fe	1.15-1.4	0.062-0.187	0.80-0.92	8.19	480	1.073
4-79 Permalloy Sp Permalloy 80 Sq Mu 79	79% Ni 17% Fe	0.66-0.82	0.025-0.82	0.80-1.0	8.73	460	1.144
Supermalloy	78% Ni 17% Fe 5% Mo	0.65-0.82	0.0037-0.01	0.40-0.70	8.76	400	1.148
Ferrites F N27 3C8	Mn Zn	0.45-0.50	0.25	0.30-0.5	4.8	250	0.629

*tesla = 10^4 Gauss
**g/cm^3 = 0.036 lb/in^3

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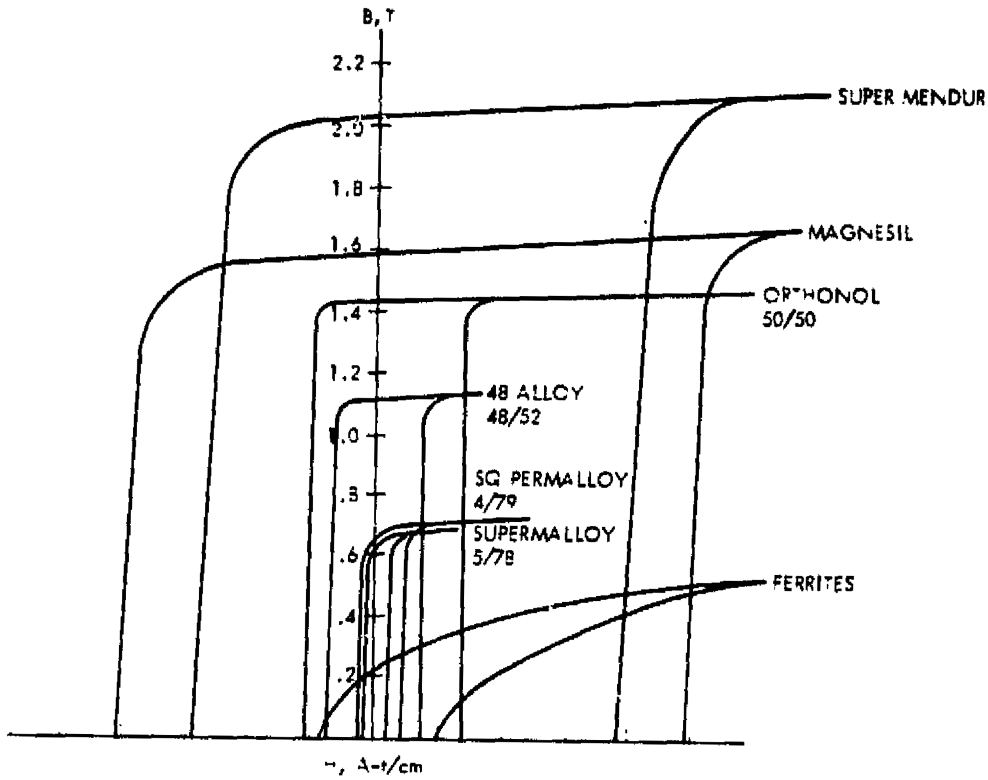

Fig. 7-9. The typical dc B-H loops of magnetic material

Choice of core material is thus based upon achieving the best characteristic for the most critical or important design parameter, with acceptable compromises on all other parameters. Figures 7-10 through 7-17 compare the core loss of different magnetic materials as a function of flux density, frequency and material thickness.

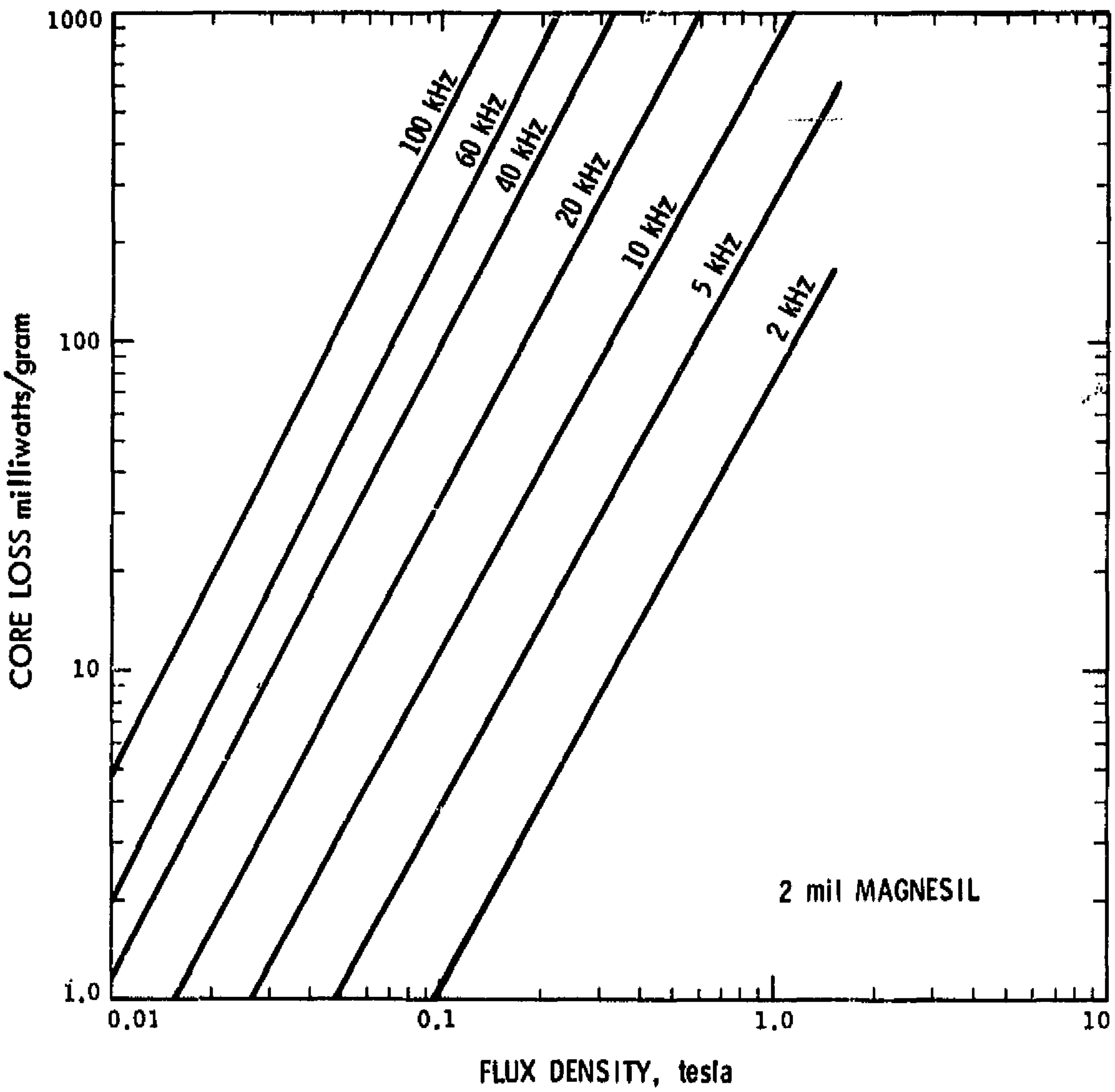

Fig. 7-10. Design curves showing maximum core loss for 2 mil silicon

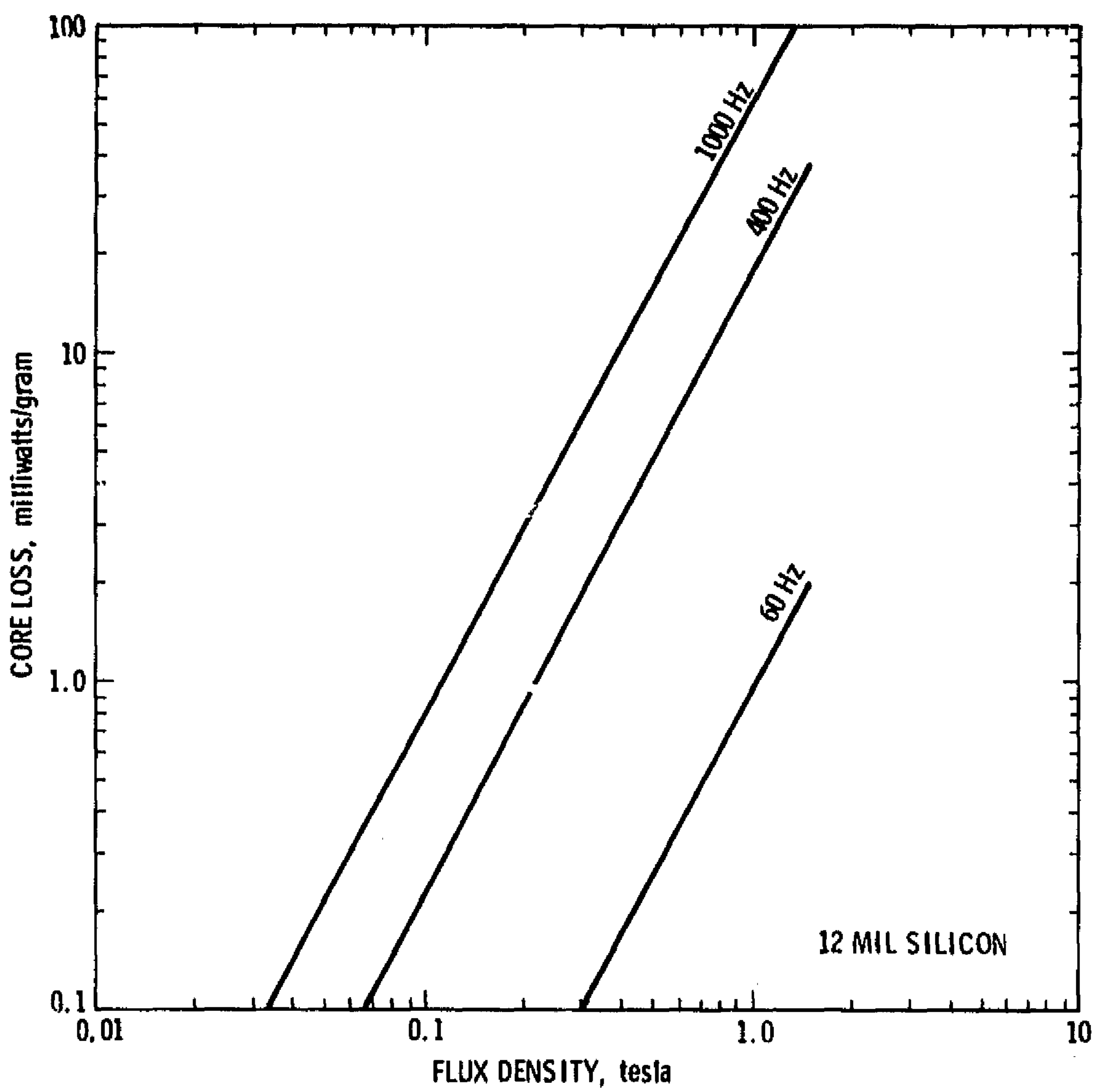

Fig. 7-11. Design curves showing maximum core loss for 12 mil silicon

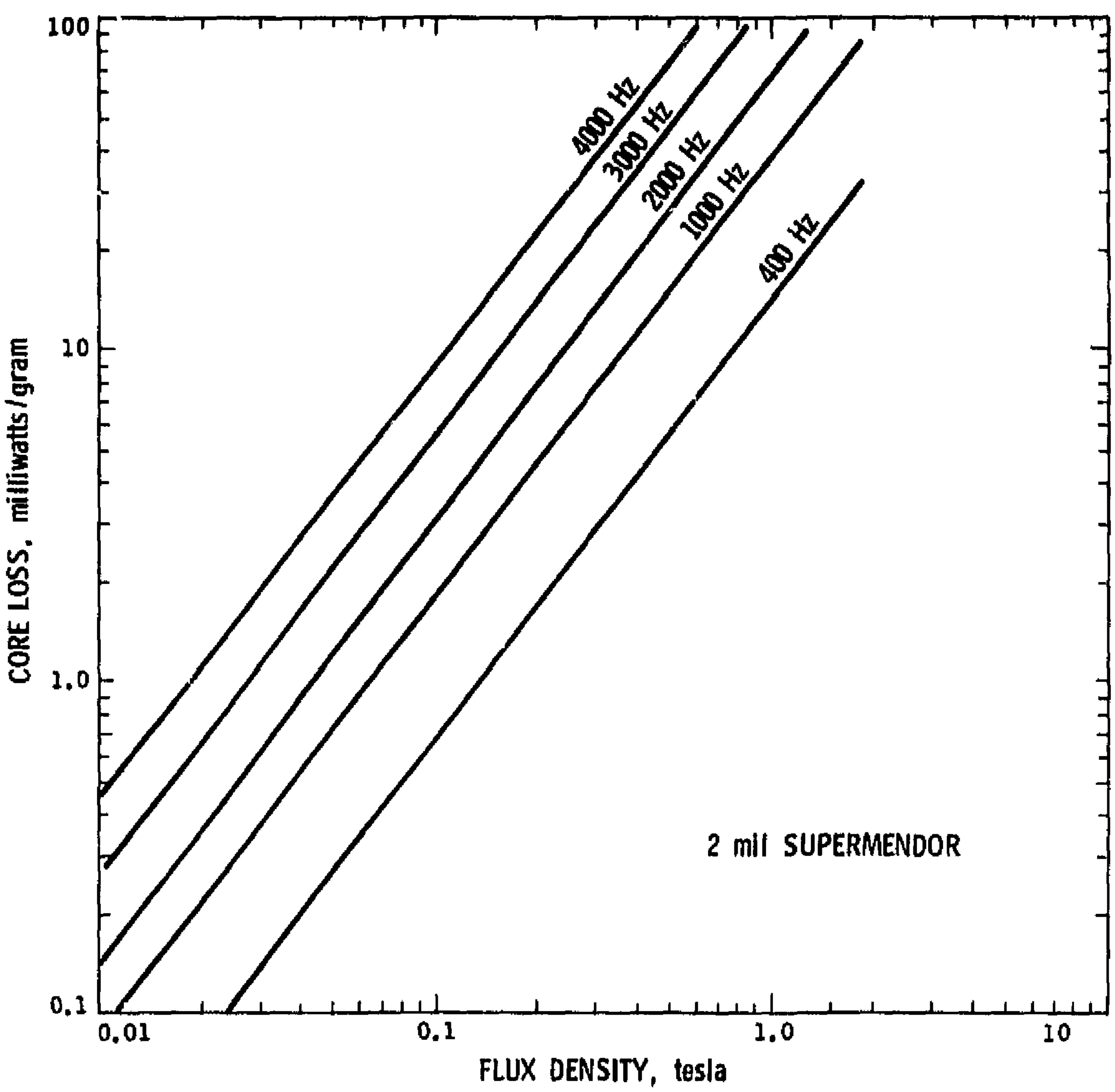

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Fig. 7-12. Design curves showing maximum core loss for 2 mil supermendor

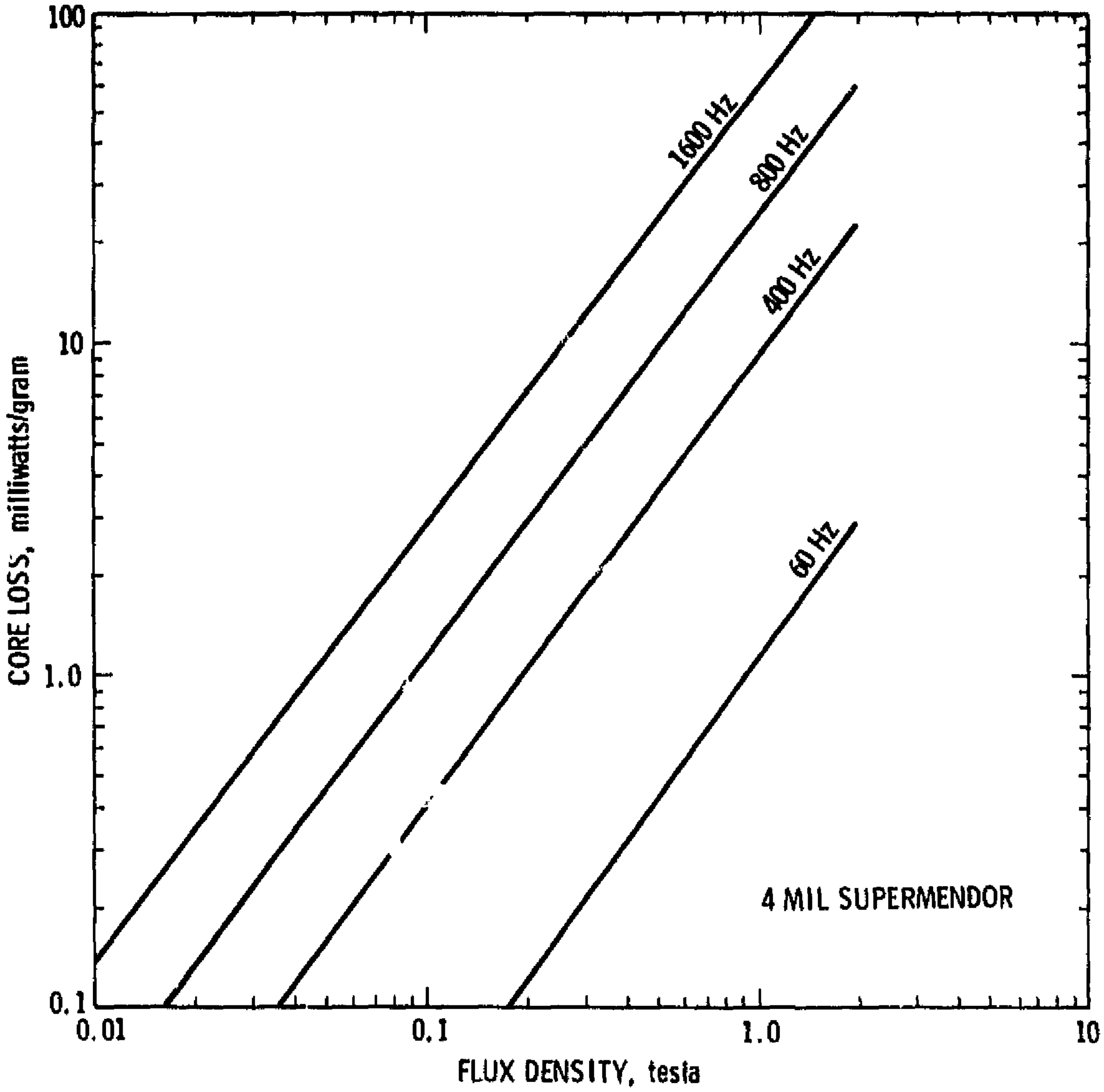

Fig. 7-13. Design curves showing maximum core loss for 4 mil supermendor

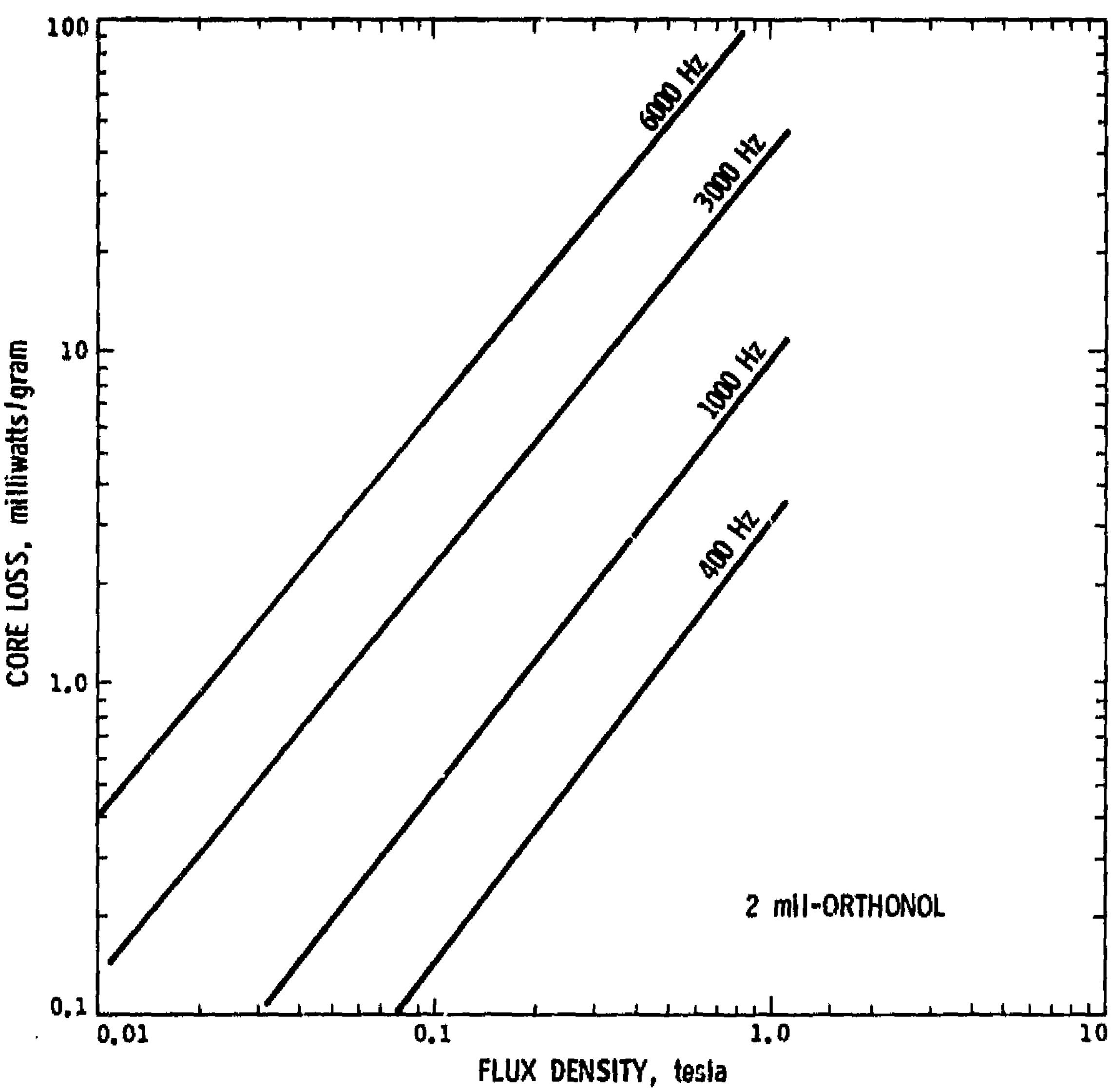

Fig. 7-14. Design curves showing maximum core loss for 2 mil 50% Ni, 50% Fe

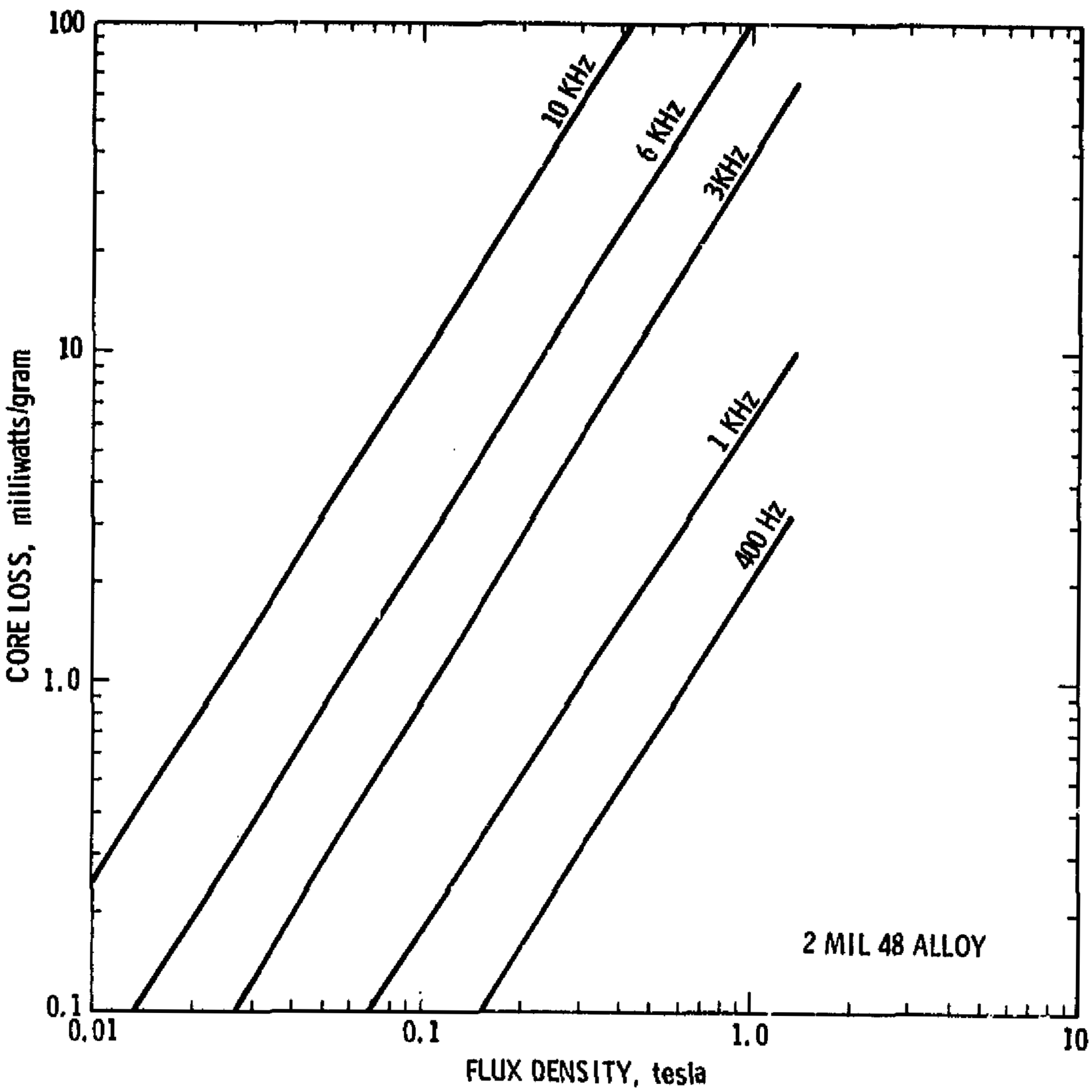

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Fig. 7-15. Design curves showing maximum core loss for 2 mil 48% Ni, 52% Fe

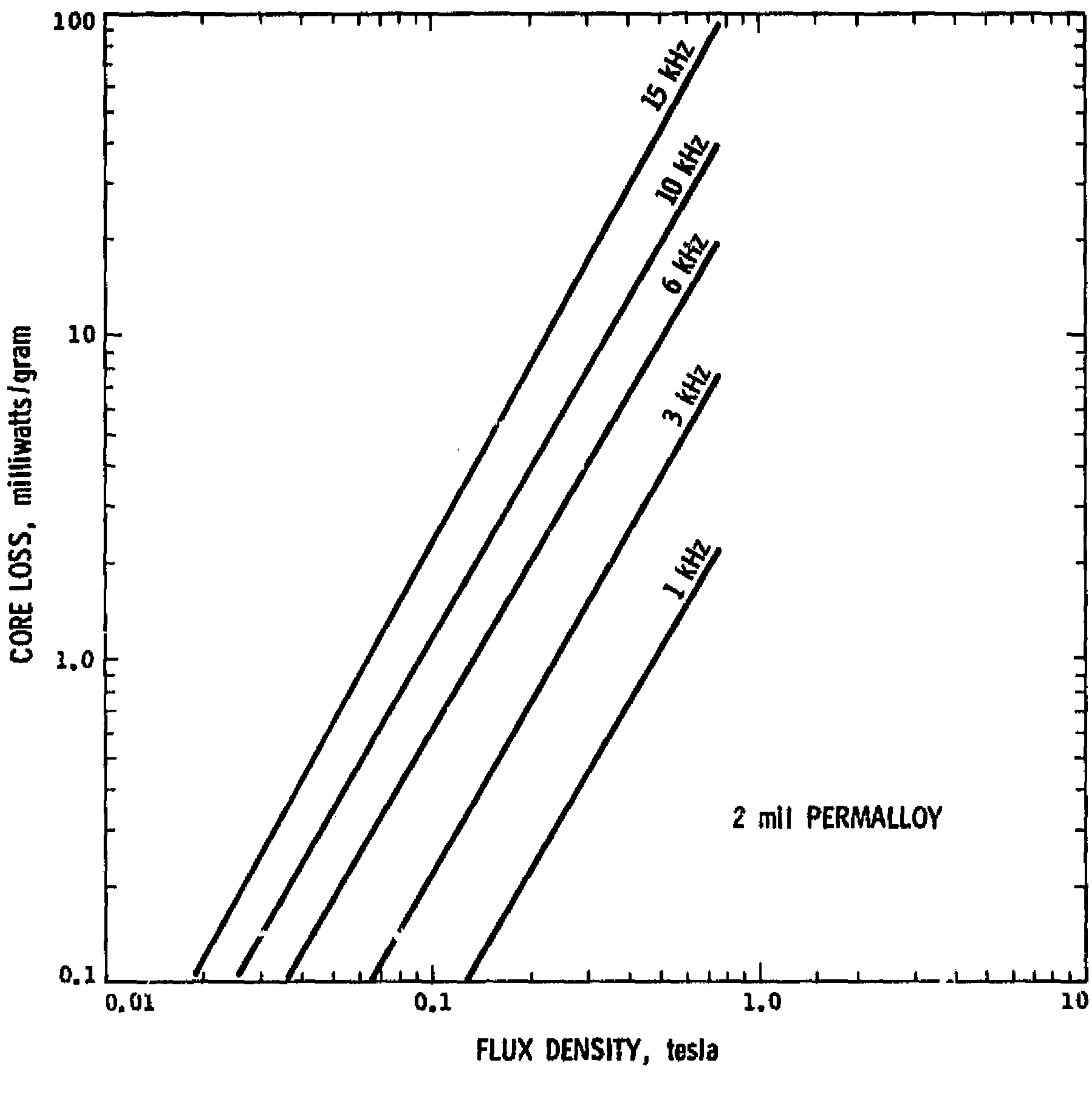

Fig. 7-16. Design curves showing maximum core loss for 2 mil 80% Ni, 20% Fe

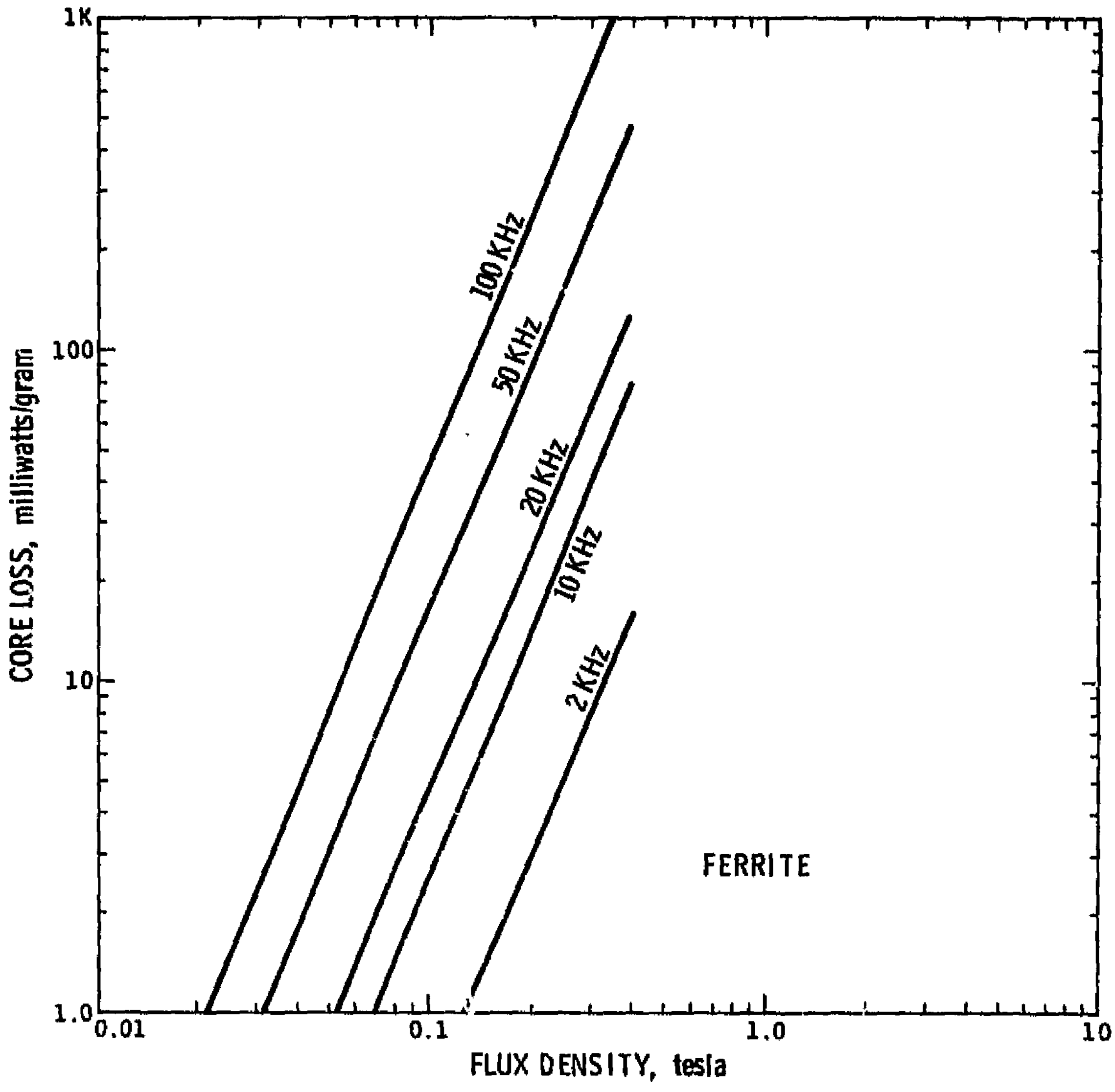

Fig. 7-17. Design curves showing maximum core loss for ferrite

Fortunately, there is such a large choice of core sizes available (Tables 2-2 through 2-7 list only a few of the different cores that are commercially available), that relative proportions of iron and copper can be varied over a wide range without changing the A_p area product.*

G. SKIN EFFECT

It is now common practice to operate dc-to-dc converters at frequencies up to 50 kHz. At the higher frequencies, skin effect alters the predicted efficiency since the current carried by a conductor is distributed uniformly across the conductor cross-section only at dc and at low frequencies. The concentration of current near the wire surface at higher frequencies is termed the skin effect. This is the result of magnetic flux lines which circle only part of the conductor. Those portions of the cross section which are circled by the largest number of flux lines exhibit greater reactance.

Skin effect accounts for the fact that the effective alternating current resistance to direct current ratio is greater than unity. The magnitudes of these effects at high frequency on conductivity, magnetic permeability and inductance are sufficient to require further evaluation of conductor size during design. The depth of the skin effect is expressed by:

$$\text{depth (cm)} = (6.61/f^{1/2})\ K \qquad (7\text{-}53)$$

in which K is a constant according to the relationship:

$$K = \left[(1/\mu r)\ \rho/\rho c\right]^{1/2} \qquad (7\text{-}54)$$

*However, at frequencies above about 20 kHz, eddy current losses are so much greater than hysteresis losses that it is necessary to use very thin (1 and 2 mil) strip cores.

in which:

μr = relative permeability of conductor material (μr = 1 for copper and other nonmagnetic materials)

ρ = resistivity of conductor material at any temperature

c = resistivity of copper at 20°C = 1.724 microhm-centimeter

K = unity for copper

Figures 7-18 and 7-19 below show respectively, skin depth as a function of frequency according to equation 7-53 above, and as related to the AWG radius, or as R_{ac}/R_{dc} = 1 versus frequency.*

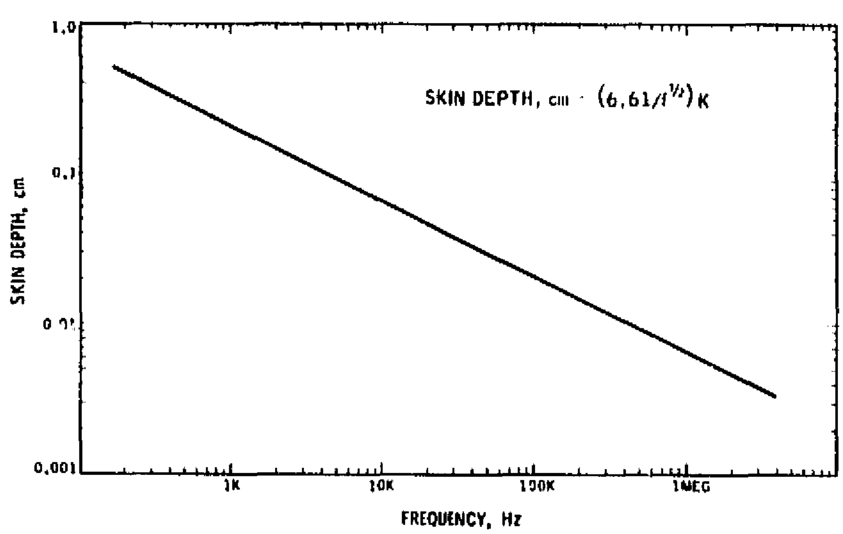

Fig. 7-18. Skin depth versus frequency

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*The data presented is for sine wave excitation. The author could not find any data for square wave excitation.

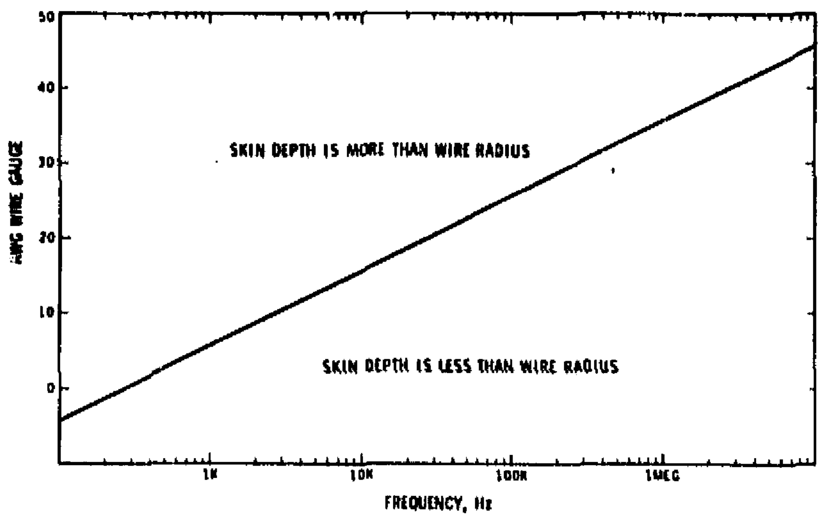

Fig. 7-19. Skin depth equal to AWG radius versus frequency

Figure 7-20 shows how the RMS values change with different waveshaps.

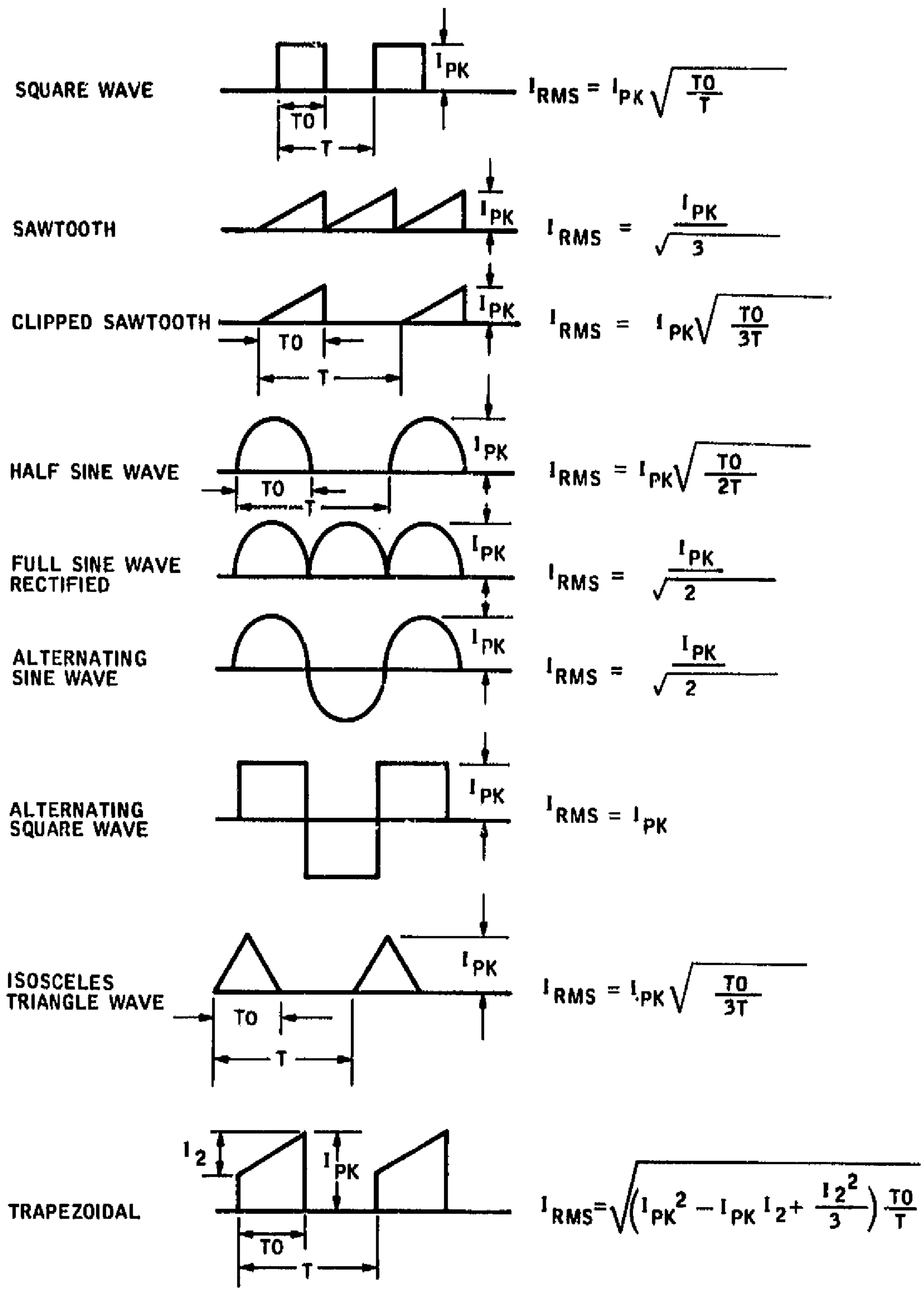

Fig. 7-20. Common waveshapes, RMS values

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REFERENCE

1. Technical Data on Arnold Tape - Wound Cover, TC-101B, Page 39, Arnold Engineer, Marengo, Ill.

APPENDIX 7.A
TRANSFORMERS DESIGNED FOR A GIVEN REGULATION

Although most transformers are designed for a given temperature rise, they can also be designed for a given regulation. The regulation and power-handling ability of a core is related to two constants:

$$VA = K_g K_e \alpha \tag{7.A-1}$$

$$\alpha = \text{Regulation (\%)}$$

The constant K_g is determined by the core geometry:

$$K_g = f(A_c, W_a, MLT) \tag{7.A-2}$$

The constant K_e is determined by the magnetic and electric operating conditions:

$$K_e' = f(f, B_m) \tag{7.A-3}$$

The derivation of the specific functions for K_g and K_e is as follows: first assume two-winding transformers with equal primary and secondary regulation, schematically shown in Figure 7.A-1. The primary winding has a resistance R_p ohms, and the secondary winding has a resistance R_s ohms:

$$\alpha = \frac{\Delta E_p}{E_p}(100) + \frac{\Delta E_s}{E_s}(100) \tag{7.A-4}$$

$$\Delta E_p = R_p I_p \tag{7.A-5}$$

$$\Delta E_s = R_s I_s \tag{7.A-6}$$

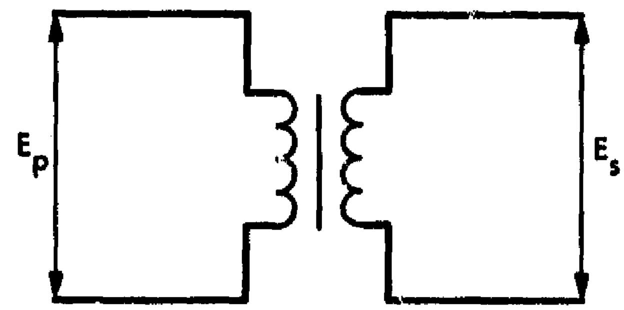

Fig. 7.A-1. Isolation transformer

$$\alpha = 2 \frac{R_p I_p}{E_p} (100) \tag{7.A-7}$$

Multiply the numerator and denominator by E_p:

$$\alpha = 200 \frac{R_p I_p}{E_p} \left(\frac{E_p}{E_p} \right) \tag{7.A-8}$$

$$\alpha = 200 \frac{R_p \, VA}{E_p^2} \tag{7.A-9}$$

From the resistivity formula, it is easily shown that

$$R_p = \frac{MLT \, N_p^2}{W_a K_p} \rho \tag{7.A-10}$$

$\rho = 1.724 \times 10^{-6}$ ohms · cm

K_p = window utilization factor (primary)

Faraday's law expressed in metric units is

$$E_p = KfNA_c B_m \times 10^{-4} \quad (7.A-11)$$

where

K = 4.0 square wave

K = 4.44 sine wave

Substituting equation 7.4-10 and 7.A-11 for R_p and E_p in equation 7.A-12,

$$VA = \frac{E_p^2}{200 R_p} \times \alpha \quad (7.A-12)$$

$$VA = \frac{\left(KfN_pA_cB_m \times 10^{-4}\right)\left(KfN_pA_cB_m \times 10^{-4}\right)}{200 \times \frac{(MLT)N_p^2 \rho}{W_a K_p}} \times \alpha \quad (7.A-13)$$

$$VA = \frac{K^2f^2A_c^2B_m^2 W_a K_p \rho \times 10^{-10}}{MLT} \times \alpha \quad (7.A-14)$$

Inserting 1.724×10^{-6} for ρ

$$VA = \frac{0.29 K^2f^2A_c^2B_m^2 W_a K_p \times 10^{-4}}{MLT} \times \alpha \quad (7.A-15)$$

Let

$$K_e = 0.29 K^2f^2B_m^2 \times 10^{-4} \quad (7.A-16)$$

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and

$$K_g = \frac{W_a K_p A_c^2}{MLT} \qquad [cm^5] \quad (7.A-17)$$

The total transformer window utilization factor is then

$$K_p + K_s = K_u \quad (7.A-18)$$

and equations 7.A-15 and 7.A-16 change to

$$K_e = 0.145 K^2 f^2 B_m^2 \times 10^{-4} \quad (7.A-19)$$

and

$$K_g = \frac{W_a K_u A_c^2}{MLT} \qquad [cm^5] \quad (7.A-20)$$

Coefficient K_g values for C cores, lamination, pot cores, powder cores, and tape-wound cores are shown in Tables 7.B-1 through 7.B-5.

Regulation of a transformer is related to the copper loss as shown in equation 7.A-21:

$$\alpha = \frac{P_{cu}}{P_o} \times 100 \qquad [\%] \quad (7.A-21)$$

The copper loss in a transformer is related to the RMS current (see Chapter 3, Power Transformer Design; also see Fig. 7-20).

Many transformers such as those used in DC-AC and DC-AC power supplys and for full wave rectifiers do not have 100% duty cycles in all windings. Proper selection of wire size based on duty cycle is, of course, necessary The following multipliers will convert these types to a VA rating based on 100% duty cycle in all windings.

PRIMARY DUTY CYCLE	SEC. DUTY CYCLE	MULTIPLY REQUIRED VA BY
100%	50%	1.41
50%	100%	1.41
50%	50%	1.82

APPENDIX 7.B
INDUCTORS DESIGNED FOR A GIVEN REGULATION

Inductors, like transformers, are designed for a given temperature rise. They can also be designed for a given regulation. The regulation and energy-handling ability of a core is related to two constants:

$$(\text{Energy})^2 = K_g K_e \alpha$$

$$\alpha = \text{Regulation } (\%) \tag{7.B-1}$$

The constant K_g is determined by the core geometry:

$$K_g = f(A_c, W_a, \text{MLT}) \tag{7.B-2}$$

The constant K_e is determined by the magnetic and electric operating conditions:

$$K_e = f(P_o, B_m) \tag{7.B-3}$$

The derivation of the specific functions for K_g and K_e is as follows for the circuit shown in Fig. 7.B-1:

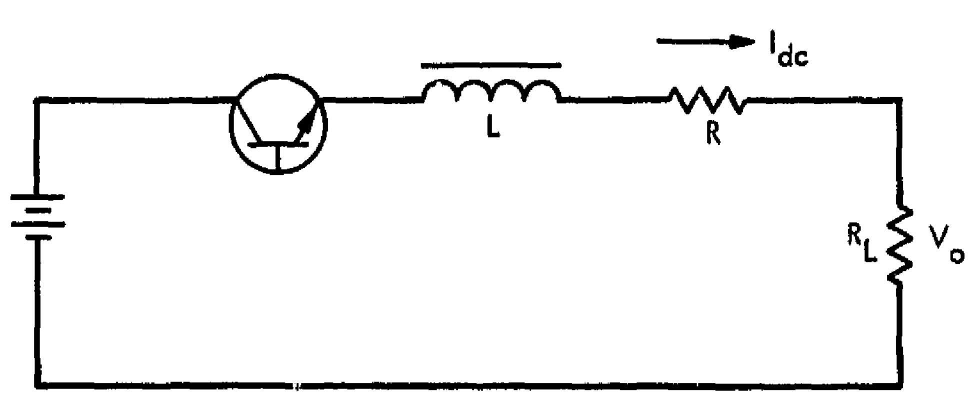

Fig. 7.B-1. Output inductor

$$P_o = I_{dc} V_o \qquad \text{[watts]} \quad (7.B\text{-}4)$$

$$\alpha = \frac{I_{dc} R}{V_o} 100 \qquad [\%] \quad (7.B\text{-}5)$$

Inductance is equal to

$$L = \frac{0.4\pi N^2 A_c \times 10^{-8}}{l_g} \qquad \text{[henry]} \quad (7.B\text{-}6)$$

Flux density is equal to

$$B_{dc} = \frac{0.4\pi N I_{dc} \times 10^{-4}}{l_g} \qquad \text{[tesla]} \quad (7.B\text{-}7)$$

Combining the two equations,

$$\frac{L}{B_{dc}} = \frac{N A_c \times 10^{-4}}{I_{dc}} \qquad (7.B\text{-}8)$$

Solving for N,

$$N = \frac{L I_{dc} \times 10^{4}}{B_{dc} A_c} \qquad (7.B\text{-}9)$$

Since the resistance equation is

$$R = \frac{\rho N^2 \, MLT}{K_u W_a} \qquad [\Omega] \quad (7.B\text{-}10)$$

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and the regulation equation is

$$\alpha = \frac{I_{dc}\ R}{V_o} \times 10^2 \qquad [\%] \quad (7.B\text{-}11)$$

Inserting the resistance equation (7. B-11) gives

$$\alpha = \frac{I_{dc}}{V_o} \times \frac{\rho N^2\ MLT}{K_u\ W_a} \times 10^2 \qquad (7.B\text{-}12)$$

$$N^2 = \left(\frac{L\ I_{dc}}{B_{dc}\ A_c}\right)^2 \times 10^8 \qquad (7.B\text{-}13)$$

$$\alpha = \frac{I_{dc}\ MLT\ \rho}{V_o\ I_u\ W_a} \times \left(\frac{L\ I_{dc}}{B_{dc}\ A_c}\right)^2 \times 10^{10} \qquad 7.B\text{-}14)$$

$$\alpha = \frac{I_{dc}\ MLT\ \rho\ (L\ I_{dc})^2}{V_o\ K_u\ W_a\ B_{dc}^2\ A_c^2} \times 10^{10} \qquad (7.B\text{-}15)$$

$$\text{Energy} = \frac{L\ I_{dc}^2}{2} \qquad [\text{watts seconds}] \quad (7.B\text{-}16)$$

Multiplying the equation by I_{dc}/I_{dc} and combining,

$$\alpha = \frac{\left(L\ I_{dc}^2\right)^2 \rho\ MLT \times 10^{10}}{V_o\ I_{dc}\ K_u\ W_a\ A_c^2\ B_{dc}^2} \qquad (7.B\text{-}17)$$

which reduces to

$$\alpha = \frac{(2\ \text{Energy})^2}{P_o\ B_{dc}^2} \times \frac{\rho\ \text{MLT}}{K_u\ W_a\ A_c^2} \times 10^{10} \qquad (7.\text{B-18})$$

$$\rho = 1.724 \times 10^{-6}\ \text{ohms} \cdot \text{cm}$$

$$\alpha = \frac{6.89\ (\text{Energy})^2}{P_o\ B_{dc}^2} \times \frac{\text{MLT}}{K_u\ W_a\ A_c^2} \times 10^4 \qquad (7.\text{B-19})$$

Solving for nergy,

$$(\text{Energy})^2 = 0.145\ P_o\ B_{dc}^2 \times \frac{K_u\ W_a\ A_c^2}{\text{MLT}} \times 10^{-4}\ \alpha \qquad (7.\text{B-20})$$

$$K_g = \frac{K_u\ W_a\ A_c^2}{\text{MLT}} \qquad \left[\text{cm}^5\right] \qquad (7.\text{B-21})$$

Coefficient K_g values for C cores, lamination, pot cores, powder cores, and tape-wound cores are shown in Tables 7.B.1 through 7.B.5.

$$K_e = 0.145\ P_o\ B_{dc}^2 \times 10^{-4} \qquad (7.\text{B-22})$$

$$\alpha = \frac{P_{cu}}{P_o} \times 100 \qquad [\%] \qquad (7.\text{B-23})$$

The regulation of an inductor is related to the copper loss, as shown in equation 7.B-24:

$$\alpha = \frac{P_{cu}}{P_o} \times 100 \qquad [\%] \qquad (7.\text{B-24})$$

The copper loss in an inductor is related to the RMS current. The RMS current in a down regulator, as shown in Figure 7.B-1, is always equal to or less than I_o:

$$I_{RMS} \leq I_o \qquad (7.B\text{-}25)$$

Table 7.B-1. Coefficient K_g for C cores[a]

Core	$10^{-3} K_g$	W_a, cm^2	A_c, cm^2	MLT, cm	G, cm	D, cm
AL-2	6.27	1.006	0.264	4.47	1.587	0.635
AL-3	14.4	1.006	0.406	5.10	1.587	0.952
AL-5	30.5	1.423	0.539	5.42	2.22	0.952
AL-6	47.8	1.413	0.716	6.06	2.22	1.27
AL-124	63.1	2.02	0.716	6.56	2.54	1.27
AL-8	106	2.87	0.806	7.06	3.015	0.952
AL-9	173	2.87	1.077	7.69	3.015	1.27
AL-10	248	2.87	1.342	8.33	3.015	1.587
AL-12	256	3.63	1.260	9.00	2.857	1.27
AL-135	273	4.083	1.260	9.50	2.857	1.27
AL-78	399	4.53	1.340	8.15	5.715	1.91
AL-18	530	6.30	1.257	7.51	3.927	1.27
AL-15	648	5.037	1.80	10.08	3.967	1.587
AL-16	869	5.037	2.15	10.72	3.967	1.905
AL-17	1380	5.037	2.87	11.99	3.967	2.54
AL-19	1600	6.30	2.87	12.98	3.967	2.54
AL-20	2370	6.30	3.58	13.62	3.967	2.54
AL-22	2940	7.804	3.58	13.62	4.92	2.54
AL-23	4210	7.804	4.48	14.98	4.92	3.175
AL-24	3910	11.16	3.58	14.62	5.875	2.54

[a]Where $K_u = 0.4$.

Table 7.B-2. Coefficient K_g for laminations[a]

Core	10^{-3} K_g	W_a, cm^2	A_c, cm^2	MLT, cm	G, cm	D, cm
EE 3031	0.103	0.176	0.0502	1.72	0.714	0.239
EE 2829	0.356	0.252	0.0907	2.33	0.792	0.318
EI 187	2.75	0.530	0.204	3.20	1.113	0.478
EE 2425	8.37	0.807	0.363	5.08	1.27	0.635
EE 2627	51.1	1.11	0.816	5.79	1.748	0.953
EI 375	63.8	1.51	0.816	6.30	1.905	0.953
EI 50	144	1.21	1.45	7.09	1.91	1.27
EI 21	181	1.63	1.45	7.57	2.06	1.27
EI 625	441	1.89	2.27	8.84	2.38	1.59
EI 75	1100	2.72	3.27	10.6	2.86	1.91
EI 87	2390	3.71	4.45	12.3	3.33	2.22
EI 100	4500	4.83	5.81	14.5	3.81	2.54
EI 112	8240	6.12	7.34	16.0	4.28	2.86
EI 125	14100	7.57	9.07	17.7	4.76	3.18
EI 138	25400	9.20	11.6	19.5	5.24	3.49
EI 150	35300	10.9	13.1	21.2	5.72	3.81
EI 175	75900	14.8	17.8	24.7	6.67	4.45
EI 36	74900	21.2	15.3	26.5	6.67	4.13
EI 19	135000	33.8	17.8	31.7	7.62	4.45

[a]Where K_u = 0.4.

Table 7.B-3. Coefficient K_g for pot cores[a]

Core	10^{-3} K_g	W_a, cm^2	A_c, cm^2	MLT, cm
9 × 5	0.109	0.065	0.10	1.85
11 × 7	0.343	0.095	0.16	2.2
14 × 8	1.09	0.157	0.25	2.8
18 × 11	4.28	0.266	0.43	3.56
22 × 13	10.9	0.390	0.63	4.4
26 × 16	27.9	0.530	0.94	5.2
30 × 19	71.6	0.747	1.[illegible]	6.0
36 × 22	171	1.00	2.[illegible]	7.3
47 × 28	584	1.80	3.12	9.3
59 × 36	1683	2.77	4.85	12.0

[a]Where K_u = 0.31.

Table 7.B-4. Coefficient K_g for powder core[a]

Core	10^{-3} K_g	W_a, cm^2	A_c, cm^2	MLT, cm
55051	0.901	0.381	0.113	2.16
55121	4.00	0.713	0.196	2.74
55848	8.26	1.14	0.232	2.97
55059	17.4	1.407	0.327	3.45
55894	55.3	1.561	0.639	4.61
55586	77.7	4.00	0.458	4.32
55071	108	2.93	0.666	4.80
55076	134	3.64	0.670	4.88
55083	316	4.27	1.060	6.07
55090	639	6.11	1.32	6.66
55439	852	4.27	1.95	7.62
55716	712	7.52	1.24	6.50
55110	1123	9.48	1.44	7.00

[a]Where K_u = 0.4.

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Table 7.B-5. Coefficient K_g for tape-wound toroids[a]

Core	10^3 Kg	W_a, cm^2	A_c, cm^2	MLT, cm
52402	0.0472	0.502	0.022	2.06
52153	0.254	0.502	0.053	2.22
52107	0.0860	0.982	0.022	2.21
52403	0.107	1.28	0.022	2.30
52057	0.456	1.56	0.043	2.53
52000	1.07	0.982	0.086	2.70
52063	1.62	1.56	0.086	2.85
52002	1.81	1.76	0.086	2.88
52007	10.6	1.56	0.257	3.87
52167	17.4	1.56	0.343	[illegible].23
52094	20.8	1.56	0.386	4.47
52004	12.7	4.38	0.171	4.02
52032	44.3	4.38	0.343	4.65
52026	87.7	4.38	0.514	5.28
52038	138	4.38	0.686	5.97
52035	203	6.816	0.686	6.33
52055	276	9.93	0.686	6.76
52012	587	6.94	1.371	8.88
52017	459	18.3	0.686	7.51
52031	668	29.2	0.686	8.23
52103	1570	18.3	1.371	8.77
52128	2220	28.0	1.371	9.49
52022	4870	18.3	2.742	11.30
52042	6790	27.1	2.742	12.0
52100	18600	27.1	5.142	15.4
52112	68100	73.6	6.855	20.3
52426	159000	73.6	10.968	22.2

[a]Where K_u = 0.4.

APPENDIX 7.C
TRANSFORMER AREA PRODUCT AND GEOMETRY

The geometry K_g of a transformer, which can be related to the area product A_p, is derived in Chapter 7 and is shown here in equation 7.C-1. Derivation of the relationship is according to the following: Geometry K_g varies in accordance with the fifth power of any linear dimension ℓ (designated ℓ^5 below), whereas area product A_p varies as the fourth power:

$$K_g = \frac{W_a A_c^2 K_u}{MLT} \quad (7.C\text{-}1)$$

$$K_g = K_{10} \ell^5 \quad (7.C\text{-}2)$$

$$A_p = K_2 \ell^4 \quad (7.C\text{-}3)$$

$$\ell = \left(\frac{K_g}{K_{10}}\right)^{0.20} \quad (7.C\text{-}4)$$

$$\ell^4 = \left[\left(\frac{K_g}{K_{10}}\right)^{0.20}\right]^4 = \left(\frac{K_g}{K_{10}}\right)^{0.8} \quad (7.C\text{-}5)$$

$$A_p = K_2 \left(\frac{K_g}{K_{10}}\right)^{0.8} \quad (7.C\text{-}6)$$

$$K_p = \frac{K_2}{K_{10}^{0.8}} \quad (7.C\text{-}7)$$

$$A_p = K_p K_g^{0.8} \quad (7.C\text{-}8)$$

The area product/geometry relationship is

$$A_p = K_p K_g^{0.8}$$

in which K_p is a constant related to core configuration, shown in Table 7.C-1, which has been derived by averaging the values in Tables 2-2 through 2-7 (see Chapter 2) and Tables 7.B-1 through 7.B-5.

The relationship between area product A_p and core geometry is given in Figures 7.C-1 through 7.C-5. It was obtained from the data shown in Tables 2-2 through 2-7 for area product A_p and Tables 7.B-1 through 7 B-5 for K_g.

Table 7.C-1. Constant K_p relationship

Core type	K_p
Pot cover	8.87
Powder cores	11.8
Lamination	8.3
C cores	12.5
Tape-wound cores	

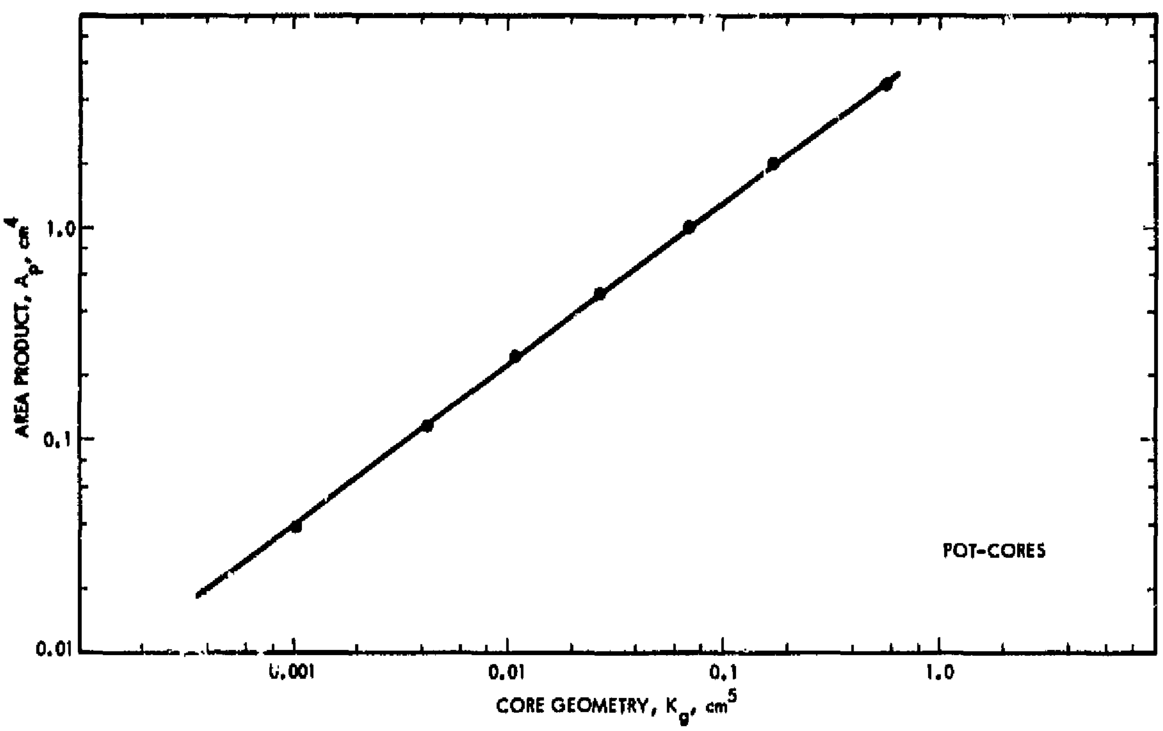

Fig. 7.C-1. Area product versus core geometry for pot cores

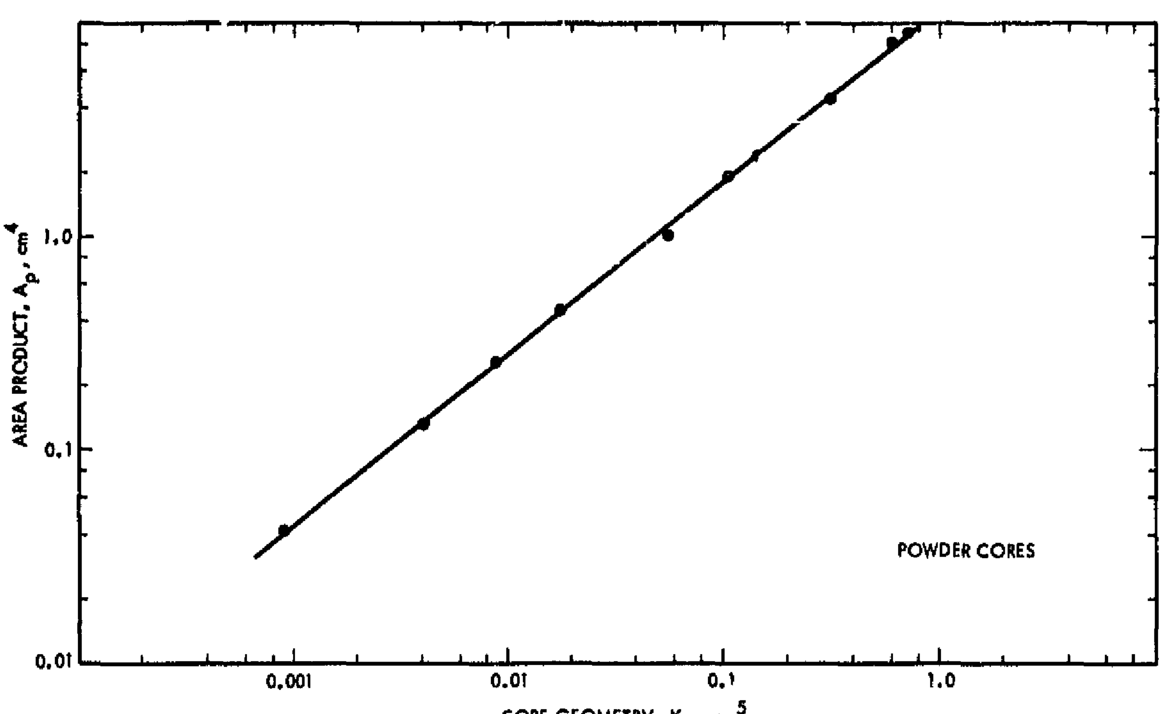

Fig. 7.C-2. Area product versus core geometry for powder cores

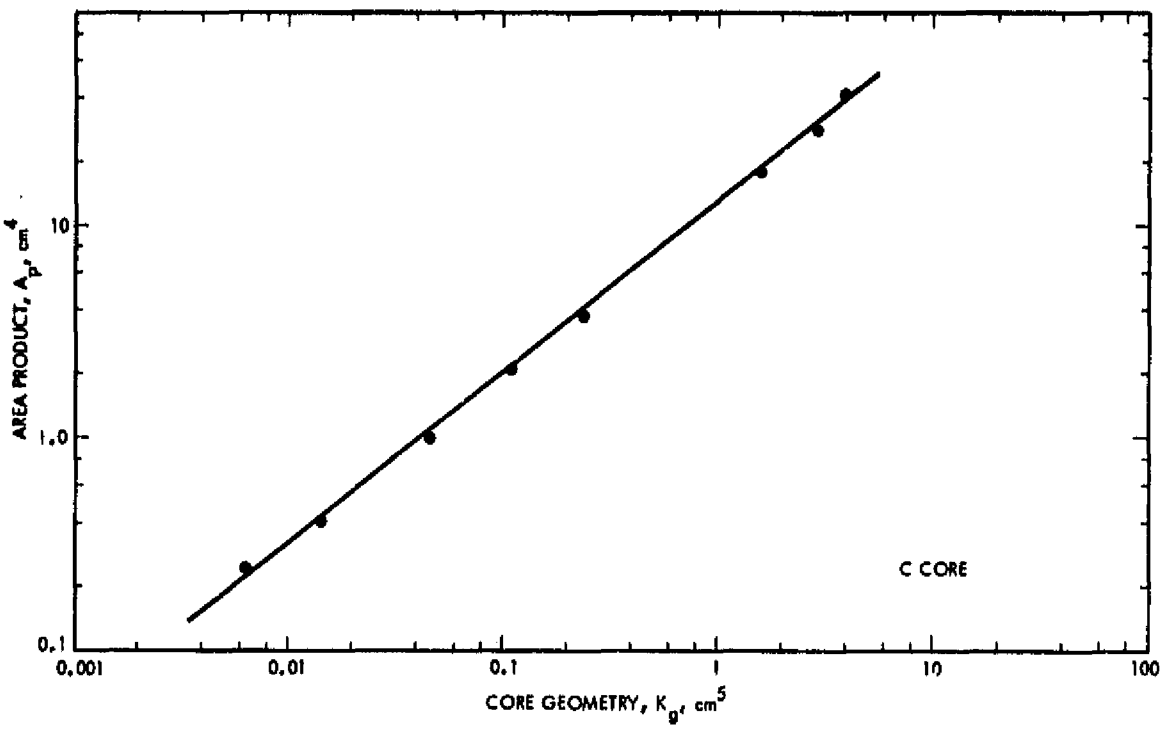

Fig. 7.C-3. Area product versus core geometry for C cores

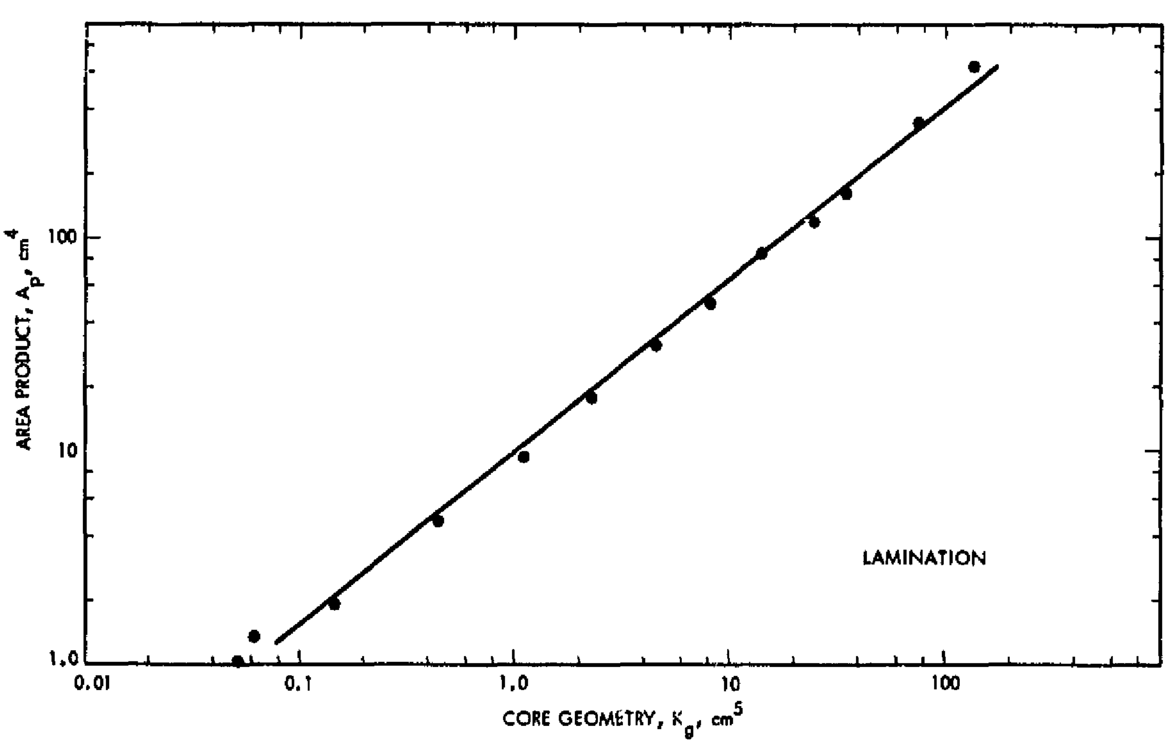

Fig. 7.C-4. Area product versus core geometry for laminations

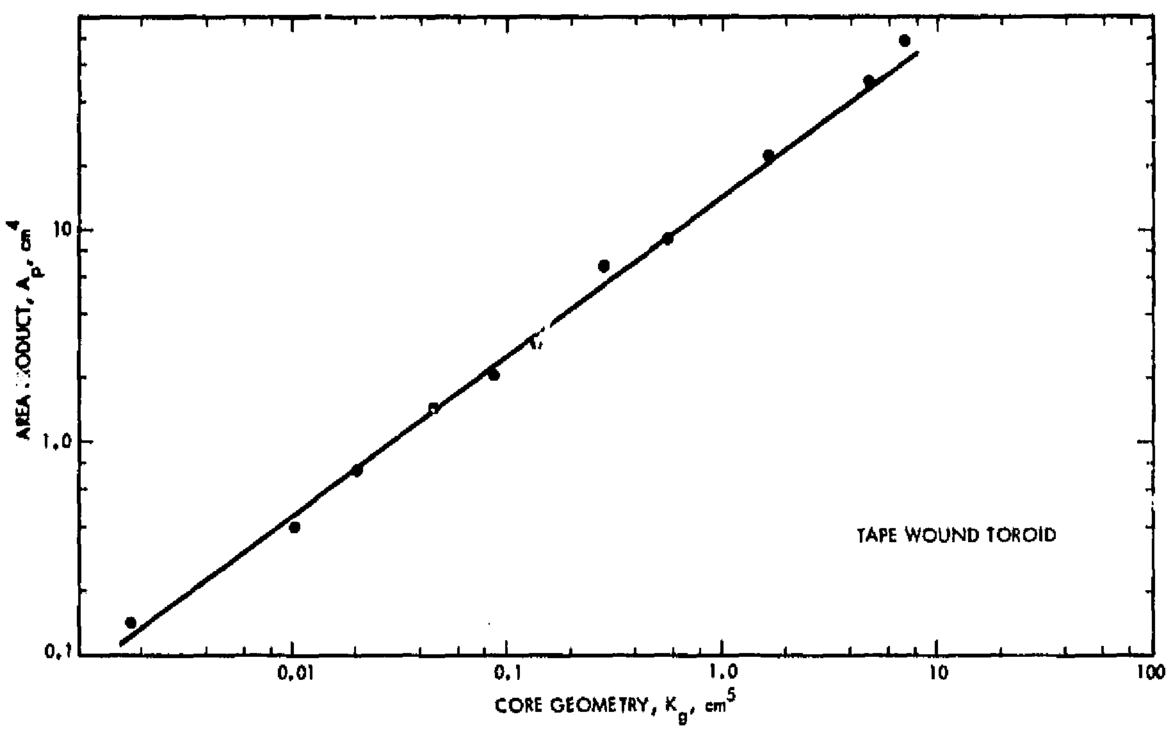

Fig. 7.C-5. Area product versus core geometry for tape-wound toroids

Chapter VIII

INDUCTOR DESIGN WITH NO DC FLUX

A. INTRODUCTION

The design of an ac inductor is quite similar to designing a transformer. If there is no dc flux in the core the design calculations are straightforward.

The apparent power P_t of an inductor is the VA of the inductor; that is, the excitation voltage and the current through the inductor:

$$P_t = VA \tag{8-1}$$

B. RELATIONSHIP OF A_p TO INDUCTOR VOLT-AMPERE CAPABILITY

According to the newly developed approach, the volt-ampere capability of a core is related to its area product A_p by an equation which may be stated as follows:

$$A_p = \left(\frac{VA \times 10^4}{4.44\ B_m f\ K_U\ K_j}\right)^{1.14} \tag{8-2}$$

K_j = current density coefficient (see Chapter 2)

K_U = window utilization factor (see Chapter 6)

f = frequency, Hz

B_m = flux density, tesla

From the above it can be seen that factors such as flux density, window utilization factor K_u (which defines the maximum space which may be occupied by the copper in the window), and the constant K_j (which is related to temperature rise), all have an influence on the inductor area product. The constant K_j is a new parameter that gives the designer control of the copper loss. Derivation is set forth in detail in Chapter 2.

B. FUNDAMENTAL CONSIDERATIONS

The design of a linear inductor depends upon four related factors:

(1) Desired inductance

(2) Applied Voltage

(3) Frequency

(4) Operating flux density

With these requirements established, the designer must determine the maximum values for B_{ac} which will not produce magnetic saturation, and make tradeoffs which will yield the highest inductance for a given volume. The core material selected determines the maximum flux density that can be tolerated for a given design. Magnetic saturation values for different core materials are given in Table 4-1.

The number of turns is calculated from the Faraday law, which states:

$$N = \frac{E \times 10^4}{4.44\ B_m\ f\ A_c} \qquad (8\text{-}3)$$

The inductance of an iron-core inductor having an air gap may be expressed as

$$L = \frac{0.4\pi N^2\ A_c \times 10^{-8}}{l_g + \frac{l_m}{\mu_r}} \qquad \text{[henry]} \quad (8\text{-}4)$$

Inductance is dependent on the effective length of the magnetic path which is the sum of the air gap length (l_g) and the ratio of the core mean length to relative permeability (l_m/μ_r).

When the core air gap (l_g) is large compared to relative permeability (l_m/μ_r), because of the high relative permeability (μ_r), variations in μ_r do not substantially effect the total effective magnetic path length or the inductance.

The inductance equation then reduces to:

$$L = \frac{0.4\pi N^2 A_c \times 10^{-8}}{l_g} \qquad \text{henry (8-5)}$$

Final determination of the air gap requires consideration of the effect of fringing flux, which is a function of gap dimension, the shape of the pole faces and the shape, size and location of the winding. Its net effect is to make the effective air gap shorter than its physical dimension.

Fringing flux decreases the total reluctance of the magnetic path and therefore increases the inductance by a factor F to a value greater than that calculated from equation (8-5). Fringing flux is a larger percentage of the total for larger gaps. The fringing flux factor is:

$$F = \left(1 + \frac{l_g}{\sqrt{A_c}} \log_e \frac{2G}{l_g}\right) \qquad (8\text{-}6)$$

where G is a dimension defined in Chapter 2. (Equation 8-6 is also valid for laminations; this equation is plotted in Figure 4-3).

Inductance L computed in equation (8-5) does not include the effect of fringing flux. The value of inductance L' corrected for fringing flux is:

$$L' = \frac{0.4\pi N^2 A_c F \times 10^{-8}}{l_g} \qquad \text{[henry] (8-7)}$$

The losses in an ac inductor are made up of three components:

(1) Copper loss, P_{cu}

(2) Iron loss, P_{fe}

(3) Gap loss, P_g

The copper loss and iron loss have been previously discussed. Gap loss* is independent of core strip thickness and permeability. Maximum efficiency

*Reference

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is reached in an inductor, as in a transformer, when the copper loss P_{cu} and the iron loss P_{fc} are equal but only when the core gap is zero. The loss does not occur in the air gap itself, but is caused by magnetic flux fringing around the gap and re-entering the core in a direction of high loss. As the air gap increases, the flux across it fringes more and more, and some of the fringing flux strikes the core perpendicular to the laminations and sets up eddy currents which cause additional loss. Distribution of fringing flux is also affected by other aspects of core geometry, the proximity of coil turns to the core, and whether there are turns on both legs. Accurate prediction of gap loss depends on the amount of fringing flux. For laminated cores it can be estimated from

$$P_g = K_i 2 D l_g f B_m^2 \quad \text{[watts]} \quad (8\text{-}8)$$

K_i = 0.0388

D = lamination tongue width, cm

l_g = gap length, cm

f = frequency, Hz

B_m = flux density, tesla

The fringing flux is around the gap and re-entering the core in a direction of high loss as shown in Figure 8-1.

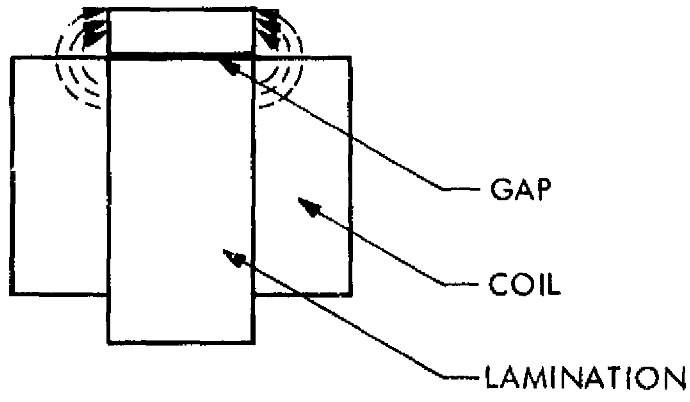

Fig. 8-1. Fringing flux around the gap of an inductor designed with lamination

D. DESIGN EXAMPLE

For a typical design example, assume:

(1) Constructed with laminations

(2) Applied voltage, 115 V

(3) Frequency, 60 Hz

(4) Alternating current, 0.5 amps

(5) $25^{o}C$ rise

The design procedure would then be as follows:

Step No. 1. Calculate the apparent power P_t from equation 8-1:

$$P_t = VA$$

$$P_t = (115)(0.5)$$

$$P_t = 57.5$$

Step No. 2. Calculate the area product A_p from equation 8-2:

$$A_p = \left(\frac{VA \times 10^4}{4.44\ B_m f K_u\ K_j}\right)^{1.14}$$

B_m = 1.2 tesla

K_u = 0.4 (see Chapter 6)

K_j = 366 (see Chapter 2)

$$A_p = \left(\frac{57.5 \times 10^4}{4.44\ (1.2)(60)(0.4)(366)}\right)^{1.14}$$

$$A_p = 17.4$$

<u>Step No. 3.</u> Select a size of lamination from Table 2-4 with a value A_p closest to the one calculated.

$$\text{E1-87 with an } A_p = 16.5$$

<u>Step No. 4.</u> Calculate the number of turns using Faraday's law, equation 8-3:

$$N = \frac{E \times 10^4}{4.44\ B_m f A_c}$$

The iron cross-section A_c is found ı Table 2-4:

$$A_c = 4.45 \qquad [cm^2]$$

$$N = \frac{115 \times 10^4}{(4.44)(1.2)(60)(4.45)}$$

$$N = 808 \qquad [turns]$$

<u>Step No. 5.</u> Calculate the impedance:

$$X_L = \frac{E}{I}$$

$$X_L = \frac{115}{0.5}$$

$$X_L = 230 \qquad [\Omega]$$

Step No. 6. Calculate the inductance:

$$L = \frac{X_L}{2\pi f}$$

$$L = \frac{230}{(6.28)(60)}$$

$$L = 0.610 \qquad [henry]$$

Step No. 7. Calculate the air gap from the inductance, equation 8-5:

$$l_g = \frac{0.4\pi N^2 A_c \times 10^{-8}}{L} \qquad [cm]$$

$$l_g = \frac{(1.26)(808)^2(4.45)(10^{-8})}{0.610}$$

$$l_g = 0.060 \qquad [cm]$$

Gap spacing is usually maintained by inserting Kraft paper. However this paper is only available in mil thicknesses. Since l_g has been determined in cm, it is necessary to convert as follows:

$$cn \times 393.7 = \text{mils (inch system)}$$

Substitucing values:

$$0.060 \times 393.7 = 23.6 \qquad [mils]$$

When designing inductors using lamination, it is common to place the gapping material along the mating surface between the E and I. When this method of gapping is used, only half of the material is required. In this case a 10 mil and a 2 mil thickness were used.

Step No. 8. Calculate the amount of fringing flux from equation 8-6; the value for G is found in Table 7-B2:

$$F = \left(1 + \frac{l_g}{\sqrt{A_c}} \log_e \frac{2G}{l_g}\right)$$

$$F = \left(1 + \frac{0.060}{\sqrt{4.45}} \log_e \frac{2(3.33)}{0.060}\right)$$

$$F = 1.13$$

After finding the fringing flux F, insert it into equation 8-7, rearrange and solve for the correct number of turns:

$$N = \sqrt{\frac{l_g L}{0.4\pi A_c F \times 10^{-8}}}$$

$$N = \sqrt{\frac{(0.060)(0.610)}{(1.26)(4.45)(1.13) \times 10^{-8}}}$$

$$N = 760$$

The design should be checked to verify that the reduction in turns does not cause saturation of the core.

Step No. 9. Calculate the current density using Table 2-1:

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$$J = K_j A_p^{-0.12}$$

$$J = (366)(16.5)^{-0.12}$$

$$J = 261 \qquad [\text{amps/cm}^2]$$

<u>Step No. 10.</u> Determine the bare wire size $A_{w(E)}$

$$A_{w(B)} = \frac{I}{J}$$

$$A_{w(B)} = \frac{0.5}{261}$$

$$A_{w(B)} = 0.00192 \qquad [cm^2]$$

<u>Step No. 11.</u> Select an AWG wire size from Table 6-1, column A.

$$\text{AWG No. 24} = 0.00205 \qquad [cm^2]$$

The rule is that when the calculated wire size does not fall close to those listed in the table, the next smaller size should be selected.

<u>Step No. 12.</u> Calculate the resistance of the winding using Table 6-1, column C, and Table 2-4 for the MLT:

$$R = MLT \times N \times (\text{column C}) \times \zeta \times 10^{-6}$$

$$R = (12.3)(760)(842.1)(1.098) \times 10^{-6}$$

$$R = 8.64 \qquad [\Omega]$$

<u>Step No. 13.</u> Calculate the power loss in the winding:

$$P_{cu} = I^2 R$$

$$P_{cu} = (0.5)^2 (8.64)$$

$$P_{cu} = 2.16 \qquad [\text{watts}]$$

From the core loss curves (Figure 7-10), 12 mil silicon at a flux density of 1.2 tesla has a core loss of approximately 1.0 milliwatts per gram. The lamination El-87 has a weight of 481 grams:

$$P_{fe} = (0.001)(481)$$

$$P_{fe} = 0.481 \quad \text{[watts]}$$

Step No. 14. Calculate the gap loss from equation 8-8; the value of D is found in Table 7-B-2:

$$P_g = K_i 2D\, l_g\, f\, B_m^2 \quad \text{[watts]}$$

$$P_g = (0.0388)(4.44)(0.060)(60)(1.2)^2$$

$$P_g = 0.894 \quad \text{[watts]}$$

Step No. 15. Calculate the combined losses, copper, iron, and gap:

$$P_\Sigma = P_{cu} + P_{fe} + P_g$$

$$P_\Sigma = 2.16 + 0.481 + 0.894$$

$$P_\Sigma = 3.53 \quad \text{[watts]}$$

In a test sample made to verify these example calculations, the measured inductance was found to be 0.592 henry with a current 0.515 ampere at 115 volt, 60 Hz, and the inductor had a coil resistance of 8.08 ohms.

REFERENCES

1. Ruben, L., and Stephens, D. Gap Loss in Current-Limiting Transformers. Electromechanical Design, April 1973, Pages 24-26.

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