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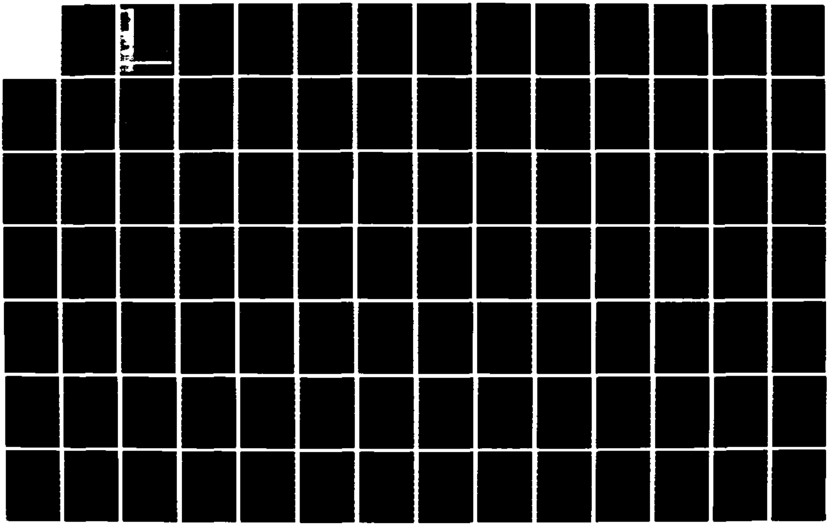

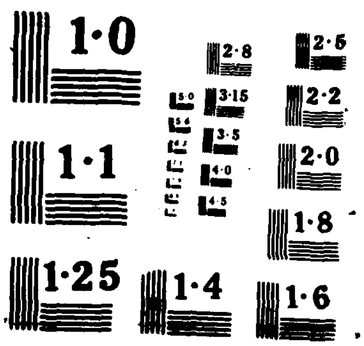
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MAGNETIC DESIGN GUIDELINES FOR ELECTRONIC POWER SUPPLIES

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NAVAL AVIONICS CENTER

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This report addresses the subject of magnetics design and fabrication for electronic power supplies. The two types of magnetics discussed are transformers and inductors. This report is intended as a guide for the design of power supply magnetics at the Naval Avionics Center.

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PREFACE

The purpose of this report is to provide guidelines for the design of magnetic components (transformers and inductors) for electronic power supplies. As a design guideline, this report does not present a comprehensive theoretical treatment of magnetics. Many textbooks are available for such a theoretical treatment. This report is an assimilation of theory, practical experience, and knowledge which is intended to be a useful design guideline for selecting materials, cores and winding configurations, and applying the design equations leading to the fabrication of reliable and producible magnetic components.

The author, Mr. Robert K. Davis, has developed his magnetics design expertise from over 25 years experience, beginning in the early 1960's up to his retirement from the Naval Avionics Center in December 1985.

Thanks are accorded to Mr. John Jentz and Mr. Chris K. Hagan, who helped in organizing and editing the text. And most importantly, the typists, Ms. Effie Jones and Mrs. Susan Daughtery deserve a special thanks for their considerable patience, excellent support and diligence in typing this report.

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1.0 INTRODUCTION

This report addresses the subject of magnetics design and fabrication for power supplies for military electronics equipment. The two types of magnetics discussed are the transformer and the inductor. Twenty-five years ago most of the magnetics designs were based on the common power line sine wave frequencies of 60 and 400 hertz. Because of the emphasis on switchmode power supplies, which directly rectify the power line frequency to get a DC voltage that is converted at a higher frequency (20 kilohertz and greater) to the desired DC voltage levels, most of today's magnetic designs are based on rectangular waves at frequencies greater than 20 kilohertz.

The magnetic design and fabrication techniques discussed are most applicable to a power range from a few watts up to two kilowatts and a frequency range from 50 hertz to 500 kilohertz.

As indicated in the Table of Contents, the magnetics information is presented in the following order and a brief description of each section is provided:

Section 2 - Magnetics Overview

Provides an overview of the basic magnetic design process.

Section 3 - Cores

Provides a discussion of applicable magnetic materials, core geometries, and the selection of core material and geometry based on the application.

Section 4 - Windings

Provides design considerations related to the different winding configurations, the selection of magnet wire and interlayer winding insulation.

Section 5 - Transformer Design

Addresses the design of two major classes of transformers:

a. Power line frequency (60 and 400 hertz) transformers using silicon iron magnetic core materials.

b. Switchmode power transformers using primarily ferrite core materials.

Section 6 - Inductor Design

Addresses the design of power line frequency (60 and 400 hertz) inductors using silicon iron magnetic core materials and switchmode power inductors using ferrite core materials.

Section 7 - Design Practices

Discusses a number of techniques in the construction of magnetics which, when properly applied, will result in producible magnetic designs.

Section 8 - Transformer and Inductor Testing

Discusses the various tests and identifies the appropriate test equipment to verify that the magnetic component meets the specification and design requirements.

2.0 MAGNETICS OVERVIEW

Although this report covers both power line frequency and higher switchmode frequency magnetic design, most power supply magnetic components, because of increasing use of switchmode power supplies in military applications, operate at frequencies greater than 20 kilohertz. Regardless of the operating frequency, the magnetic components, design equations, known parameters, and design approach are very similar.

Five of the magnetic components that may be found in a power supply are the input filter inductor, output filter inductor, power transformer, drive transformer, and auxiliary transformer (Figure 1). For a given supply it is possible that components will be designed for use at different frequencies, e.g., a 400 Hz input inductor and a 100 kHz switchmode transformer.

At the outset of a magnetics design some of the parameters required in the design equations will not be known. However, input voltage, usable maximum flux density, core cross-sectional area, and the frequency are usually available. The designer must determine the primary and secondary turns, wire sizes, and minimum core size. The design approach, which is outlined below, is the same regardless of the frequency.

1. Determine the transformer and/or inductor requirements.
2. Select the core material and geometry based on the application and frequency.
3. Estimate the core window area required for the windings.
4. Estimate the power which the magnetic component must process and select a trial core for initial design calculations.
5. Iterate the design process (2 through 4) until a compatible design is determined.
6. Finalize the magnetic specification for electrical, construction, and test details.

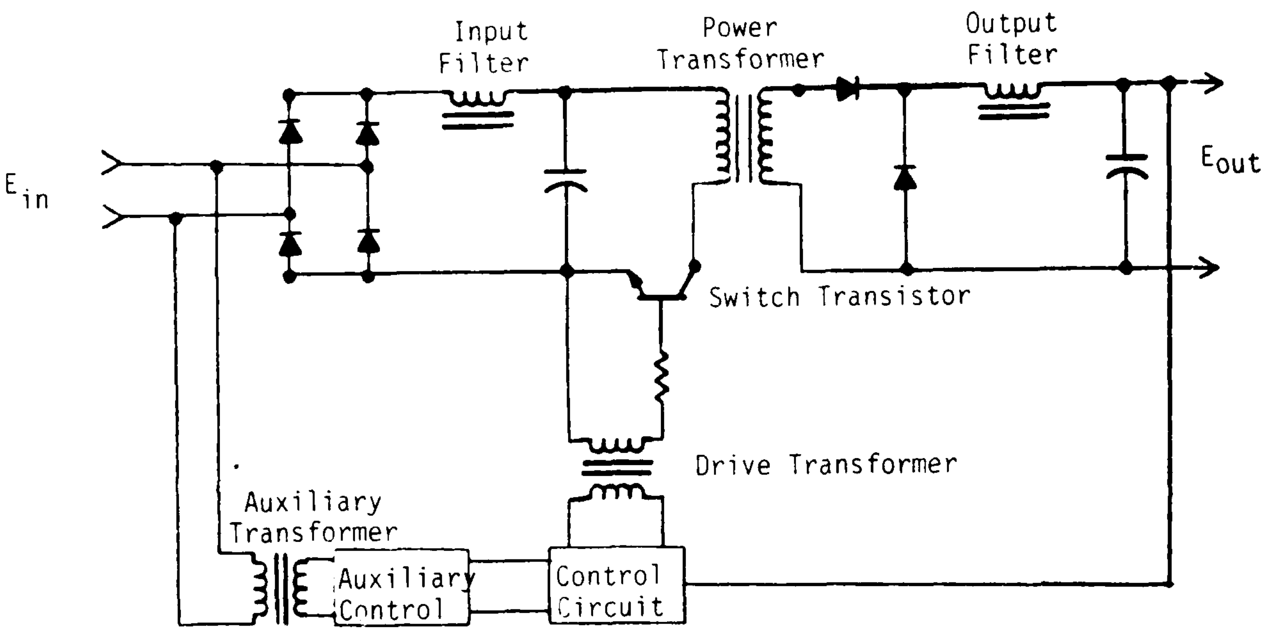


Figure 1. Basic Switchmode Power Supply Circuit

3.0 CORES

3.1 Introduction

The magnetic core materials used in most military power supply applications are silicon iron and ferrites; other core materials, e.g., molybdenum permalloy powder (MPP) and powdered iron, are used less often. The core material characteristics, which are most important to the magnetics designer, include the following:

- saturation flux density
- permeability and its temperature coefficient
- core hysteresis power loss*
- core eddy current power loss*
- Curie Temperature (temperature above which the magnetic properties deteriorate rapidly)

* In practice, these are usually lumped together as "power loss".

A larger maximum flux density reduces the number of turns the magnetic component requires. In military applications, a high Curie temperature is needed for most designs. Hysteresis and eddy currents result in magnetic losses that dissipate power and cause core heating. High permeability is also important because it translates into high inductance and thus lower magnetizing current. Core materials, core configurations, and their applications are discussed in the following paragraphs.

3.2 Silicon Iron

3.2.1 Silicon Iron Characteristics

Silicon iron cores are classified as laminated cores which are either tape wound (also referred to as strip wound) cores or punched lamination cores. Typical characteristics of silicon iron cores are the following:

Maximum Flux Density (Bmax) = 15,000 gauss (or greater depending upon frequency)

Permeability (m) = 10,000 or more gauss per oersted (Oe) depending upon flux density and frequency

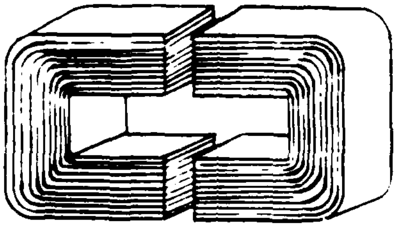
Core loss = 0.9 watts per pound at 60 hertz and 15000 gauss flux density or 9.0 watts per pound at 400 hertz

Core loss in silicon iron alloys, at least for grain oriented materials, decreases as the temperature is raised from zero to 600^{o}C. The decrease is about 10% from 0^{o}C to 150^{o}C.

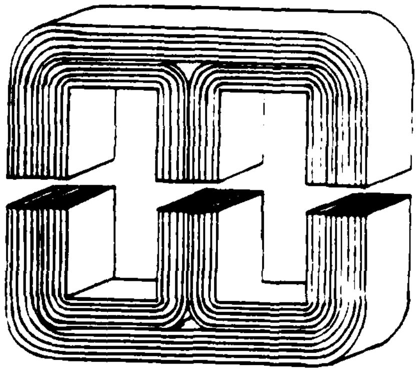
3.2.2 Core Configurations

1. Tape Wound Laminated Silicon Iron Cores

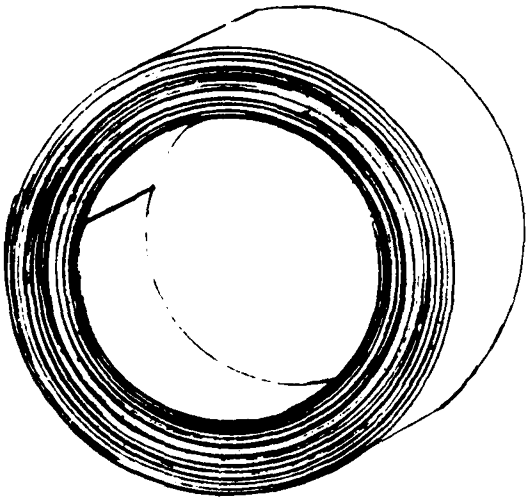
The tape wound laminated cores are made from silicon iron that is rolled into different thicknesses and cut into strips. The strips are then wound into different core configurations, e.g., C-cores or ring cores with rectangular core cross sections. See Figure 2 for illustrations of the different core configurations. The tape wound core was developed to take full advantage of the characteristics of grain-oriented material. Its best characteristics are in the direction of rolling which is in the length of the strip. National Magnetics Inc. has over 200,000 mandrel sizes. On any mandrel they can wind a core with any thickness material available, any width, and any buildup, all without a tooling charge. If they do not have on hand the required mandrel, there is a tooling charge of about $20.00. The National Magnetics Inc. [ref. 1] catalog has about 100 pages of tape wound cores listed (single-phase, three-phase, and ring in A, H, L, and Z material). Westinghouse [ref. 2] lists about 25 pages of tape wound cores in their handbook. Arnold Magnetics [ref. 3] lists about 100 pages of tape wound cores in their part number index (not their catalog).



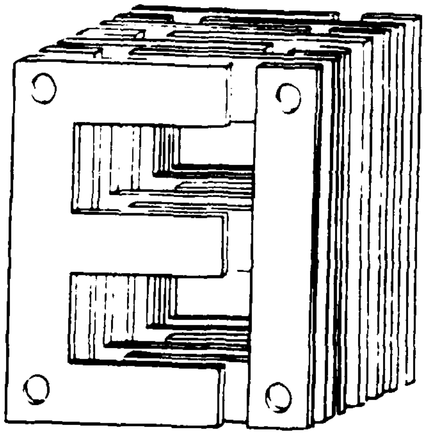
A. Tape Wound C-core



B. Tape Wound 3-Phase Core



C. Tape Wound Ring Core



D. Punch Laminated E-I Core

Figure 2. Example Silicon Iron Core Configurations

2. Punched Lamination Silicon Iron Cores

Punched lamination cores are made from silicon iron that is rolled into different thicknesses and the piece patterns required are then punched out. These punched laminations are then physically stacked and assembled to make the core with the desired cross-sectional area. See Figure 2.D.

Punched laminations are available in a variety of styles. There are EI, EE, U, F, etc., types in various thicknesses in various alloys for different applications. Many of the available types are "scrapless" configurations (no waste material when punched). Inevitably, these give a poor ratio of window area to core cross section unless the stack is small, and then the proportions are far from optimum. Without automatic stacking machines, which the Naval Avionics Center (NAC) does not have, punched laminated cores are expensive to use. They are also difficult to store in order to avoid damaging. Only partial use is made of the grain orientation feature available in silicon iron alloys. Finally, nonstandard sizes and shapes are essentially out of the question because expensive new dies would be required.

In view of the above drawbacks, punched laminations are not very satisfactory compared to tape wound cores. Punched laminations, when used, are purchased for the application and are not stocked at NAC.

3.2.3 Applications and Core Selection

Laminated cores are most suited for the low end of the frequency spectrum i.e., power line frequencies. The tape or punched lamination thickness determines the highest frequency at which the core is normally used. The available tape and punched lamination thicknesses and the nominal upper frequency of application are listed below:

- 12-mil* tape or punched lamination: 60 hertz
- 4-mil tape or punched lamination: 400 hertz
- 2-mil tape lamination: used primarily for pulse transformers and charging chokes for frequencies from 1 to 20 kilohertz
- 1/8, 1/4, 1/2 and 1 mil tape: used primarily for toroids in square wave and pulse applications at high frequencies

*1 mil = .001 inch

For 60 hertz single-phase and three-phase transformers and chokes for filtering rectified single-phase power (120 hertz ripple), 12-mil laminated cores (either tape wound C-cores or punched laminations) are ordinarily used.

Rectified three-phase power (6-pulse or 12-pulse), either 60 hertz or 400 hertz, does not oruinarily need a choke input filter because the ripple magnitude is low (13% for 6-pulse systems and 3.5% for 12-pulse systems). A three-pulse system would have 50% ripple and would need a choke input filter.

For 400 hertz single-phase and three-phase transformers and chokes for filtering rectified single-phase power (800 hertz ripple), 4-mil laminated cores (either strip wound C-cores or punched laminations) are usually used. When a choke is required for filtering rectified 3-pulse, 6-pulse, or 12-pulse power, the same core material that is appropriate for the preceeding transformer, if present, is usually used.

Two-mil C-cores may be used for the input choke of a switching regulator power supply operating at 20 kilohertz. At higher frequencies, ferrite is more suitable.

It should be emphasized that chokes for filtering rectified power line frequencies will usually require silicon iron laminated cores. Chokes to prevent feedback of switching frequency harmonics to the power line will usually require ferrite cores. Thus a choke of each type may be required in certain power conditioning systems.

For silicon iron cores, the first core selection choice should be from the CODED-T listing. CODED-T is a computer program developed at NAC that aids in transformer design [ref. 4]. After the CODED-T consideration, select one of the following in descending order of preference:

- Preferred core from the National Magnetics catalog [ref. 1], Arnold catalog [ref. 3] or Westinghouse catalog [ref. 2].
- Nonstandard core from National Magnetics catalog [ref. 1] or from Arnold Magnetics catalog [ref. 3].
- Special core made on an existing mandrel.
- Special core made on a new mandrel.

The CODED-T core list does not include tape wound, 12-mil, 3-phase cores for 60 hertz applications. Punched lamination part numbers will appear on CODED-T print-outs for this application. The procedure for forcing the CODED-T program to provide a 3-phase, 60 hertz tape wound core design is discussed in the transformer design section (Section 5.1.3).

National Magnetics gives good delivery on special cores. The ordering procedure is very simple. Using an existing mandrel if possible, determine the required core dimensions and then call National Magnetics. If they have no catalog number for a core of the required dimensions, they will assign a number and quote the weight, price, and delivery.

3.3 Ferrites

3.3.1 Ferrite Characteristics

Ferrite is a dense, homogeneous ceramic structure made by mixing, pressing, and sintering iron oxide (Fe_2O_3) with oxides or carbonates of one or more metals such as manganese, zinc, nickel, or magnesium. Ferrite magnetic cores are normally molded, but can also be machined from block material. The greatest advantages of ferrites are their low eddy current and hysteresis losses over a wide frequency range. Typical characteristics of ferrite cores are as follows:

Bmax = 3,000 Gauss approximately
μ = 2,500 to 6,500 Gauss/Oe depending upon temperature and flux density

Total loss is about 90 watts per pound or about 1 watt per cm^3 at 3k Gauss flux density, 50k hertz frequency, and $100^{\circ}C$.

Curie Temperature is greater than $230^{\circ}C$ for power materials.

3.3.2 Ferrite Core Configurations

Ferrite cores are available in many configurations: Pot, EC, EI, U, etc., and many variations of some of them. The possibilities are limitless. Figure 3 illustrates some of the various ferrite core configurations. Ferrite pot cores

a) Pot Core

b) EI Core

c) EC Core

d) U Core

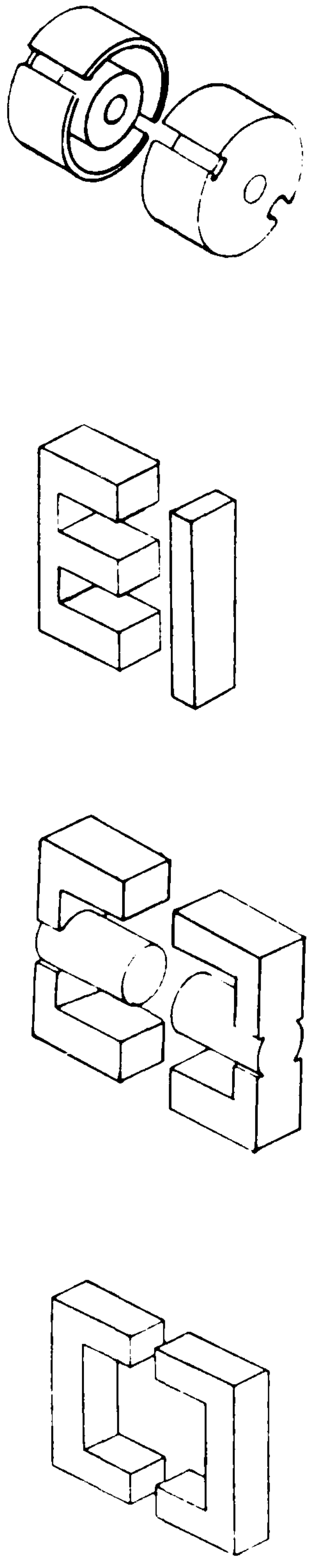


Figure 3. Ferrite Core Configurations

are available in nine standard sizes (Figure 4). The last two digits in the pot core number represent the core height, in millimeters, for the two pot core halves, and the first one or two digits represent the pot core diameter in millimeters.

3.3.3 Ferrite Applications and Core Selection

Ferrite is the material ordinarily used for transformers and chokes for applications at 20 kilohertz and up. In this frequency range, its losses are lower than those of metal cores (silicon iron, permalloy, orthinol, etc.). Because ferrites have fairly low Curie temperatures, only a few ferrite material types are usable up to 125°C. The common ferrite core material designations vary between companies. A comparison of the preferred power ferrite materials and their manufacturers is shown in Table 1.

TABLE 1. PREFERRED POWER FERRITE MATERIALS

MFG.	MATERIAL DESIGNATION	POWER LOSS*	INITIAL PERMEABILITY	CURIE TEMP. DEG.C
Magnetics	P	300	2500	230
Inc.	F	300	3000	250
Ferroxcube	3C6A	N/A	2000	≥200 (new material)
Inc.	3C8	500	2700	≥210
TDK Inc.	H7C1	250	2500	≥230

* Power loss in milliwatts (mW) per cm^3 @2K gauss, 100°C, and 50K hertz

Pot Core Number	OD (Max)	H (Max)	D max/min	Core Gross Cross-sectional Area, Ac nom. (min)	Weight	Eff. Avg. Path Length (le)	Eff. Vol. (Ve)	Eff. Core Area (Ae)
	in	in	in	cm^2	oz	cm	cm^3	cm^2
905	.366	.212	.083/.079	.0788 (.0729)	.046	1.25	.126	.101
1107	.445	.262	.083/.079	.1316 (.124)	.064	1.55	.251	.167
1408	.559	.334	.126/.118	.1979 (.183)	.113	1.98	.495	.251
1811	.717	.422	.126/.118	.3604 (.337)	.226	2.58	1.12	.433
2213	.858	.536	.181/.173	.5094 (.483)	.425	3.15	2.00	.635
2616	1.024	.642	.221/.213	.7610 (.720)	.707	3.76	3.53	.948
3019	1.201	.748	.221/.213	1.147 (1.10)	1.20	4.52	6.19	1.38
3622	1.418	.866	.221/.213	1.744 (1.66)	1.91	5.32	10.7	2.02
4229	1.697	1.174	.221/.213	2.136 (2.05)	3.68	6.81	18.2	2.66

Figure 4. Standard Ferrite Pot Core Mechanical Characteristics

The ferrite core manufacturers' catalogs [ref. 5 and ref. 6] contain curves which show core loss per unit volume versus flux density with frequency as a parameter. The core loss is given in milliwatts per cm^3. These curves were derived from measurements using toroid cores. Comparisons between toroid cores and other core configurations are based on effective dimensional parameters; effective core area, effective core volume, and effective core magnetic path length. These effective dimensional parameters are a measure of the magnetically active core material for the different core configurations. The differences between these effective parameters and the actual dimensions for toroids are negligible. However, for other core configurations these differences are appreciable, and thus are accounted for by using the effective parameters. Therefore, power loss calculations, using the curves found in the catalogs, should be based on the core effective volume rather than the actual core volume.

Pot cores provide nearly complete electric and magnetic shielding for transformers. For chokes, the magnetic shielding is less complete because of the air gap in the core unless pregapped cores with the air gap only in the center leg are used.

Custom ferrite cores are available from the following manufacturers and possibly others:

Ceramic Magnetics, Fairfield, NJ
Elna Ferrite Labs, Woodstock, NY
Magnetics Inc., Butler, PA

3.4 Molybdenum Permalloy Powdered (MPP) & Powdered Iron

3.4.1 MPP and Powdered Iron Characteristics

Both MPP and Powdered Iron are molded cores that use a nonmagnetic binder. The effective permeability of these cores can have a wide range which is controlled by varying the amount of nonmagnetic binder used. A larger amount of nonmagnetic binder increases the effective distributed air gap which lowers the permeability, but permits higher DC magnetic force. A range of characteristics for MPP cores is listed below:

$$B_{max} = 1500 \text{ gauss for typical designs}$$

$$\mu = 14 \text{ to } 550 \text{ gauss/Oe}$$

Curie Temperature = $460^{o}C$, but the temperature limitation is the temperature rating of the core insulation.

3.4.2 Core Configurations

MPP is available in toroidal form only. Powdered iron and its variations are available in toroidal and other shapes.

3.4.3 Applications and Core Selection

MPP and powdered iron cores are more suited to upper audio and radio frequencies, respectively. MPP cores are not ordinarily used for transformers because of their low permeability (<550 gauss/Oe). The low permeability results in low inductance for the primary and therefore high magnetizing current. If DC flux is present in the core, and low effective permeability is needed, MPP cores may be useful. They are excellent for chokes that must have a well controlled inductance, especially in the audio frequency range, i.e., below the range used for switching power supplies.

4.0 WINDINGS

This section provides design considerations related to the different winding configurations, the size selection of magnet wire, and interlayer winding insulation.

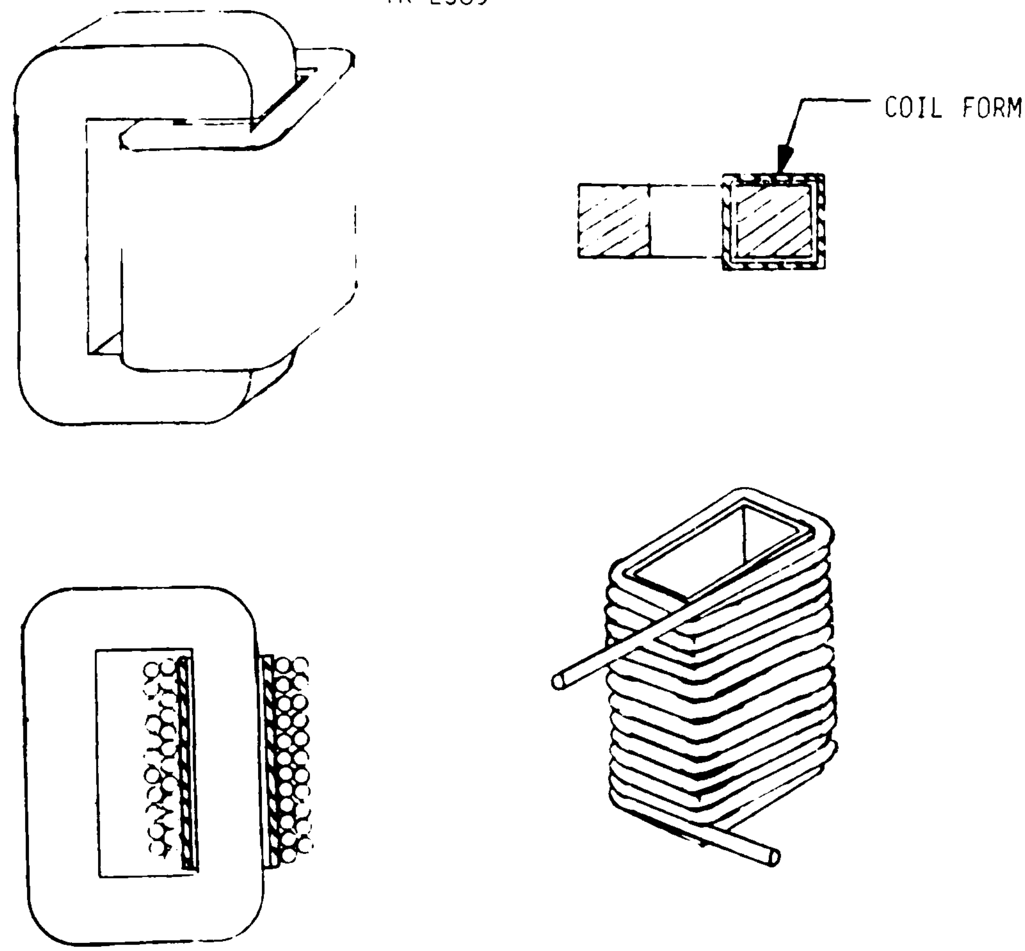
4.1 Winding Configurations

There are many types of winding configurations used in transformer and inductor constructions. Each winding configuration has characteristics which affect its suitability for certain applications. For example, winding configurations that minimize distributed capacitance are not used for power line frequencies except possibly in the high kilovolt or megavolt range. The type of winding used is dependent upon the application, frequency, core, voltage, and wire size.

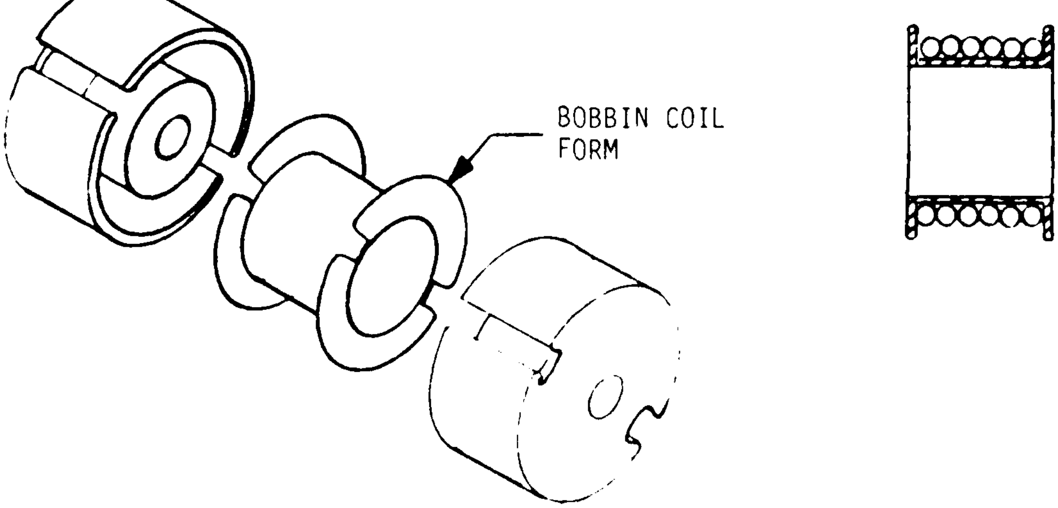
The winding configurations described are used on two principal coil form types. C-cores and punched lamination cores normally use a coil form of rectangular cross section whose size just fits over the magnetic core cross section area (Figure 5.A). The coil form material commonly used is silicone impregnated woven or felted fiber glass sheet. NAC drawing number 200AS118 lists the standard rectangular coil forms. Ferrite pot cores use a bobbin for the coil form (Figure 5.B). Standard pot core bobbin coil forms and their dimensions are shown in Figure 6. Bobbin materials commonly used are delrin, fiber glass filled nylon, and Kel-F for high temperature applications. Nylon is not a material of choice because it cannot be readily etched to allow potting materials of epoxy types to wet the surface. If necessary, nylon bobbins can be fine abrasive blasted to provide for good mechanical adhesion of epoxy.

A number of different winding configurations (Figure 7) are described below (see reference 8 for a more complete description of coil winding configurations):

- Layer windings (Figure 7.A) are neat and utilize the winding space to good advantage. Adjacent turns are as close together as the winding equipment can put them. There may be one or more layers in a winding.

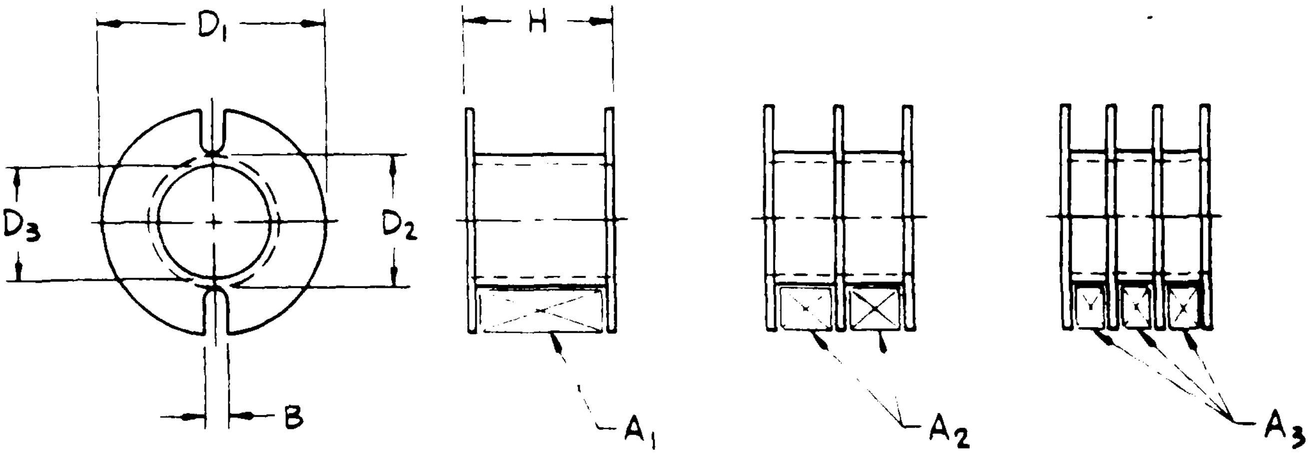


A - RECTANGULAR COIL FORM FOR C-CORE OR PUNCHED LAMINATION CORE



B - BOBBIN COIL FORM FOR FERRITE POT CORE

Figure 5. Coil Forms



											EXAMPLE OF STANDARD							
SIZE (mm)	NO. OF SECT	TOL.	A_1		A_2		A_3		B		D_1		D_2		D_3		H	
			mm²	in²	mm²	in²	mm²	in²	mm	in	mm	in	mm	in	mm	in	mm	in
9x5	1	MIN MAX	3 17 —	00492 —					1 6 —	060 —	7 23 7 34	285 289	4 67 4 78	184 188	4 01 4 11	158 162	3 40 3 50	134 138
11x7	1	MIN MAX	4 78 —	00742 —					1 6 —	060 —	8 69 8 89	342 350	5 59 5 69	220 224	4 81 4 91	189 193	4 09 4 19	161 165
	2	MIN MAX			2 16 —	00335 —												
	3	MIN MAX					1 26 —	00195 —										
14x8	1	MIN MAX	8 81 —	0136 —					1 6 —	060 —	11 3 11 5	444 454	6 98 7 24	275 285	5 97 6 10	235 240	5 28 5 49	208 216
	2	MIN MAX			3 92 —	00608 —												
	3	MIN MAX					2 35 —	00365 —										
18x11	1	MIN MAX	17 1 —	0265 —					1 8 —	070 —	14 6 14 8	574 584	8 59 8 84	338 348	7 70 7 82	303 308	6 88 7 09	271 279
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	3	MIN MAX					4 66 —	00722 —										
22x13	1	MIN MAX	26 2 —	0406 —					1 8 —	070 —	17 6 17 8	694 702	10 3 10 6	407 417	9 50 9 75	374 384	8 89 9 09	350 358
	2	MIN MAX			12 5 —	0194 —												
	3	MIN MAX					7 87 —	0122 —										
26x16	1	MIN MAX	37 5 —	0582 —					1 8 —	070 —	20 9 21 1	824 832	12 4 12 7	489 499	11 6 11 7	457 462	10 7 10 9	421 429
	2	MIN MAX			17 3 —	0269 —												
	3	MIN MAX					10 8 —	0168 —										
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	3	MIN MAX					15 9 —	0246 —										
36x22	1	MIN MAX	71 3 —	110 —					2 8 —	110 —	29 5 29 8	1 160 1 172	17 9 18 2	705 715	16 4 16 6	645 653	14 2 14 4	560 568
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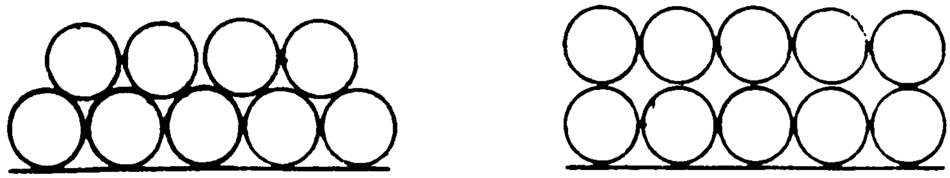
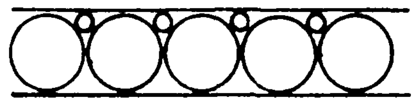
	2	MIN MAX			55 6 —	0862 —												

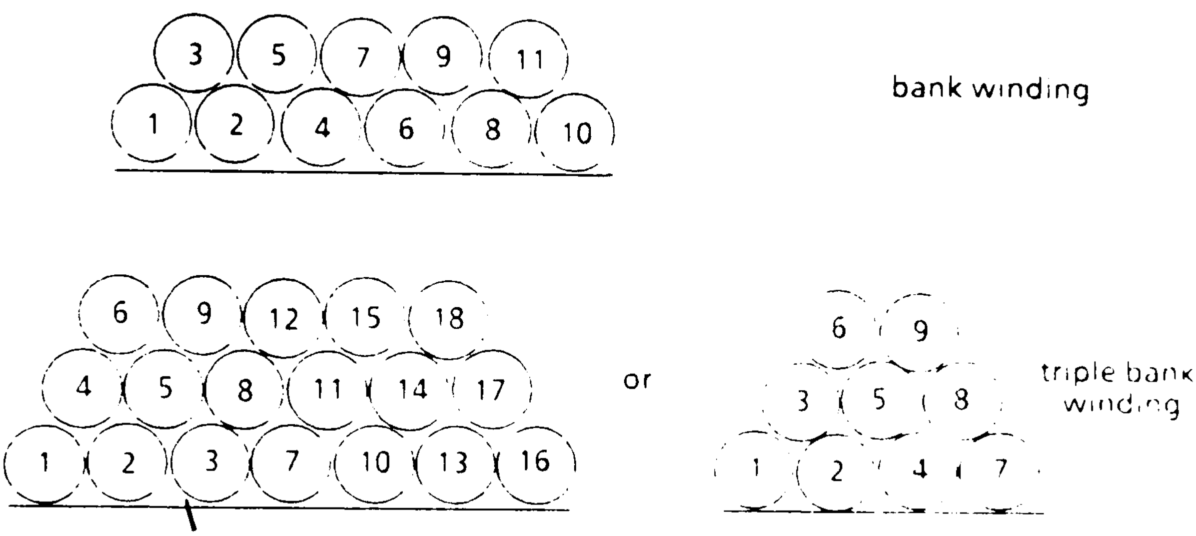
Figure 6. Standard Pot Core Bobbins and Dimensions [Ref. 7]



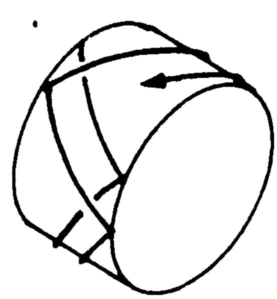
A. Layer Windings



B. Bifilar Winding



C. Bank Windings



D. Universal Winding

Figure 7. Examples of Winding Configurations

- Pseudo layer winding is a multiple layer winding and often times results from attempting to layer wind many layers of fine wire without interlayer insulation.

- Multiple strand windings are referred to as multifilar, e.g., bifilar, trifilar, etc. Multiple strands may be used to better utilize the available space or may be used to maximize the magnetic coupling between the strands in order to improve circuit performance. A sense winding may be bifilar wound with the main winding in order to obtain a signal which is electrically isolated. Because of nearly perfect magnetic coupling, the signal voltage will closely match the voltage of the main winding. If the sense winding uses wire that is 14 AWG wire sizes smaller than the main winding, it takes up no additional space (Figure 7.B). The sense winding may be put on either before or after the main winding. However, the latter method is easier.

- Bank winding (Figure 7.C), basket winding, and universal winding (Figure 7.D) and its variations are types of windings used to reduce the distributed capacitance of the winding. Capacitance is reduced by so disposing the turns or layers of turns that the potential difference between adjacent turns or layers is minimized or the projected areas are minimized. Capacitance reduction is achieved in universal windings by winding the turns in successive layers at a considerable angle to each other, instead of the normal uniform helical winding construction. In addition, the winding may progress axially from the starting area so that the start and finish ends are not close to each other.

- Scramble winding can be deliberate or the result of hand feeding from the supply spool where neatness and space considerations are subordinate. When scramble winding is deliberate, it is used to reduce distributed capacitance. It may be used when universal winding equipment is not available, but the distributed capacitance will be greater than for universal.

- A solenoidal winding, consisting of two layers wound left to right in the first layer and then right to left in the second layer, has the turns having the highest potential difference adjacent. Bank winding avoids this by having the third, fifth, seventh, etc. turns wound on top of the first, second, fourth, sixth, eighth, etc. turns. Bank winding requires a threaded form and/or wire insulation that keeps each turn in place.

- Staggering bifilar winding starts and finishes can be advantageous. Transformers for DC to DC converters usually operate at frequencies that require ferrite cores (usually pot cores), with the windings on bobbin coil forms. In order to best utilize the available space when the winding is to be bifilar, the starts and finishes may be "staggered"; the start ends of the two strands are placed in separate slots located 180 degrees apart on the same side of the bobbin. Each strand is wound with the same integral number of turns and will end in the slot where it started or at the opposite end of the bobbin. Staggering is most useful when the wire diameter is greater than about 5% of the bobbin width or larger than about AWG 20.

C-core transformers and inductors commonly use true layer windings with interlayer and interwinding insulation. If a bobbin is used (large volume production at low or medium power), the windings may be pseudo layer wound. Bobbins are usually used on ferrite cores unless the winding is a self-supporting type such as a universal winding.

The reduction of distributed capacitance becomes important when the resonant frequency of the winding is to be controlled, especially at higher voltages and higher frequencies. Higher voltage at a given frequency requires more turns for that winding, and more turns on a winding increases the winding capacitance. The resonant frequency is inversely proportional to the square root of the product of the winding inductance and the distributed capacitance, and the inductance is proportional to the number of turns squared. Thus, the resonant frequency is inversely proportional to the turns when the total winding capacitance is fixed. But since the capacitance increases with the number of turns, the resonant frequency decreases at a faster rate than that due to just an increase in inductance resulting from an increase in the number of turns. The bank winding configurations and the universal winding configurations discussed above can be used to minimize the distributed capacitance of a winding.

4.2 Magnet Wire

4.2.1 Wire Size

The choice of appropriate wire size for magnetic designs is based on the voltage drop and the heating effect ($I^2 rmsR$) in the magnetics' windings. Root Mean Squared (RMS) current is given by:

$$\text{True RMS Current} = \text{Irms} = (I^2 ac + I^2 dc)^{0.5}$$

A very important consideration is to calculate the true RMS current based on the distorted load current waveforms that result from rectification and filtering, e.g., capacitor and inductor-capacitor filters. Any load current distortion from the normal sine wave normally results in increased RMS current. These RMS current considerations are discussed in Appendix B.

In low voltage applications, voltage drop considerations dominate the choice of wire size. If heating were the only consideration, one could use as little as 400 circular mils per ampere RMS (CM/A) in small transformers or inductors and 500 to 600 CM/A in larger units. Still larger units will require 800 to 1000 CM/A.

However, 400 CM/A may give higher resistance than desired unless the resistance is desired in an inductor for damping within the circuit. Designing resistance into the winding for damping is permissible only in small units because of temperature rise. Also, copper has a large temperature coefficient of resistance (0.39%/$^{\circ}$C), thus the damping will change with temperature if the winding resistance dominates. It is preferable to use wire of suitable cross-sectional area on the required core and use an external resistor to make up the additional ,istance required for circuit damping. The change in resistance, due to change of temperature, for copper wire is given by:

$$R_2/R_1 = 1 + .00393\ (t_2 - t_1)$$

Where R_1 = resistance at temperature t_1 = + 20°C

R_2 = resistance at temperature t_2

In applications for high currents at high frequencies, the skin effect may require the use of strip copper rather than round wire. Alternatively, two or more strands of smaller round wire may be paralleled for increased ratio of surface area to cross-sectional area.

Litzentraht ("Litz") wire is another alternative for the reduction of skin effect in high frequency windings. Litz wire consists of many strands of insulated fine wire woven together in such a way as to have each strand appear at the surface at regular intervals along the length of the combination. This construction is an attempt to improve the ratio of the AC resistance to the DC resistance for frequencies up to about 500 KHz. At higher frequencies capacitance effects negate the advantages of Litz wire.

True Litz wire is very difficult to use if it has high temperature enamel insulation. Each strand must be stripped and soldered at each end. Wire coated with a material which will melt at soldering temperatures and also meets military temperature requirements is available.

Another multistrand wire is made up of twisted strands and, therefore, is not the equal of Litz wire for the amount of copper present. It is an improvement over solid wire and can consist of fewer strands of wire which are larger than those in true Litz wire. These fewer strands can be individually stripped and soldered if the extra effort is justified.

4.2.2 Wire Insulation

Magnet wire is available with various coating materials, e.g. polyester, nylon, polyester-amide, polyimide, and in combinations of them. The rated maximum operating temperatures for magnet wire with these coating materials are found in Federal Specification J-W-1177. The insulation for each wire size and coating thickness has a specified minimum breakdown voltage. Over a portion of the size range the coating can be single, double, or triple. A double coating is commonly used. At the Naval Avionics Center, grade H2 is the only magnet wire stocked. Grade H2 is double coated with polyester-amide-imide and is designated Class 180, i.e., rated for 180^{o}C.

The minimum breakdown voltage ratings for grade H2 vary from 3700 volts RMS at 60 Hertz for AWG 4 to 1300 volts RMS for AWG 44. It is good practice to operate at no more than about 20% of the minimum breakdown voltage rating.

4.3 Winding Insulation

4.3.1 Winding Insulation Materials

Insulation used in transformers and inductors serves several purposes. The primary function is to provide electrical isolation between points of different potential. Another purpose is to facilitate construction by providing a level surface for the next winding. For this latter case, insulation, in the form of sheet material, is used between layers or windings which also serves as interlayer or interwinding insulation.

A low viscosity liquid may be used for impregnation to give overall protection against voltage breakdown. If the impregnant solidifies during cooling or curing by chemical reaction of constituents, the impregnant holds the windings and turns in place.

Viscous fluids are used for potting or molding. Other fluids can be used for one-step impregnation and potting. Glass in the form of cloth, fiber, or mat when impregnated with epoxy, silicone resin or melamine is a good material for coil forms, terminal boards, printed circuit board, etc.

Another excellent insulating material is parylene, an organic chemical compound applied by vapor deposition. There are no pin-holes and coverage of sharp points is the same as flat or concave surfaces. The process is complete when the coated parts are removed from the deposition chamber. The temperature of the parts during this process does not rise more than a few degrees above room temperature.

4.3.2 Characteristics Required of Insulating Materials

The primary requirements for winding insulating materials are adequate temperature rating, dielectric strength, flexibility in strip form for interwinding and interlayer insulation, non-brittleness, resistance to tearing, availability in required thicknesses, and compatibility with the other materials present.

Materials for bobbins must be moldable or machinable if small quantities are to be made. Short glass fibers are used as filler for moldings such as bobbins. The glass fibers increase the rigidity and useable temperature limit of moldable materials such as delrin and nylon. Nylon is not a material of choice because of its moisture absorption characteristic. Delrin is much better, having low moisture absorption and good dimensional stability. Torlon, by Amoco, is a relatively new material with excellent properties for bobbins.

Naturally occurring organic materials such as cotton and silk are very unsatisfactory because they absorb water, mildew, or rot, are attacked by vermin, and have low temperature limits. The newer materials are synthetics and surpass the older materials in performance characteristics. Some naturally occurring inorganic materials are quite satisfactory if sufficiently pure. Mica is an outstanding example, but synthetic mica is better. Some forms of clay such as bentonite are used as fillers for encapsulants for transformers and inductors. Sulfur is a good insulator within its temperature limitations and applicability.

Some of the chemical classes of insulating materials are polyamides (nylon), polyimides (kapton), copolymers of these, polyesters, and many others. There are both liquids and solids consisting of chlorine, fluorine, hydrogen, and carbon. Fluorinert, by 3M, is an excellent insulator available in a wide range of boiling points.

Teflon by Dupont and Kel-F by 3M are excellent materials, especially for applications requiring rod or sheet material. Kel-F is somewhat easier to machine because of its higher modulus of elasticity. Teflon can be improved by heat treatment to harden it. This material is especially good for high frequency and high voltage applications because of its low dielectric constant, high dielectric strength, and low moisture absorption.

5.0 TRANSFORMER DESIGN

This section deals with the design of two types of transformers: 60 and 400 hertz power line frequency transformers, and high frequency switchmode power transformers.

All applicable information on the input, output, performance requirements, etc., should be available before starting the transformer design. Information should include power line frequency, voltages, secondary currents with due allowance for the nature of the load (see Appendix B), environment, cooling methods, and size and weight limitations. The design time will be reduced by knowing various details of the application. It is difficult to completely specify magnetic components for some applications. Hence, any additional information permits the magnetic designer to use his ingenuity in the transformer design.

5.1 Design of 50/60 and 400 Hertz Power Transformers

Silicon iron is the preferred core material for these power line frequencies. In most cases, tape wound silicon iron cores will be used. The transformer design challenge for a single-phase transformer is to select the proper C-core and obtain a proper fit of the required primary and secondary windings on the C-core. Figure 8 shows a C-core with the standard dimensioning designators. Figure 9 shows a cross section of a C-core and winding. The basic equations used in the design process are stated below.

$$E_{rms} = \pi\sqrt{2}\, B \times A_i \times N \times f \times 10^{-8}$$

$$\frac{N_s}{N_p} = \frac{E_s}{E_p} = \frac{\hat{E}_s}{\hat{E}_p}$$

Where:

Erms = Primary voltage in Vrms
B = Peak flux density in gauss
N = Number of turns on winding

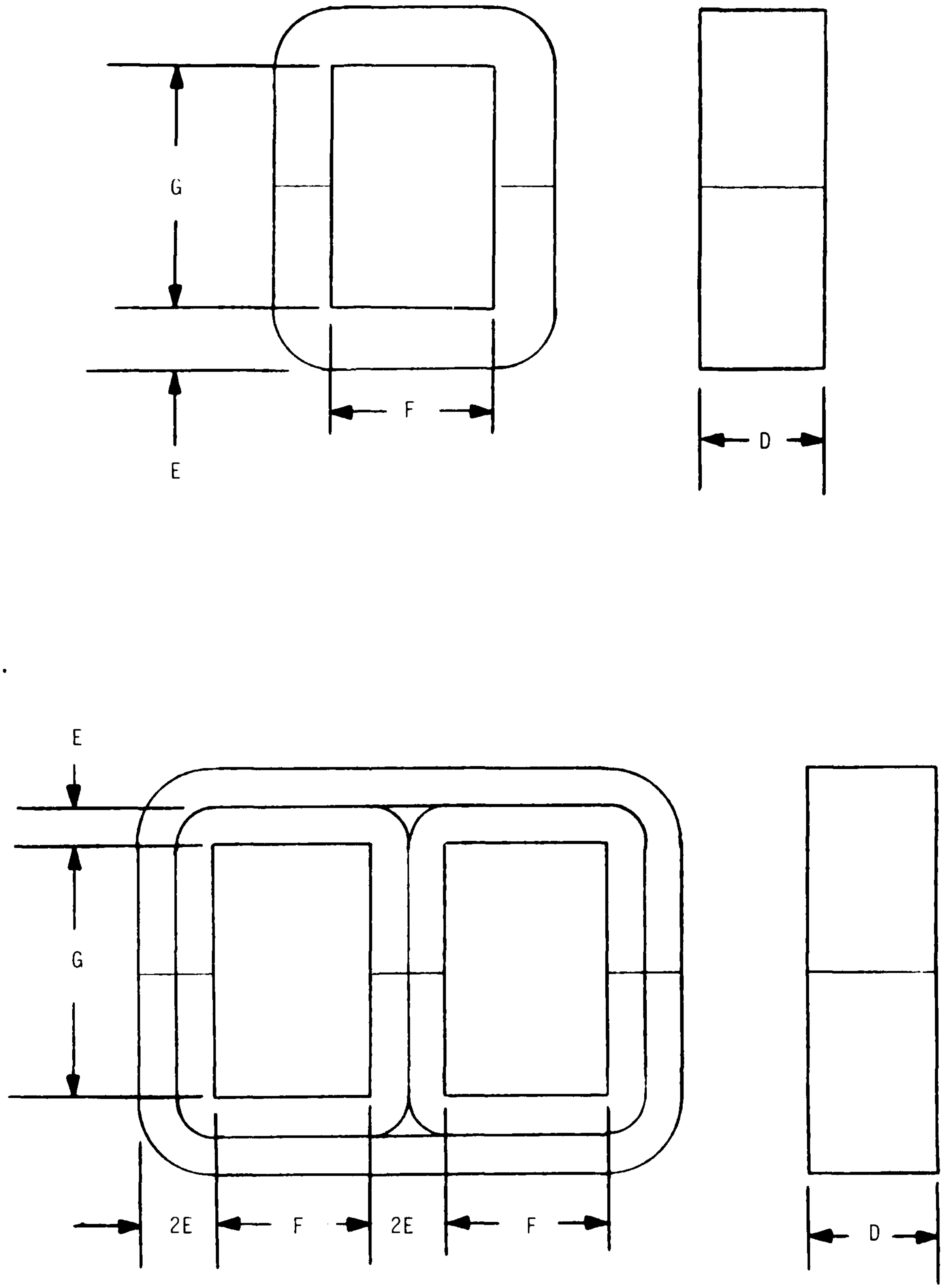
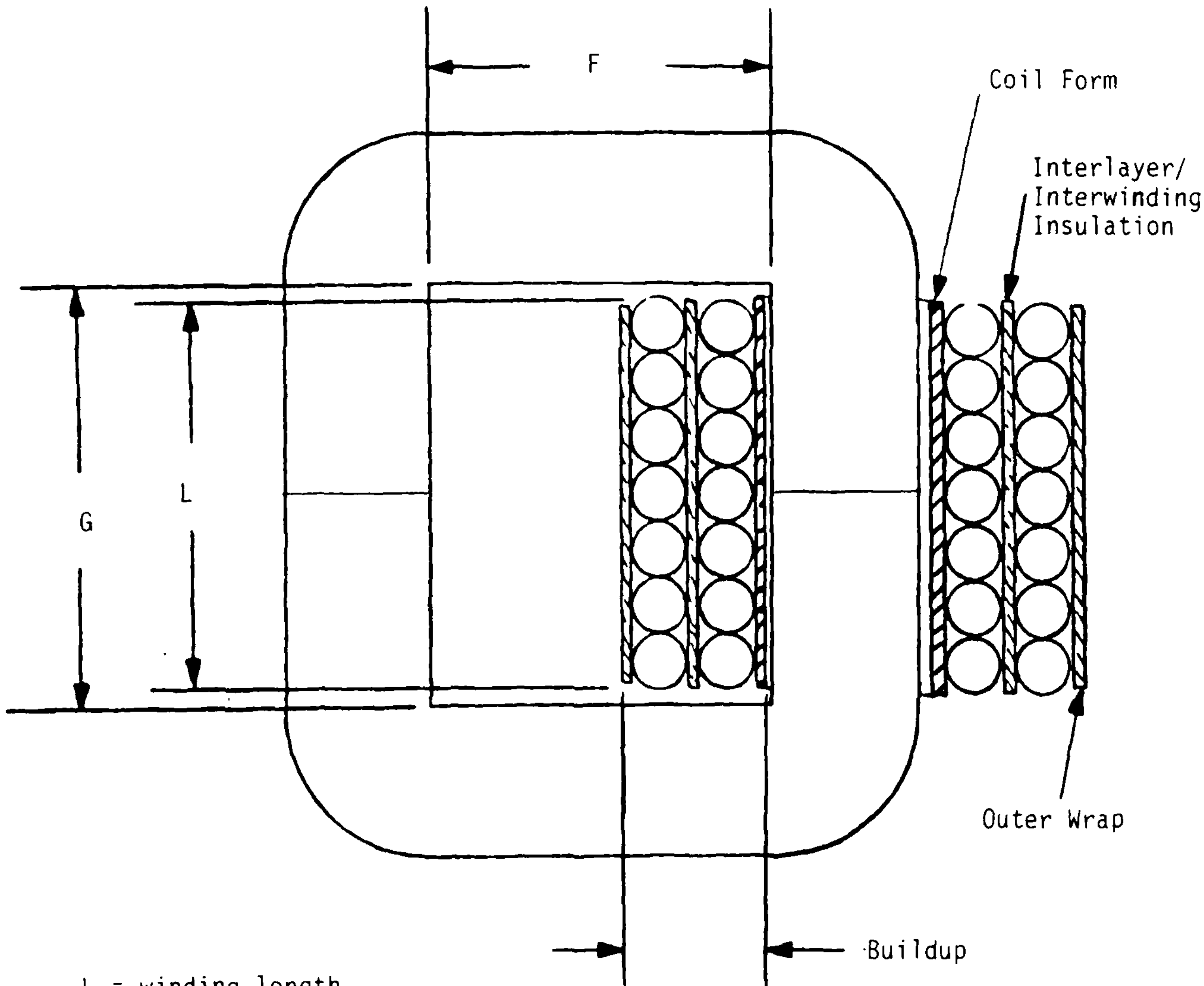


Figure 8. Tape Wound C-core Standard Dimension Designators for Single-phase and Three-phase Cores



L = winding length

$$\text{Margin} = \frac{(G - L)}{2}$$

Figure 9. Cross-Section of C-core and Winding

N_p = Number of primary turns
f = Frequency of applied waveform in hertz
N_s = Number of secondary turns
E_s = Secondary RMS voltage
E_p = Primary RMS voltage
$\hat{E}_s$ = Secondary peak voltage
$\hat{E}_p$ = Primary peak voltage
Ai = Net core cross-sectional area in cm^2*

*The net core cross-sectional area, Ai, is the actual area of the iron at the cross section. This net area differs from the actual core cross-sectional area, Ac, or gross area, by a factor of Kc, the core stacking factor. The stacking factor accounts for the cross-sectional area of insulating adhesive between strips or laminations of iron for tape wound and punch laminated cores respectively. Thus, the gross cross-sectional area for such cores is the sum of the iron area Ai and the inter-laminate adhesive. For cores of homogeneous materials such as ferrites, the net area Ai is equal to the gross area Ac.

Flux produced by the secondary ampere turns cancels the flux produced by the load component of the primary ampere turns. That is, no net flux is produced by the load currrent. Therefore, saturation is not produced by overloading. The peak flux is produced by the exciting current which equals (peak voltage)/ωL plus the current to support the core loss.

5.1.1 Manual Approach to Power Line Frequency Transformer Design

The following procedure is suggested for the design of a power line frequency transformer:

1. Calculate the total Volt-Ampere (VA) output power (DC component and ripple components combined as the square root of the sum of the squares) from the rectifiers or directly from the secondaries if rectifiers are not used. Primary magnetizing current may be neglected because it is a quadrature component and is small compared to the primary load current in a well designed and constructed transformer.

The VA rating of the primary is not necessarily equal to the sum of the secondary VA ratings. For example, the primary VA rating of a transformer is the same whether the secondary supplies a bridge rectifier or is center-tapped for a center-tap rectifier circuit, where the load circuit following the rectifiers is the same in each case. The VA rating of the center-tapped winding is $\sqrt{2}$ times the actual VA output of the bridge rectifier circuit (rectifier drop and IR drop of windings neglected).

2. Divide the actual VA output by the estimated transformer efficiency. The result is the primary VA input. Typical transformer efficiency ranges between 0.8 and 0.95.

3. Calculate the design circular mils per ampere (CM/A) for each winding using Equation (56) from Appendix A. The required CM/A increases for increasing current in the winding.

4. Using the true RMS current for each winding, calculate the needed CM and thus the wire size for each winding.

5. Obtain a preliminary estimate of the core cross-sectional area and the available window area (AcAw) product required using Equation (49) from Appendix A.

6. Using the "Rule of 2s" as a guide, choose a trial core from the manufacturer's catalog. Some compromise may be necessary. The "Rule of 2s" relationship is based on a mathematical analysis used to minimize size or weight of a single-phase transformer (or inductor) on a C-core, and is expressed as follows:

$$D \cong (2.2 \text{ to } 2.0) \times E$$

$$G \cong (2.2 \text{ to } 2.0) \times F$$

$$G \times F \cong (2.2 \text{ to } 2.0) D \times E$$

where D, E, G, and F are the standard C-core dimension designators (Figure 8).

The constants are not hard and fast. They depend somewhat upon the fill factor in the window. Fill factor is defined as the actual total cross-sectional area of conductor in the window divided by the window area. An approximate fill factor is 0.4 for laminated cores, particularly grain oriented silicon iron cores.

7. Calculate the primary turns for trial core selected in Step 6 using Equation (1) from Appendix A. The maximum flux density used for silicon iron core designs is about 16,000 gauss.

8. Calculate corresponding turns for the secondaries. Increase secondary turns by a factor of 1.02 for 400 Hz supply (or 1.05 for 60 Hz) to allow for IR and reactance drop. Thus: $N_s = N_p \times (E_s/E_p) \times (1.02)$ or (1.05) for 400 Hz and 60 Hz transformers respectively. Alternatively, the primary turns may be decreased, provided the flux density is acceptable.

9. Calculate the approximate margin for each size wire being used.

Margin $\cong 0.8 \times (\text{wire diameter})^{0.5}$ to $1.0 \times (\text{wire diameter})^{0.5}$ at each end of each layer.

Note: Margin should not be smaller than $\cong$ .06 inches and need not be greater than $\cong$ .25 inches. When wire larger than AWG 12 is required, it is better to parallel two or more strands of wire to give required cross section.

G - (2) x (margin) = maximum length of one layer of a winding.

Maximum effective diameter of the wire is approximately (1.02) x (maximum wire diameter listed in the wire table). Thus, the approximate maximum length of one layer of a given wire size is:

(N+1) x (1.02) x (maximum wire diameter listed in tables)

Where N = the number of turns in the layer.

For purposes of calculating layer length, margin, and buildup, "maximum wire diameter" is the maximum diameter including the wire insulation.

Design trade-offs include: using more than one layer, using smaller wire, using a slightly smaller margin, or using a core with larger G dimension.

A technique that conserves window space by reducing coil buildup is to use two strands of wire three AWG numbers smaller and bifilar wind them. The coils can then be connected in parallel or in some cases they can be used as a center-tapped winding. The latter avoids having to tap in the middle of a layer.

10. Calculation of Coil "buildup"

buildup = coil form maximum wall thickness

+ summation of wire diameter multiplied by number of layers of that diameter
+ total of interlayer insulation, if used
+ total of interwinding insulation
+ outer wrap on coil
+ 5% to 10% of the sum of the above to allow for bulge*

* If conductors are small so that each winding is compact, 5% is enough. If the conductors are large and therefore do not make a coil with nearly plane faces, allow 10%. Also, silicone varnished fiber glass for interlayer and interwinding insulation gives more bulge than Kapton tape.

Multiply the buildup sum by number (1 or 2) of coils in the window.

Along with filling the layers reasonably well in the G dimension, one must not exceed the F dimension with buildup of coil depth of the one or two coils that are to occupy the window. Calculated total buildup should not exceed approximately 90% of the F dimension of the window. The other 10% is an allowance for the difference between the calculated buildup and actual results. If less than about 75% of the window width is full, a different core should be used. Alternatively, larger wire could be used for better regulation and lower temperature rise. More turns could be used on each winding for lower flux density and thus lower core loss.

11. Iterate as required to get desired final result.

5.1.2 Alternate Ways (In Order of Preference) To Choose The Core Dimensions

1. Use the CODED-T program to obtain a preliminary design, a core part number, and wire sizes.

2. Use Equation (56) from Appendix A to pick CM/A and Equation (49) from Appendix A to pick the AcAw product.

3. Given fixed maximum overall dimensions, use cut-and-try design changes until acceptable results are achieved.

4. Sometimes changing the D and E dimensions of the core while keeping the DE product constant or nearly so will be the key. For example, keeping F, G, and DE product constant, a ΔE change in E changes the overall dimension in the F direction by $2\Delta E$ for a single-phase transformer or $3\Delta E$ for a three-phase unit.

5. Another approach is to design the windings and dimension the core to fit (while satisfying all requirements).

6. Sometimes one has no choice but to use the core used in the previous design for the application, even though the requirements (power or number of secondaries) have been changed (increased, usually). Sometimes this will result in greater temperature rise than the previous design unless novel

techniques and refinements are used to better utilize the window space. As a last resort let it run hotter (use higher temperature wire and insulation) or provide fins for extracting heat.

5.1.3 CODED-T Approach To Power Line Frequency Transformer Design.

The Naval Avionics Center developed a computer program for the design of C-core and punched lamination silicon iron cores [ref. 4]. This program can now be accessed via any remote CRT terminal that is linked to NAC's central computer. See Reference 9 for instructions for using this program on the NAC central computer. A copy of Reference 9 can be obtained from the Computer Operations Branch 723. A single-phase transformer design example using the CODED-T program is included in Appendix C.

Because 12-mil tape wound cores for the design of 60 hertz three-phase transformers are not included in the CODED-T data base, a modified design procedure must be employed. Use the CODED-T program to design a single-phase split coil transformer which has 2/3 the required three-phase power rating, twice the primary and secondary leg voltages required, and currents equal to those required. Specify 5 to 10^{0}C lower temperature rise than normally permissible because the center leg of the three-phase unit doesn't cool as well as the outer legs. Add another leg to the core and put the same number of turns of each wire size on the third leg.

The same design procedure may also be used to design a 400 hertz three-phase transformer. If a unit is needed for which there is no three-phase 4-mil strip wound core in the data base, one can design around a single-phase core which is in the data base and then refer to a catalog for a suitable three-phase core or order a special core as required.

If one calls for a tape wound core (either three-phase or single-phase, 4-mil or 12 mil), but a suitable core is not available in the data base, the program will default to a punched lamination. If that is not acceptable to the user, he may use the above procedure to design a unit using a tape wound core.

The question of delta versus wye for connecting windings is left to the user of the CODED-T program. The user must specify the voltages and currents of the windings according to their use in delta or wye connections. The computer program does not address the subject of connections. It simply computes the number of turns, wire size, etc, required to make the individual windings for the user's transformer.

The above statements apply to all CODED-T transformer and inductor designs whether single-phase or three-phase, at any frequency, and on any tape wound or punched lamination core.

5.2 Design of Switchmode Power Transformers

As discussesd previously, the switchmode power supply operates at frequencies above 20 kilohertz and ferrite core materials are normally used for magnetics at these higher frequencies. Because the switchmode supply is normally based on rectangular pulse voltage waveforms, the basic magnetic design equation is modified as shown below along with other pertinent design equations.

$$\text{Epeak} = 4B \times Ai \times N \times f \times 10^{-8} \qquad \text{square wave}$$

$$(\text{Epeak}) \times (t_{on}) = 2B \times Ai \times N \times 10^{-8} \qquad \text{rectangular wave}$$

$$E\ dc = (\text{Epeak}) \frac{t_{on}}{t_{on} + t_{off}}$$

Where:

Epeak = Peak primary voltage in volts

B = Peak flux density in gauss

Ai = Net core cross-sectional area in cm^2 = Ac, gross cross-sectional area

N = Number of turns on winding

Edc = Output voltage in Vdc

t_{on} = On time of rectangular voltage waveform
t_{off} = Off time of rectangular voltage waveform
f = Frequency of rectangular waveform

5.2.1 Manual Approach To Switchmode Power Transformers.

The following procedure is suggested for the design of switchmode transformers:

1. Calculate the power output required.

   a. Add 20% if Edc output = 5 volts
   b. Add 10% if Edc = 15 volts or higher

2. Pick a tentative core size, material, and geometry, taking into account the following:

   a. Power output - refer to Equation (70) of Appendix A or to the Magnetics Inc. ferrite catalog, p 9.5 [ref. 5].
   b. Input, output, and standoff voltages required. (High voltage designs require extra window space.)
   c. Frequency

3. Calculate the number of primary turns required using the relationship:

$$(E_p) \times (t_{on}) = 2B \times Ai \times N_p \times 10^{-8}$$

Where E_p = Peak voltage applied to primary
t_{on} = duration of one conduction period in seconds
N_p = Number of primary turns

4. Determine the secondary voltage required based on the output voltage required, an estimate of the IR drop in the choke, and the duty factor.

5. Calculate secondary turns using the relationship

$$N_s = \frac{(E_s)}{(E_p)} \times (N_p) \times (1.02)$$

Where N_s = Number of secondary turns
E_s = Secondary RMS voltage
E_p = Primary RMS voltage

a. Because of imperfect coupling between windings, the turns ratio must be greater than the voltage ratio, hence the 1.02 factor. This factor is not rigid and can be a little larger to give an integral number of turns provided the output voltage is not too high.

b. Otherwise, change the primary turns to come closer to the ideal ratio.

c. If it is necessary to decrease primary turns, recalculate the flux density. A higher flux density may require using a different core material, e.g., Magnetics Inc., P material instead of F material.

d. An alternative is to use a larger core (larger cross section).

6. Calculate the RMS current for each winding based on the circuit configuration to be used and the applicable form factor.

7. Calculate wire cross-sectional areas based on 500 or 600 CM per ampere or more in large devices. If required, due to space limitations, as low as 400 CM per ampere may be used if the heating and IR drop is acceptable.

8. Attempt to fit the windings into bobbin, if used, or the available window area. For pot cores, use Table 2 or manufacturers' catalog information giving winding space available on bobbins. Table 2 applies mainly where wire is larger than AWG 26. Also, see graph No. 7 in the Magnetics Inc., ferrite catalog, p. 5.12 [ref. 5] which applies mainly where wire size is smaller than AWG 26. Iterate steps 2, 3, 5, 7, and 8 as required.

TABLE 2. TURNS PER LAYER FOR BOBBIN COIL FORMS FOR VARIOUS WIRE GAUGES AND VARIOUS FERRITE CORE SIZES

BIFILAR TURNS IN FIRST LAYER STAGGERED STARTS	DIAMETERS PER LAYER	WIRE GAUGES FOR INDICATED CORE SIZE								
		905	1107	1408	1811	2213	2616	3019	3622	4229
2 1/2	6	27	25	22	19	17	15	13	13	10
3	7	28	26		21	18	17	15	14	11
3 1/2	8	29	27		22	19	18	16	15	12
4	9	30	28		23	20	19	17	16	13
4 1/2	10	31	29	27	24	21	20	18	17	14
5	11	32	30	28	25	22	21	17	18	15
5 1/2	12	33	31	27	26	23	22	20	19	16
6	13	34	32	29	27	24	22	20	20	17
6 1/2	14							21	20	17
7	15								21	18
7 1/2	16									19
2	5					15		12	11	
2	17								22	
9						26				

6.0 INDUCTOR DESIGN

This section discusses the design of inductors for 60 and 400 hertz frequency inductors and switchmode power supply frequency inductors, and includes a discussion of the benefits of using coupled inductors. Almost all the applications for power inductors associated with power supplies involve their use in the inductor-capacitor (LC) filter which either follows a rectifier circuit or is located at the dc input to prevent harmonic currents from feeding back on the power distribution lines. Just like the transformer design, all applicable information is desirable, e.g., the input/output voltages, current range, frequency, environment, and size and weight limitations.

6.1 Design of 60 and 400 Hertz Power Line Inductors

The actual frequency seen by the inductor in an LC filter is some integer multiple of the power line frequency. For example, an LC filter for a single-phase input voltage that is full-wave rectified sees twice the power line frequency. Thus, a three-phase input voltage that is bridge rectified will provide six times the power line frequency to the LC filter.

The concept of critical inductance is important in defining the inductance value required in LC filter applications. Critical inductance is the minimum inductance value that will ensure continuous (uninterrupted) current through the inductor at the minimum current condition. At load currents equal to or greater than the minimum load current for which the inductor has the critical value, the inductor output voltage equals the average value of the inductor input voltage waveform. Use of an inductance less than the calculated critical value, or a load current less than that required for the designed critical inductance permits the filter capacitor (which follows the inductor) to charge to a voltage approaching the peak value of the inductor input voltage waveform. Thus, the voltage regulation is poorer than if the load current did not drop below the value for which the critical inductance was calculated. Furthermore, the ripple voltage, peak current, and RMS current are higher for a given value of filter capacitor. Expressions for calculating the critical inductance for a rectified sine wave and rectangular wave are found in Appendix A.2.

As shown in the derivation of the value for critical inductance, the instantaneous current just drops to zero, but is not interrupted. For a symmetrical current wave form (e.g., sine wave or triangular) the peak value of current is twice the average current.

6.1.1 Manual Approach To Power Line Frequency Inductor Design

The following procedure is suggested for the design of a power line frequency inductor.

1. Calculate the critical inductance, Lc, using the appropriate equation from Appendix A.2. The desired inductance will be approximately 2 x Lc.

2. Calculate the AcAw product which is the product of the gross core cross-sectional area and the gross core window area. The AcAw is calculated using Equation (64) from Appendix A.

$$AcAw = \frac{(10^2)\ (L)\ (Idc + \hat{I}ac)\ (Irms)\ (CM/A)}{(\hat{B}KcKw)\ (2.54)^2} \times \frac{\pi}{4}$$

$$Kw = \frac{(\pi/4)\ (d^2)(N)}{10^6\ (Aw)} \simeq .45 \text{ to } .65$$

Ac = gross core cross-sectional area in square inches
Aw = gross window area in square inches
Kc = stacking factor of core
Kw = window fill factor of winding
L = inductance in henries
$\hat{B}$ = peak flux density in gauss
d = wire (conductor) dia in mils
CM = d^2 = circular mils
Irms = RMS current in amperes
Idc = DC current in amperes
$\hat{I}ac$ = Peak Ac current referenced from Idc in amperes (see Figure A.1)

3. Based upon the calculated AcAw product, select a trial silicon iron core that has about the same AcAw product. The "Rule of 2s" discussed previously under power line frequency transformer design also applies here.

$$D \simeq (2.2 \text{ to } 2.0) \times E$$
$$G \simeq (2.2 \text{ to } 2.0) \times F$$
$$G \times F \simeq (2.2 \text{ to } 2.0)\ D \times E$$

4. Using Equation (12) from Appendix A, determine the number of turns required for the inductance desired.

$$N = \frac{L\hat{I} \times 10^8}{Ai \times B}$$

5. Determine the appropriate wire using Irms and the desired CM/A.

6. Determine if the required number of turns will fit into the core using the wire size selected. This is known as calculating the "buildup" and requires several steps.

a. Determine the approximate margin for the wire size being used. Margin = (0.8 to 1.0) x $(\text{wire diameter})^{.5}$ at each end of the layer. Margin should be between .06 and .25 inches. If wire larger than AWG 12 is required, it is preferable to parallel two smaller strands of wire to give the necessary cross-section.

b. Determine the maximum length of one layer of a winding.

$$L_{max} = G - 2 \times (\text{margin})$$

The maximum effective diameter of a wire is approximately 1.02 x (maximum wire diameter listed in the wire table). The maximum effective diameter of the wire should include wire insulation. Thus, the approximate maximum length, Lw, of one layer of a given wire size is:

$$L_w = (N+1) \times (1.02) \times (\text{max wire dia})$$

where N = turns per layer

These equations are used to determine the number of layers required to accommodate the specified number of turns.

c. Determine the number of layers of windings and the thickness required for interlayer, interwinding, and final insulation.

d. Determine bulge. Bulge is a percentage of the sum of the winding(s) thickness, the interlayer, interwinding, and outer insulation, and the coil form wall thickness. If conductors are smaller than AWG #26, so that each winding is compact, 5% is enough. If the conductors are large and therefore do not make a coil with nearly plane faces, allow 10%. Also, silicone varnished fiber glass for interlayer and interwinding insulation gives more bulge than Kapton tape.

e. Determine coil buildup.

Buildup for each coil equals:

coil form maximum wall thickness
+ wire diameter x number of layers per winding
+ total of interlayer insulation
+ total of interwinding insulation
+ outer insulation on coil
+ bulge

Calculated total buildup should not exceed approximately 90% of the F dimension of the window. If buildup is less than 75% of the F dimension, another core should be selected.

Design trade-offs include: using more than one layer, using smaller wire, using a slightly smaller margin, using bifilar windings, or using a core with a larger G dimension.

7. Iterate as required to get desired final result.

8. The air gap, ℓg, for the core is calculated using Equation (7) from Appendix A,

$$\ell g + \frac{\ell c}{\mu \Delta} = \frac{0.4\pi N \times \hat{I}}{B} \text{ centimeters}$$

In practice the $\ell c/\mu\Delta$ term is negligible compared to ℓg. The actual air gap should be five or more times the inherent gap of the core, otherwise the air gap is too critical. Because of flux fringing at the air gap, the actual gap required in each leg will usually be larger than one-half the calculated gap. The larger the gap in comparison to the D and E dimensions of the core (especially the E dimension of a laminated core), the greater the fringing. Fringing also increases with core flux density. Windings that are close to the core legs (minimum radial clearance between windings and core) reduce fringing by forcing some of the fringing flux back into the core. The material used for the air gap spacers must be non-magnetic and electrically non-conductive.

6.1.2 CODED-T Approach to Power Line Frequency Inductor Design

Inductors may be designed using the CODED-T program, but the input data must be presented as for a transformer design. The inductor design will be optimized if the data is scaled before inputting to the computer. If the data is not scaled, the computer-generated inductor design will base core selection on the frequency of the ripple component and require a more expensive core than is actually needed. Instead of selecting a core based on the 400 Hz power line frequency, the CODED-T program core selection is based on 1200 Hz for 3-pulse rectification, 2400 Hz for 6-pulse rectification, or 4800 Hz for 12-pulse rectification. The situation for 60 Hz is analogous. The break point for the use of 12-mil and 4-mil core materials is 100 Hz. Two mil material is used instead of 4-mil material at frequencies above 1000 Hz. Scaling ensures that the transformer and inductor for a given circuit and frequency will be designed using the same material.

Application of the scaling factors to the 400 Hz, 6-pulse rectified case is shown in Table 3; the 60 Hz, 6-pulse rectified case is shown in Table 4. The factor of 10, and its square, is used in this procedure because it permits simple manipulation of the decimal point and because a change of 10 numbers in the wire gage table gives almost exactly a change of 10 to 1 in cross-sectional area.

The factor of 10/3 and its reciprocal for the 400 Hz case are derived as follows, after an arbitrary choice of 3 as the frequency modification factor.

From $$Eac = Iac \times 2 \times \pi \times f \times L$$

where

Eac = RMS voltage of the lowest frequency ripple component

Iac = RMS ac current in the inductor

f = minimum frequency in hertz

L = inductance in henries

Then

$$\frac{Eac_2}{Eac_1} = \frac{16.83 \times 2\pi \times 760 \times .05}{168.3 \times 2\pi \times 2280 \times .0005} = \frac{10}{3}$$

The mathematical justification for the inductance and direct current scaling factors is that the stored energy ($.5LI^2$) is the same for the the scaled and unscaled case.

For the 60 Hz case a factor of 10/4 is used for voltage scaling and a factor of 1/4 for frequency scaling.

Actual input to the CODED-T program will be from columns 2 through 8 in the line "Input Data" in Tables 3 and 4.

TABLE 3. APPLICATION OF SCALE FACTORS TO INDUCTOR VALUES FOR 6 PULSE RECTIFICATION OF 400 Hz(±5%) FOR CODED-T INPUT

1	2	3	4	5	6	7	8	9	10
	INDUC-TANCE in HENRIES	Eac in RMS VOLTS	Iac in mA	Idc in mA	Rdc in ohms	f min Hz	f max Hz	AWG	TURNS
REQUIRED	[illegible]	[illegible]	[illegible]	15000	.06	2280	2520	---	-----
FACTOR	100	[illegible]	[illegible]	1/10	100	1/3	1/3	---	-----
INPUT DATA	[illegible]	[illegible]	[illegible]	1500	6.0	760	840	---	-----
PRINT-OUT	.05	[illegible]	[illegible]	1500	6.0	760	840	GAGE #	N
FACTOR	1/100	[illegible]	15	10	1/100	3	3	-10	1/10
BUILD	.0005	[illegible]	179	15000	.06	2280	2520	G-10	N/10

TABLE 4. APPLICATION OF SCALE FACTORS TO INDUCTOR VALUES FOR 6 PULSE RECTIFICATION OF 60Hz (±5%) FOR CODED-T INPUT

1	2	3	4	5	6	7	8	9	10
	INDUC-TANCE in HENRIES	Eac in RMS VOLTS	Iac in mA	Idc in mA	Rdc in ohms	f min Hz	f max Hz	AWG	TURNS
REQUIRED	.0005	1.4	1303.0	15000	.06	342	378	---	---
FACTOR	100	10/4	1/10	1/10	100	1/4	1/4	---	---
INPUT DATA	.05	3.5	130.3	1500	6.0	85.5	94.5	---	---
PRINT OUT	.05	3.5	130.3	1500	6.0	85.5	94.5	AWG	T
FACTOR	x1/100	4/10		x10	x1/100	x4	x4	-10	÷10
BUILD	.0005	1.4	1303.0	15000	.06	342	378	AWG-10	T/10T

6.2 Design of Switching Regulator Power Supply Output Inductor

1. Calculate the critical inductance, Lc, using Equation (28) from Appendix A, and then the desired inductance, approximately 2 x Lc.

Because the steady magnetic flux (from the Idc) lowers the core permeability and hence the inductance below the desired value, start with an inductance approximately 10% greater than the required operating value. Thus, the design inductance L = 1.1 x 2 x Lc.

Note: In a closed loop system, the actual inductance must not vary too much or instability may result.

2. From Equation (37), Appendix A, calculate $\hat{I}$ (peak current), and from Equation (40), Appendix A, calculate Irms.

3. Using Equation (12) in Appendix A calculate the Ai x N product,

$$Ai \times N = \frac{L \times \hat{I} \times 10^8}{B}$$

Where L = inductance from step 1

$\hat{I}$ = peak current in amperes from step 2 above

B = peak flux density in gauss = 2000 to 3000 gauss

Higher flux density is normally permitted in chokes than in a transformer using same core material because only a portion of the flux is varying, i.e., there is no core loss from the D.C. or steady flux

Ai = minimum net cross-sectional area of magnetic path of chosen core in square centimeters = Ac (gross cross-sectional area, Figure 4, page 13)

N = number turns of wire

4. Pick a tentative core.

Start with the same size core as used for the output transformer. Calculate N from the Ai x N product (step 3).

5. Choose the wire size based on the Irms from step 2 using a current density of 500 to 600 CM/A or more for large units. 400 CM/A can be used if the heating and the voltage drop are acceptable.

6. If a bobbin is used, check the fit of the winding. For pot core, use Table 2 or the manufacturers' catalog information giving winding space on bobbins. (Applies mainly where wire is larger than AWG 26.) Also, see graph No. 7 in the Magnetics Inc., ferrite catalog, p. 5.12 [ref. 5] which applies mainly where wire is smaller than AWG 26.

7. Iterate steps 3, 4, 5, and 6 to get an acceptable design.

8. Calculate the air gap from $\ell g + \frac{\ell c}{\mu \Delta} = \frac{0.4\pi N\hat{I}}{B}$ in centimeters

For inductor construction, use 1.5 to 2.0 times the calculated air gap value to compensate for fringing at the gap. Normally, the air gap should not be smaller than 5 mils, otherwise the inductance value is too sensitive to small variations of the air gap.

6.3 Benefits of Coupled Inductors

Coupling of inductors is a technique for placing two or more inductor windings on a single magnetic core. The principle advantage of this design is that the critical inductance for each of the inductors on the core is based on the total magnetization of the core. The benefit is analogous to the improvement in regulation in a circuit consisting of two or more loads taking current from a single source instead of individual sources each of which must supply the peak current. With two or more load circuits, the magnetization of the core is held up by all windings instead of being solely dependent upon any one winding.

The number of turns in each winding must be in the same ratio as the turns in the windings of the preceeding transformer whose outputs are rectified and taken to the respective windings of the coupled inductor. An incidental benefit is that only one core is required instead of two or more.

7.0 DESIGN PRACTICES

A number of techniques are discussed in the design and construction of magnetics which, properly applied, will result in producible magnetic designs.

7.1 Rules of Thumb for C-core Magnetics

As a general rule, coils will be neater and therefore more symmetrical and will better fit the magnetic core if the successive windings are started at the side (left or right) at which the previous layer ended. This is contrary to the instruction on the CODED-T printout.

When it is feasible, wind each layer of the winding or coil out to the minimum acceptable margin. In some cases, the last layer of a winding may be unfilled. Sometimes it is desirable to "pyramid" the layers, i.e., one or two turns less on each successive layer while achieving the correct total turns and adhering to the margin requirements. Do not increase the number of layers to do this. In summary, choose the sequence of the windings and the turns per layer so that the margins increase with each successive layer. Do not violate a more important requirement.

As a general rule, put half of each winding on each leg of a single-phase C-core. This is called split coil construction and will give approximately minimum volume, weight, resistance of windings, external magnetic flux, and internal coil temperature.

On transformers with two secondaries, put one secondary on first, then the primary, and finally the other secondary. This will give better coupling, that is lower reactance drop in each secondary, than with the primary first or last.

On transformers with three or more secondaries, the more important windings should be next to the primary.

Generally each secondary must produce a specified minimum voltage under low line voltage and maximum load current conditions even if this means a higher than desirable secondary voltage when line voltage is high and load current is low. Those secondaries that are not close enough to the primary to have good coupling must produce a little extra voltage to compensate for the reactance drop caused by the poorer coupling. This may mean an extra turn, or turns, or at least the extra fractional turn over the theoretical number required.

As a general rule, core proportions for chokes and transformers using C-cores should approximately follow the "Rule of 2s". That is:

$D \cong 2 \times E$

$G \cong 2 \times F$

$G \times F \cong 2 \times D \times E$

These proportions approximately minimize the weight and volume of the completed component (excluding mounting brackets, encapsulant, etc.) Mathematically, the ideal ratios are in the range of about 2:1 to 2.2:1. The ratios are partially dependent upon the window fill factor (the fraction of the window area actually occupied by the conductors). The above ratios assume a fill factor of about 0.4 and copper conductors.

High voltage components require more insulation than medium or low voltage units and therefore a greater ratio of GF to DE and thus will have a smaller fill factor than 0.4.

When possible, fill the window nearly completely, leaving only enough space to accommodate tolerances, bulge, etc. Bulge should usually be 5% to 10% of the total of conductor diameters (max), insulation thicknesses, and coil form wall thickness.

If window space left over is greater than about 25%, use a more suitable core or larger wire. Unused window space increases the leakage reactance and thus degrades the regulation of a transformer.

When attempting to design a transformer or inductor to fit a limited space, several compromises with the wire gage are possible. Instead of the wire gage indicated by the RMS current, two bifilar strands (to be connected in parallel) that are each three sizes smaller than the original gage may be used. These techniques have several advantages: better coupling due to the longer layers, reduced buildup which may permit more layers, use of a smaller margin, and less winding bulge.

7.2 Fabrication of Power Line Frequency Transformers and Inductors

a. Cut coil forms.

b. Wind coils on coil forms using specified margins and appropriate interwinding and interlayer winding insulation.

c. Assemble coils on a silicon iron C-core incorporating any nonmagnetic, electrically nonconducting spacers for appropriate air gap if needed.

d. Assemble electrical terminals, e.g., using fiber glass board.

e. Assemble steel band around C-core and mounting base.

f. Perform basic electrical tests as discussed in Section 8.

g. Complete impregnation and incapsulation.

h. Repeat tests described in step f.

7.3 Rules of Thumb for Pot Core Magnetics

From the standpoint of maximum number of turns on a bobbin, it is usually advantageous to stagger the starts and finishes of a bifilar winding. The two strands of wire start at slots that are diametrically located and finish in the same slot (either right or left side) for an integral number of turns in each strand.

Bobbin windings cannot easily be wound with margins and do not need them because of the flanges. The effective wire diameter should be taken as equal to the maximum diameter. Ordinarily no allowance need be made for the fact that the winding is a helix and the effective diameter is minutely larger than the actual diameter.

As mentioned under Section 7.1 Rules of Thumb for C-core Magnetics, multifilar windings may provide better space utilization and better magnetic coupling.

7.4 Fabrication of Pot Core Magnetics

The steps for pot core magnetics fabrication are described below:

a. Wind coils on bobbin applying interlayer winding insulation if required.

b. Assemble winding bobbin, pot core, and nonmagnetic, nonconducting spacer as an air gap, if required.

c. Assemble mounting hardware using the appropriate torque as shown in Table 5.

d. Perform basic electrical test on transformer as outlined in Section 8.

e. Apply RTV to the slots in the core where the transformer winding leads exit.

TABLE 5. RECOMMENDED MOUNTING HARDWARE AND TORQUE FOR VARIOUS POT CORE SIZES

CORE	METAL WASHER ID (INCH)	METAL WASHER OD (INCH)	MOUNTING SCREW	P.D. of SCREW (INCH)	TORQUE OZ-IN	MICA WASHER ID (INCH)	MICA WASHER OD (INCH)
905	.067	.156	0-80	.0519	2.7	.065	.150
1107	.078	.1835	1-64	.0629	4	.075	.180
1408	.121	.248	4-40	.0958	7	.120	.230
1811	.121	.248	4-40	.0958	7	.120	.290
2213	.160	.310	6-32	.1177	10	.170	.360
2616	.186	.373	8-32	.1437	13	.170	.440
3019	.219	.437	10-32	.1697	15	.210	.520
3622	.219	.437	10-32	.1697	15	.210	.625
4229	.219	.437	10-32	.1697	15	.210	.685

8.0 TRANSFORMER AND INDUCTOR TESTING

Before connecting a transformer or inductor into a circuit, the construction should be checked. The core should be checked for correct orientation and damage. Measurement of the inductance of each winding serves as a check of the turns ratio and the continuity of the windings. The turns ratio equals the square root of the inductance ratio.

$$N2/N1 = (L2/L1)^{.5}$$

Equipment for transformer and inductor testing must be chosen carefully. Meters which respond to and read true RMS should be used for all voltage and current measurements. Winding resistance measurements can be affected by frequency and distributed capacitance and require an instrument which applies DC to the windings. Digital impedance meters use low AC voltage or current for resistance, capacitance, and inductance measurements resulting in impedance, rather than resistance measurements, if the coil is on the core.

When testing high frequency (20 kHz and above) magnetics, additional considerations are required. Special test equipment can be used to duplicate the application. It is also possible to make measurements on acceptable components with conventional instruments. The results are used to develop acceptance tests.

8.1 Transformers

Voltages and currents should be measured with the transformer loaded with the actual circuit to be used. All secondary windings should have a rectifier circuit, filter circuit and load in place. Testing may reveal connection errors or a connection where there should be isolation.

Primary exciting current and no load voltage should be measured at maximum line voltage and nominal frequency. The exciting current is difficult to accurately predict and varies among units. On a three-phase transformer the exciting current may vary from leg to leg. The total exciting current for a three-phase transformer could be about 25% higher than in a single-phase transformer of equal power.

To make measurements using a 1000 hertz signal on a transformer intended to operate at 50 kHz requires 1/50th the normal operating voltage on the primary in order to achieve the same flux density in the core. A large increase in flux density will cause disproportional increases in core loss and exciting current.

8.2 Inductors

In any magnetic device having a core of magnetic material, the permeability of the core, and thus the inductance, is dependent upon the magnetic flux density in the core. In addition to carrying AC current, both power line frequency and high frequency inductors will carry DC current which affects core permeability. Therefore, tests should be performed with an instrument which measures inductance while the inductor is carrying the specified DC and AC currents at the specified frequency. Designing in 10% additional inductance increases the chances of meeting the AC and DC specifications.

Ideally, inductance should be measured at three or more current levels: maximum, minimum, and a typical value. The test signal should be at the frequency and voltage corresponding to the application.

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APPENDIX A

DERIVATIONS

A.1 Standard Magnetic Design Expressions

Symbols

Erms	primary voltage in Vrms
B	flux density in gauss
Ac	core gross cross-sectional area in in^2
Ai	core net cross-sectional area in in^2
Kc	core stacking factor, $Kc = \frac{Ai}{Ac}$
N	number of turns
f	frequency of applied waveform in hertz
E	peak voltage in volts
t_{on}	on-time in seconds; equals duty cycle x 1/f
L	inductance in henries
μ	permeability in gauss/oersted
H	magnetizing force in oersteds
ℓg	length of air gap in cm
ℓc	length of magnetic path in cm
$\mu\Delta$	incremental permeability in gauss/oersted
E_p	primary voltage in volts
E_s	secondary voltage in volts
N_p	primary turns
N_s	secondary turns
I_p	primary current in amps
I_s	secondary current in amps
Z	impedance in ohms; equals square root of the sum of resistance squared plus reactance squared
V_{NL}	no-load voltage in volts
V_{FL}	full load voltage in volts
T	period in $seconds^{-1}$
E_{H2}	peak amplitude of voltage at 2nd harmonic of supply frequency in rectified, unfiltered waveform in volts

Edc	direct or average output voltage in volts
R_L	load resistance in ohms
Lc	critical inductance in henries
ω_{H2}	angular frequency of second harmonic of supply frequency in radians
Idc	direct current in amps
$\hat{I}$	peak current in amps
e	instantaneous inductor voltage in volts
I	instantaneous current in amps
t	time in seconds

$$Erms = \pi\sqrt{2}\,B \times Ai \times N \times f \times 10^{-8} \quad \text{for sine wave.} \tag{1}$$

$$E = 4B \times Ai \times N \times f \times 10^{-8} \quad \text{for square wave.} \tag{2}$$

$$E \times t_{on} = 2B \times Ai \times N \times 10^{-8} \quad \text{for square or rectangle wave.} \tag{3}$$

$$L = \frac{.4\pi \times Ai \times N^2 \times 10^{-8}}{\ell g + \frac{\ell c}{\mu \Delta}} \quad \text{henries} \tag{4}$$

$$H = .4\pi\, N \times \hat{I} \text{ oersteds} \tag{5}$$

$$B = \mu H \text{ gauss} \tag{6}$$

$$B = \frac{.4\pi\, N \times \hat{I}}{\ell g + \frac{\ell c}{\mu \Delta}} \text{ gauss} \tag{7}$$

$$\frac{E_p}{E_s} = \frac{N_p}{N_s} = \frac{I_s}{I_p} \tag{8}$$

$$\frac{Z_1}{Z_2} = \frac{(E_1)^2}{(E_2)^2} = \frac{(N_1)^2}{(N_2)^2} \tag{9}$$

$$\text{Percent regulation} = 100\,\frac{(V_{NL} - V_{FL})}{(V_{FL})} \tag{10}$$

$$Irms = \left(\frac{1}{T}\int_0^T I^2 dt\right)^{1/2} \text{ Arms} \tag{11}$$

$$Ai \times N = \frac{L\hat{I} \times 10^8}{B} \text{ centimeters squared X turns} \tag{12}$$

The above is the basic equation for design of chokes for power supplies. It is not applicable to charging choke design.

A.2 Derivation of Expression for Critical Inductance of Inductor in an LC Filter for a Rectified Sinusoidal Waveform and a Rectangular Waveform.

Depicted in Figure A.1 is the current waveform for the case of critical inductance if the second harmonic only of the supply were present. This is never true, but it is an appropriate assumption for a simplified analysis.

In practice at least two times the calculated critical value should be used because the current waveform extends farther below the DC line than above it. At two times critical inductance the waveform approximates the second harmonic of the supply frequency.

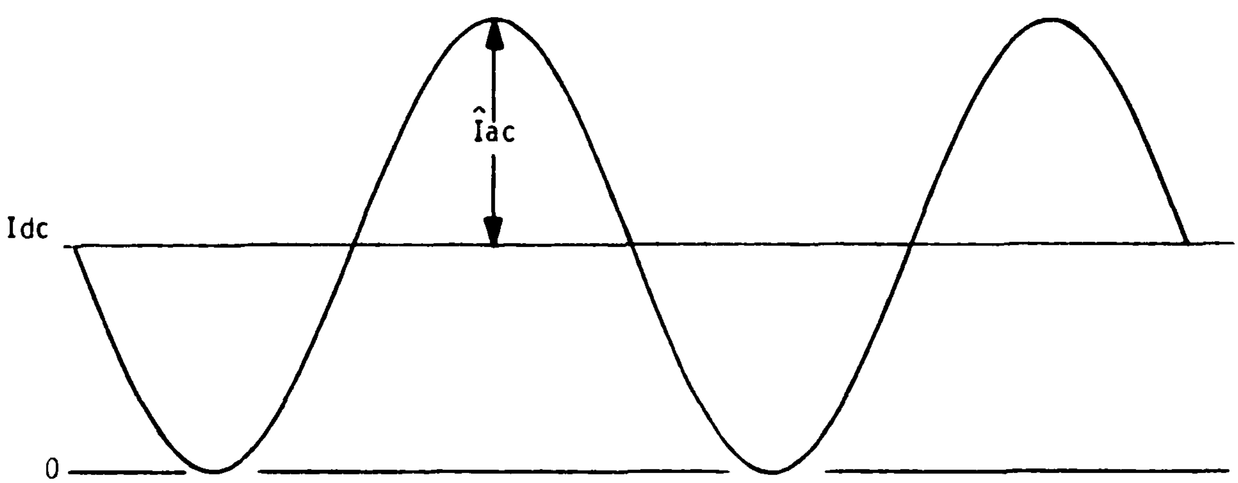


Figure A.1 Relationship of Idc and $\hat{I}ac$

From Figure A.1 above:

$\hat{I}ac$ = Idc for critical inductance

$$\text{i.e.,} \quad \frac{E_{H2}}{L\omega_{H2}} = \frac{Edc}{R_L} \tag{13}$$

The equation for the AcAw product, which is applicable to single-phase and three-phase power transformers built on laminated cores, is given by,

Gross AcAw product in $(\text{inches})^4$

$$= \frac{(25)(EI)(K_R)(4.205)(1 + .00427T_A)}{(\sqrt{2})(2.54^2)BKcKwf}\left\{40 \; Log_{10}\left[(EI)\left(\frac{400}{f}\right)^{0.8}\right] - \Delta T + 52\right\} \qquad (49)$$

Where:

$$EI = \frac{Pout}{\eta \cos \emptyset}$$

Pout = Power output of the transformer, not of the rectifier or other circuitry which may follow.

η = Efficiency.

$\emptyset$ = Phase angle of the input.

T_A = Ambient temperature in ^oC.

ΔT = Temperature rise in ^oC.

Kc = Stacking factor of core
0.89 for 2-mil material
0.90 for 4-mil material
0.95 for 12-mil material

Kw = Window fill factor (typically in the range of 0.2 for small units, 0.3 for medium size units, up to 0.4 for large units).

K_R = 2 for pure resistive load, or $1 + \sqrt{2}$ for center-tapped secondary and resistive load, or $2\sqrt{2}$ if primary is center-tapped drive and secondary is for center-tapped rectifier circuit.

B = Peak flux density in gauss.

f = Minimum frequency of supply in hertz.

Aw = Window area in square inches.

Ac = Gross cross-sectional area of one leg of the core in square inches

Ai = Core net cross-sectional area in in^2; K_c = Ai/Ac

N = Number of turns in primary winding.

E = RMS voltage of the source supplying the primary.

I = RMS current in the primary as reflected from the secondary.

Note: If the secondary load circuit is a rectifier combination with a filter circuit, appropriate calculations must be made to find the secondary RMS current

CM = Cross-sectional area of conductor in circular mils.
= (diameter in mils)^2

1CM = Area of 1 mil diameter circle

$$= \pi\left(\frac{.001}{2}\right)^2 = \frac{\pi}{4} \times 10^{-6}\ in^2 = 0.7854 \times 10^{-6}\ in^2$$

1 mil= 0.001 inch

CM/A = Circular mils per ampere.

$$E = (\pi\sqrt{2})(2.54^2)B \times Ac \times Kc \times N \times f \times 10^{-8}$$

from which

$$\text{gross } Ac = \frac{E \times 10^8}{(\pi\sqrt{2})(2.54^2)B \times Kc \times N \times f} \quad (50)$$

$$\text{gross } Aw = \frac{CM \times N \times K_R}{Kw \times 10} \times \frac{\pi}{4} \quad (51)$$

$$\text{gross } AcAw = \frac{E \times 10^8}{(\pi\sqrt{2})(2.54^2)B \times Kc \times N \times f} \times \frac{CM \times N \times K_R}{Kw \times 10^6} \times \frac{\pi}{4} \times \frac{I}{I}$$

$$= \frac{25\ E \times I \times K_R}{(\sqrt{2})(2.54^2)B \times Kc \times Kw \times f} \times CM/A \quad (52)$$

The CODED-T program uses the following expression to estimate the size wire required for each winding.

$$CM/A = 6\left[40\ \mathrm{Log}_{10}\ (EI) - \Delta T + 51\ \frac{2}{3}\right] \quad (53)$$

For use here some refinements are made to reduce iterations required when designing by manual computation.

Equation (53) did not take into account the fact that, at a frequency other than 400 hertz, the transformer would be larger for lower frequencies or smaller for higher frequencies. To account for other frequencies, the volt-ampere product, EI, is multiplied by a factor of $(400/f)^{.8}$.

$$\text{or } Lc = \frac{E_{H2}}{Edc} \times \frac{R_L}{\omega_{H2}} \tag{14}$$

$$\frac{E_{H2}}{Edc} = \frac{2}{3} \quad \text{for 2-pulse system.} \tag{15}$$

$$\frac{E_{H2}}{Edc} = \frac{2}{35} \quad \text{for 6-pulse system.} \tag{16}$$

However, an inductor is seldom used in a 6-pulse or greater system.

$$Lc = \frac{E_{H2}}{E_{dc}} \times \frac{R_L}{\omega_{H2}} \tag{14}$$

$$= \frac{2}{3} \times \frac{R_L}{\omega_{H2}} \quad \text{for 2-pulse per cycle system}$$

$$R_L = \frac{Edc}{Idc} \tag{17}$$

$$Lc = \frac{2}{3} \times \frac{1}{\omega_{H2}} \times \frac{Edc}{Idc} \tag{18}$$

Derivation of Critical Inductance for a Rectangular Waveform.

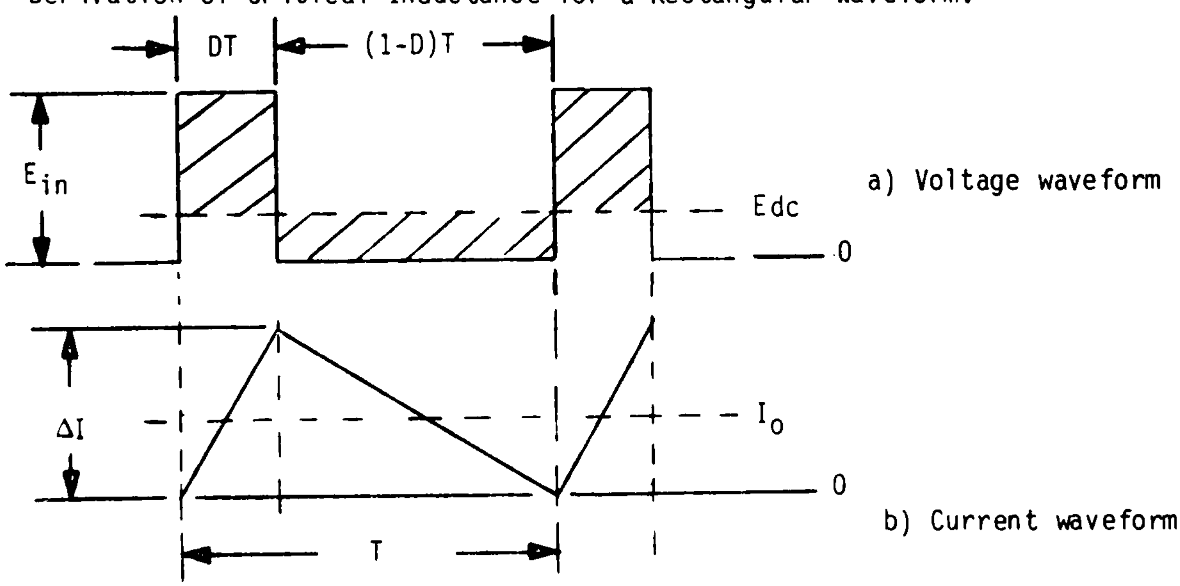


Figure A.2 Inductor Voltage and Current Waveform.

The voltage, e, across the inductor is given by,

$$e = L \frac{di}{dt} \tag{19}$$

where L = inductance
i = instantaneous current
t = time

This inductor voltage, e, is the difference between the input voltage, Ein, and the dc output voltage, Edc. From Figure A.2 a), the inductor voltage is,

$$e = Ein - Edc \tag{20}$$

during the on time D x T, where D is the duty factor and T is the period.

The slope of the current waveform, Figure A.2 b), during the on time is,

$$\frac{di}{dt} = \frac{\Delta I}{DT} \tag{21}$$

Using Equations (20) and (21) in Equation (19) gives,

$$Ein - Edc = L \frac{\Delta I}{DT} \tag{22}$$

solving for L gives,

$$L = \frac{(Ein - Edc)DT}{\Delta I} \tag{23}$$

Since the DC voltage, Edc, is the average voltage, then from Figure A.2 a) the cross-hatched areas above and below Edc must be equal. Thus,

$$(Ein - Edc)\ DT = Edc\ (1-D)T \tag{24}$$

Inserting Equation (24) into Equation (23) gives,

$$L = \frac{(1-D)T\ Edc}{\Delta I} \tag{25}$$

Equation (25) is a general equation for the waveforms in Figure A.2. By definition of critical inductance, the minimum inductance required for continuous current occurs at, for constant Edc and T, minimum current and minimum duty factor.

Thus,

$$\frac{\Delta I}{2} = Idc,min \tag{26}$$

or

$$\Delta I = 2Idc,min \tag{27}$$

where Idc, min = minimum DC current.

Using Equation (27) and the minimum duty factor, Dmin, in Equation (25), the critical inductance, Lc, is found from,

$$Lc = \frac{(1 - Dmin)T\ Edc}{2Idc,min} \tag{28}$$

A.3 Relationship of Idc, $\hat{I}$, Irms, and Inductance of Choke in Filter Circuit of a Pulsed Width Modulated Power Supply.

The current in the inductor is shown in the figure below,

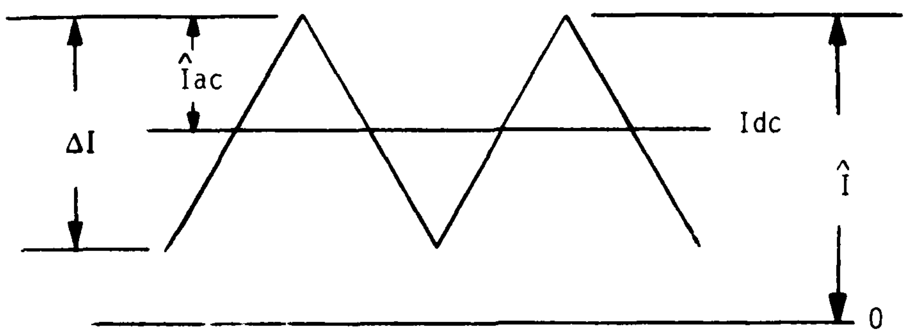


Figure A.3 Current Waveform for a Pulse Width Modulated Power Supply Output Inductor

The peak inductor current, $\hat{I}$, is given by,

$$\hat{I} = Idc + \hat{I}ac \tag{29}$$

or

$$\hat{I} = Idc + \frac{\Delta I}{2} \tag{30}$$

The maximum ΔI can be found from Equation (25), which is re-arranged as,

$$\Delta I = \frac{(1 - D)T\ Edc}{L} \tag{31}$$

or for maximum ΔI,

$$\Delta Imax = \frac{(1 - Dmin)T\ Edc}{L} \tag{32}$$

where Dmin = minimum duty factor.

Another expression for ΔI_{max} is found by solving Equation (28) for $2I_{dc,min}$ and dividing the result by Equation (32), which results in,

$$\frac{2I_{dc,min}}{\Delta I_{max}} = \frac{L}{L_c} \tag{33}$$

Defining $K = \dfrac{L}{L_c}$ gives,

$$\frac{2I_{dc,min}}{\Delta I_{max}} = \frac{L}{L_c} = K \tag{34}$$

or

$$\Delta I_{max} = \frac{2I_{dc,min}}{K} \tag{35}$$

or

$$\Delta I_{max} = 2I_{dc,min}\left(\frac{L_c}{L}\right) \tag{36}$$

The equation for the maximum peak inductor current, $\hat{I}_{max}$, is

$$\hat{I}_{max} = I_{dc,max} + \frac{\Delta I_{max}}{2} \tag{37}$$

where $I_{dc,max}$ = maximum inductor DC current

and ΔI_{max} is found from either Equation (32), Equation (35), or Equation (36).

The inductor RMS current, Irms, is given by

$$\text{Irms} = \left[(\text{Iac,rms})^2 + (\text{Idc})^2\right]^{1/2} \tag{38}$$

where Iac,rms = RMS value of the ripple current

$$\text{Iac,rms} = \frac{\hat{\text{I}}\text{ac}}{\sqrt{3}} \quad \text{for the sawtooth waveform} \tag{39}$$

Since $\hat{\text{I}}\text{ac} = \frac{\Delta \text{I}}{2}$,

then the maximum RMS current, Irms,max, is,

$$\text{Irms,max} = \left[\left(\frac{\Delta\text{Imax}}{2\sqrt{3}}\right)^2 + \left(\text{Idc,max}\right)^2\right]^{1/2} \tag{40}$$

where ΔImax is found again by either Equation (32), Equation (35) or Equation (36).

A.4 Application of Inductor Design Equations to the Design of Coupled Inductors

The fundamental equation for a single inductor is given by Equation (4). By rearranging Equation (7), and modifying the N x $\hat{\text{I}}$ product to include the ampere-turns ($N_1\hat{I}_1$, $N_2\hat{I}_2$, $N_3\hat{I}_3$, etc.) of the additional windings, gives

$$\ell g + \frac{\ell c}{\mu \Delta} = \frac{.4\pi}{B} \quad N_1\hat{I}_1 + N_2\hat{I}_2 + N_3\hat{I}_3 + \text{etc.} \tag{41}$$

$$L_1 = \frac{.4\pi \times Ai \times N_1^2 \times 10^{-8}}{\frac{.4\pi}{B} \quad N_1\hat{I}_1 + N_2\hat{I}_2 + N_3\hat{I}_3 + \text{etc.}} \tag{42}$$

$$L1 = \frac{B \times Ai \times N_1^2 \times 10^{-8}}{N_1\hat{I}1 + N_2\hat{I}2 + N_3\hat{I}3 + \text{etc.}} \tag{43}$$

$$L_1 = \frac{B \times Ai \times N_1 \times 10^{-8}}{\hat{I}_1 + \frac{N_2}{N_1}\hat{I}_2 + \frac{N_3}{N_1}\hat{I}_3 + \text{etc.}} \tag{44}$$

Equation (41) and either Equation (43) or Equation (44) may be used for coupled inductor design.

The critical inductance, L_1c, for inductor 1, is

$$L_1c = \frac{E_{01}\, t_{off}}{2 \quad I_{01} + I_{02}\frac{N_2}{N_1} + I_{03}\frac{N_3}{N_1} + \text{etc.} \quad \text{min}} \tag{45}$$

where I_{0n} = Idc,min for winding n.

As can be seen by examination of the denominator, the critical inductance for outputs having the same number of secondary turns is the same value even though their currents differ.

Since the choke windings for outputs having equal secondary turns, must have equal turns, they will have equal inductance and therefore equal values of

$$K = \frac{L}{Lc} \tag{46}$$

Maximum peak current, $\hat{I}n,max$, through each winding, n, will be

$$\hat{I}n,max = Indc,max + \frac{\Delta In,max}{2} \tag{47}$$

where Indc,max = maximum DC current through winding n

ΔIn,max = maximum ΔI current through winding n from Equation (32), Equation (35), or Equation (36).

The peak currents, $\hat{I}n,max$, from Equation (47) are used in the expression

$$L_1 = \frac{B \times Ai \times N_1 \times 10^{-8}}{\hat{I}_{1,max} + \frac{N_2}{N_1}\hat{I}_{2,max} + \frac{N_3}{N_1}\hat{I}_{3,max} + \quad etc.} \tag{48}$$

A.5 An Empirical Relationship for Calculating the AcAw Product and the Circular Mils per Ampere for a Transformer used at a Primary Power Frequency.

The empirical portion of the equation is based upon an equation used in the NAC CODED-T program. The original source of the equation is unknown; it may have been published in the Armour report on transformer design [ref. A-1]. As given below it has some refinements to make it more useful for choosing a core and the required copper/current density figure.

The CODED-T program does not need a refined equation because temperature rise and the effect of other constraints is calculated for each tentative design. If specifications are not met, iteration proceeds until they are met.

Multiplying the unmodified expression for CM/A by

$$\frac{1 + .00427T_A}{1.427} \tag{54}$$

makes the expression applicable when the ambient temperature = T_A instead of $+100^{\circ}C$. The factor is derived from the expression for change in resistance of copper with temperature.

$$\frac{R_2}{R_1} = \frac{1 + .00427t_2}{1 + .00427t_1} \tag{55}$$

where t_1 and t_2 are temperatures in $^{\circ}C$.

Thus,

$$CM/A = (4.205)(1 + .00427T_A)\left\{40 \operatorname{Log}_{10}\left[(EI)\left(\frac{400}{f}\right)^{0.8}\right] - \Delta T + 52\right\} \tag{56}$$

The circular mil per ampere relationship is applicable to single-phase transformers on E-I punched laminations or on C-cores, both single coil and split coil types.

The CM/A and AcAw product equations are applicable to three-phase transformers by calculating for 2/3 the total power required in the three-phase transformer.

A.6 Derivation of AcAw Product for an Inductor used at a Primary Power Frequency

Solving Equation (7) for $\ell g + \ell c/\mu\Delta$ and converting to inches gives

$$\ell g + \frac{\ell c}{\mu\Delta} = \frac{.4\pi N \times \hat{I}}{B \times 2.54} \tag{57}$$

Substituting into Equation (4)

$$L = \frac{B \times Ai \times N \times 10^{-8} \times (2.54)^2}{\hat{I}} \tag{58}$$

Solving Equation (58) for N and multiplying by CM gives

$$CM \times N = \frac{L \times \hat{I} \times 10^{8} \times (CM/A) \times Irms}{B \times Ai \times (2.54)^2} \tag{59}$$

Since Kc = Ai/Ac, or Ai = AcKc then,

$$CM \times N = \frac{L \times \hat{I} \times 10^{8} \times (CM/A) \times Irms}{B \times Ac \times Kc \times (2.54)^2} \tag{60}$$

or

$$Ac \times CM \times N = \frac{L \times \hat{I} \times 10^{8} \times (CM/A) \times Irms}{B \times Kc \times (2.54)^2} \tag{61}$$

Since,

$$Aw = \frac{\pi \times CM \times N}{4Kw \times 10^{6}} \tag{62}$$

then

$$Ac\,Aw = \frac{\pi \times L \times \hat{I} \times (CM/A) \times Irms \times 10^{2}}{4B \times Kc \times Kw \times (2.54)^2} \tag{63}$$

Letting $\hat{I} = Idc + \hat{I}ac$

$$AcAw = \frac{\pi \times L \times (Idc + \hat{I}ac) \times Irms \times (CM/A) \times 10^2}{4B \times Kc \times Kw \times (2.54)^2} \tag{64}$$

A.7 Determination of Volts/mil Rating

Volts per mil rating of insulation is approximately inversely proportional to the square root of the thickness ratio.

Thus,

$$\frac{V_1/mil}{V_2/mil} = \left(\frac{d_2}{d_1}\right)^{1/2} \tag{65}$$

where: d_1 is the thickness of the material having the rating of V_1 volts per mil

d_2 is the thickness of the material having the rating of V_2 volts per mil

From which can be derived:

$$V_1 = (V_2/mil)(d_1 d_2)^{1/2} \tag{66}$$

or,

$$d_1 = \frac{1}{d_2}\left(\frac{V_1}{V_2/mil}\right)^2 \tag{67}$$

A.8 Pot Core Power Capability

The power handling capability of a pot core, for a square wave, can be estimated from the equation

$$Po = \frac{4\eta \times B \times f \times Kw \times Ac \times Aw \times 10^{[illegible]}}{CM/A} \quad \text{Ref A-[illegible]} \tag{68}$$

Where Po = transformer output power in watts

n = transformer efficiency

B = flux density in gauss

f = frequency in Hz

Kw = winding fill factor

Ae = core effective area [illegible], Table 4

Aw = window area [illegible]

CM/A = current [illegible] per ampere

Multiplying Equation [illegible]

[illegible]

[illegible]

[illegible]

TABLE A.1. Table of AeAw Product For Pot Cores

POT CORE NUMBER	EFFECTIVE AREA Ae (cm^2)	WINDING AREA Aw (cm^2)	AeAw (cm^4)
905	.101	.0317	0.003202
[illegible]	[illegible]	.0478	0.007983
[illegible]	[illegible]	.0881	0.02211
[illegible]	[illegible]	[illegible]	0.07404
[illegible]	[illegible]	[illegible]	0.1664
[illegible]	[illegible]	[illegible]	0.3555
[illegible]	[illegible]	[illegible]	0.7411
[illegible]	[illegible]	[illegible]	1.4403
[illegible]	[illegible]	[illegible]	[illegible]

REFERENCES

A-1 "Research and Development of a New Design Method for Power Transformers", Armour Research Foundation of Illinois Institute of Technology, ARF Project E037, Final Report, 1956.

A-2 Ferrite Cores for Power and Filter Applications, Magnetics Inc., FC-405-12S, 1985.

APPENDIX B

RECTIFIER CIRCUITS

Rectifier Circuits

This appendix contains a brief overview of common rectifier circuits. These circuits convert AC sine waves into DC voltage. For each, the relationships between RMS currents at various points of the circuit are given.

PPC = pulses per cycle is given for each case

TCA = total conduction angle = $\theta_1 + \theta_2$ in radians

$$\text{where } \theta_2 = \tan^{-1}\left(\frac{1}{\omega RC}\right)$$

$$\text{and } \theta_1 = \left(\frac{2\pi}{PPC}\right) - \theta_2 + (\omega RC)\ln\left(\frac{\cos\theta_1}{\cos\theta_2}\right)$$

ω = frequency of input sine wave in radians per second

R = load resistance in ohms

C = filter capacitance in farads

θ_1 must be solved interatively and is best calculated by computer.

Unless otherwise noted, the following equations apply for each circuit:

$$I_{rms} = I_{dc}\sqrt{\frac{4}{3}\,\frac{\left(\frac{2\pi}{PPC}\right)}{TCA}}$$

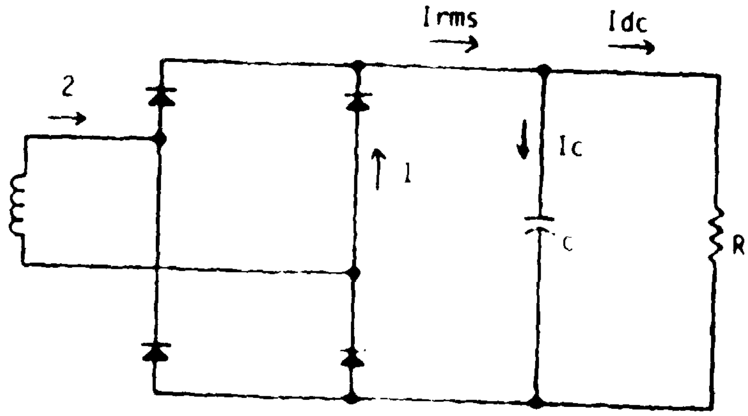
$$\hat{I} = 2I_{dc}\,\frac{\left(\frac{2\pi}{PPC}\right)}{TCA}$$

Where I_{dc} = load DC current

I_{rms} = RMS current at rectifier circuit output

$\hat{I}$ = peak AC current at rectifier circuit output

Figures B.1 through B.8 contain schematics and the equations for some commonly used rectifier circuits. For more information on rectifier circuits, see Reference B-1.

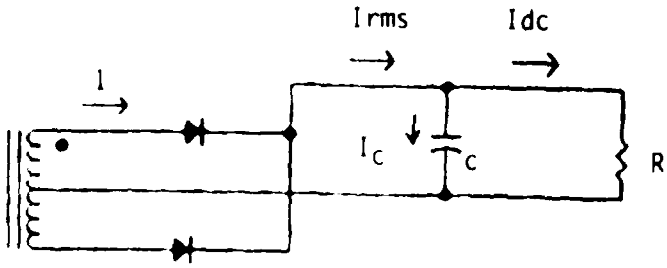


PPC = 2

$$I_1 = \frac{I_{rms}}{\sqrt{2}}$$

$$I_2 = I_{rms}$$

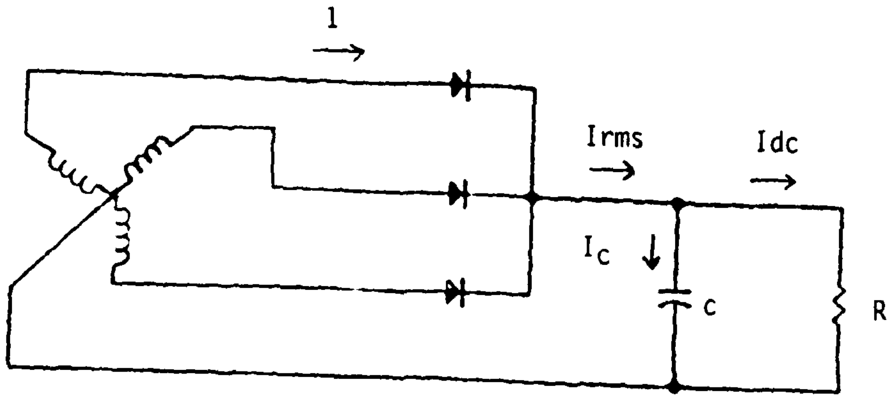
[illegible]



PPC = 2

$I = \frac{Irms}{\sqrt{2}}$

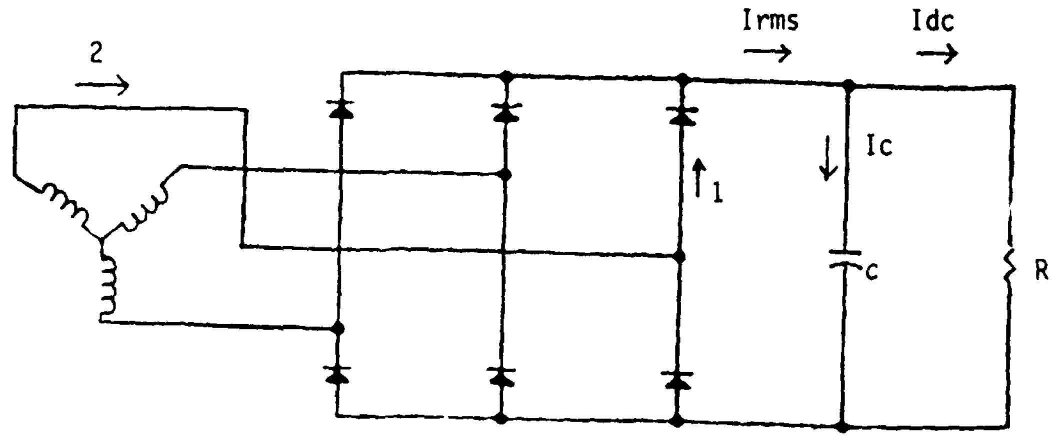
Figure ... Single-Phase Full-wave Center-Tapped



PPC = 3

$$I_1 = \frac{1}{\sqrt{3}} \text{ Irms}$$

Figure B.3. Three-Phase Half Wave

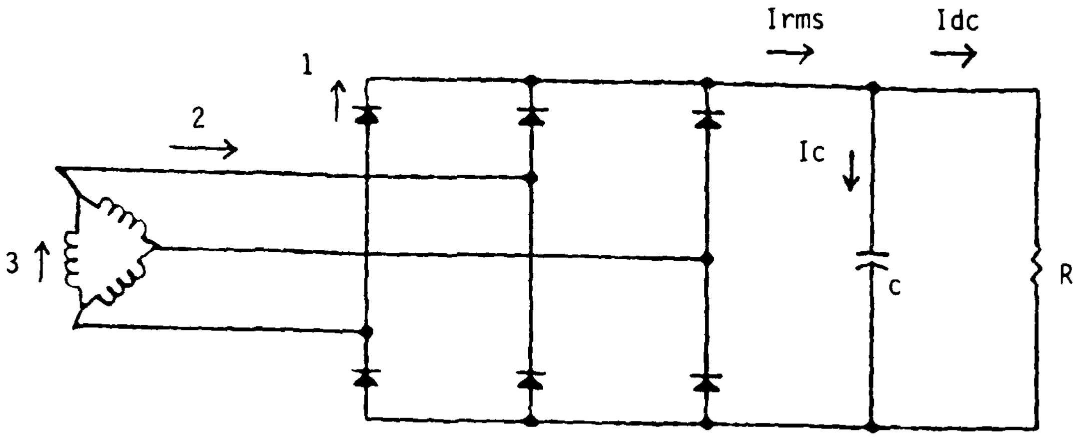


PPC = 6

$$I_1 = \frac{1}{\sqrt{3}}\,Irms$$

$$I_2 = \sqrt{\frac{2}{3}}\,Irms$$

Figure B.4. Three-Phase Full Wave



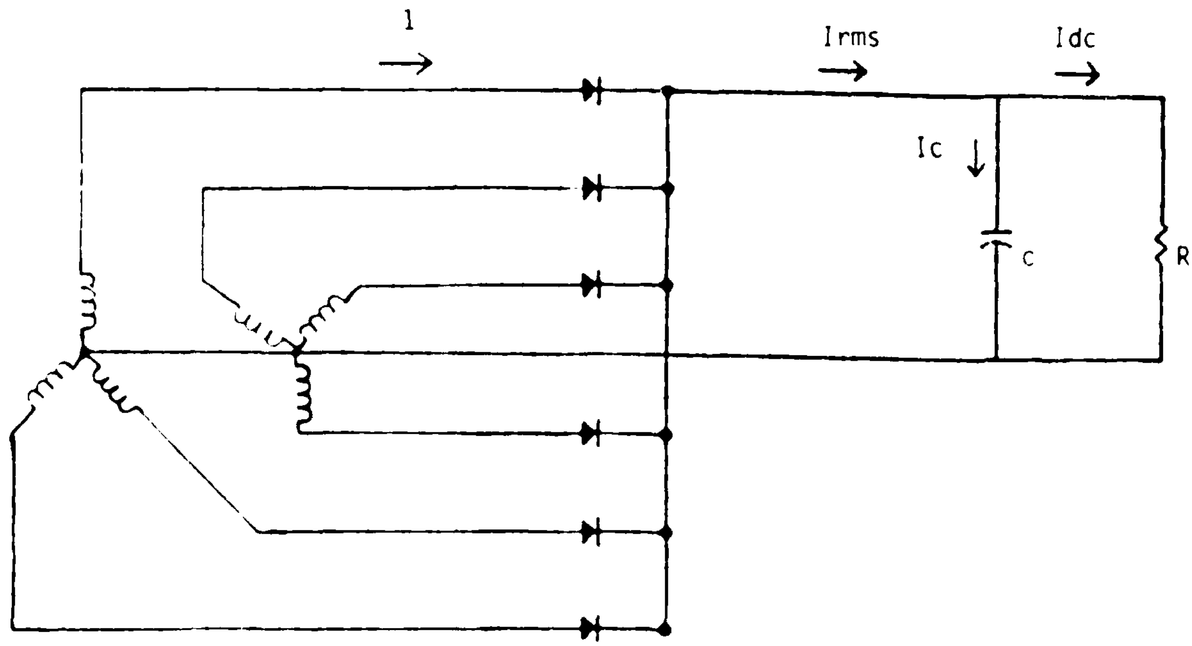
PPC = 6

$$I_1 = \frac{1}{\sqrt{3}} \text{ Irms}$$

$$I_2 = \sqrt{\frac{2}{3}} \text{ Irms}$$

$$I_3 = \sqrt{\frac{2}{3}} \text{ Irms}$$

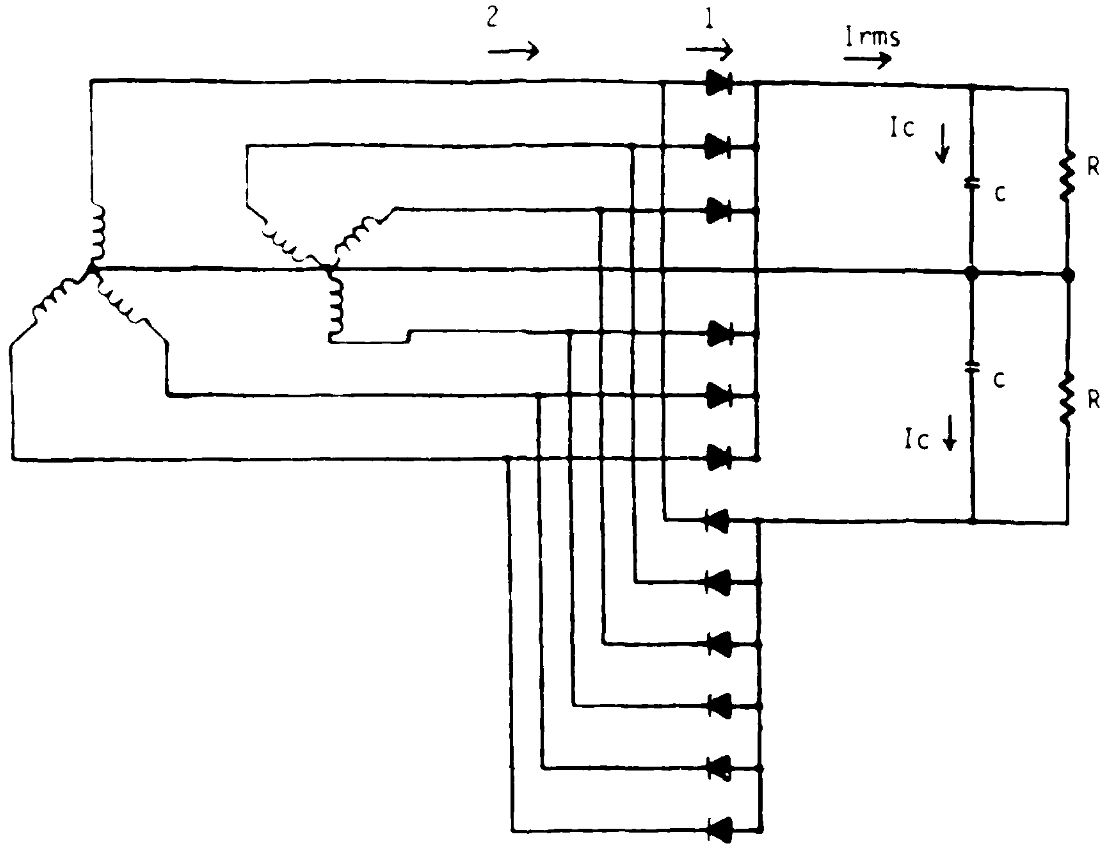
Figure B.5. Three-Phase Full Wave



PPC = 6

$$I_1 = \frac{1}{\sqrt{6}}\,\mathrm{Irms}$$

Figure B.6 Three-Phase Full-wave

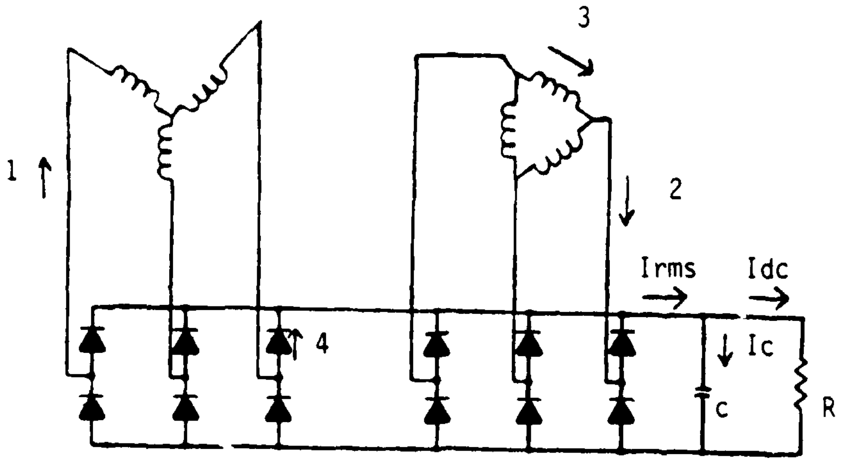


PPC = 6

$$I_1 = \frac{1}{\sqrt{6}} \text{ Irms}$$

$$I_2 = \frac{1}{\sqrt{3}} \text{ Irms}$$

Figure B.7. Three-Phase Full Wave, Positive and Negative



PPC = 12

$$Irms = \frac{Idc}{3}\sqrt{\frac{2\pi}{TCA}}$$

$$\hat{I} = Idc \; \frac{\pi}{3}\left(\frac{1}{TCA}\right)$$

$$I_1 = I_2 = \frac{1}{\sqrt{6}}\, Irms$$

$$I_3 = \frac{1}{\sqrt{18}}\, Irms$$

$$I_4 = \frac{1}{\sqrt{[illegible]}}\, Irms$$

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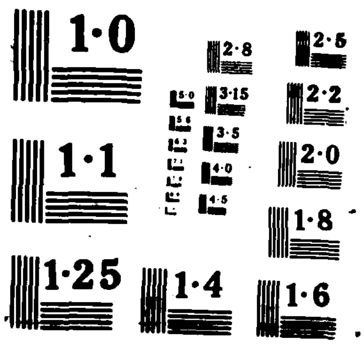
1·0
2·8
2·5
3·15
2·2
3·5
1·1
2·0
4·0
4·5
1·8
1·25
1·4
1·6

REFERENCES

B.1 Schaefer, Johannes. Rectifier Circuits, Theory and Design. New York, John Wiley and Sons, 1965.

APPENDIX C

DESIGN EXAMPLES

C.1 Design of a 60 Hz Primary Power Transformer Using the Manual Approach

In this example a 60 Hz primary power transformer is designed using the manual design approach outlined in section 5.1.1. This example assumes that the transformer electrical performance specifications have been determined and are as follows:

Primary: frequency = single-phase, 60Hz $\pm$ 5%
voltage = 115 Vrms $\pm$ 10%

Secondary: voltage = 10.45 Vrms maximum, center-tapped
current = 10 Arms maximum (at rectifier output)

A schematic of the transformer/rectifier circuit is shown below.

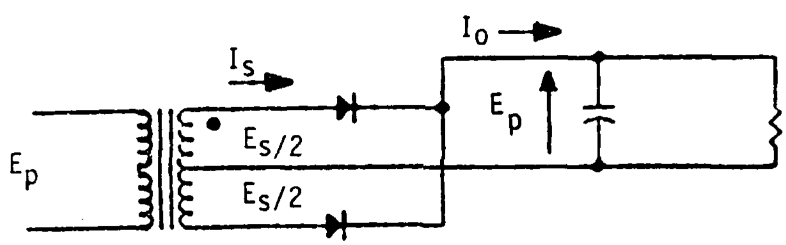

Figure C.1 Center-Tapped Transformer/Rectifier Schematic

From Appendix B, Figure B.2, the secondary current is found from,

$I_s = I_o/\sqrt{2}$

where I_s = secondary rms current

I_o = center-tap rectifier output rms current

Thus,

$I_s = 10/\sqrt{2} = 7.1$ Arms

The secondary voltage is simply twice the rectifier output voltage, thus,

$E_s = 2 \times (E_p/\sqrt{2})$

where E_s = secondary rms voltage

E_p = peak rectifier output voltage

thus,

$$E_s = 20.9 \text{ Vrms}$$

Using this information and Section 5.1.1. the transformer design is as follows below:

1. The total VA output of this transformer is the product of the secondary winding rms voltage and the secondary winding rms current, thus,

$$VA = 20.9 \text{ Vrms} \times 7.1 \text{ Arms} = 148.39 \text{ V-A}$$

2. Using an estimated efficiency of 90% gives 164.88 V-A for the primary.

3. Using Equation (56) from Appendix A, the circular mils per ampere, CM/A, is found from,

$$CM/A = 4.205 \times (1 + .0047 \times 100^{o}C) \times \left\{ 40 \times \text{Log}_{10}\left[(164.88 \text{ V-A}) \times \left(\frac{400}{57 \text{ Hz}}\right)^{.8}\right] - 50^{o}C + 52 \right\}$$

or

$$CM/A = 706$$

4. The secondary rms current is 7.1 Arms. The primary rms current is found by dividing the primary VA by the primary voltage. Thus,

$$I_p = 164.88 \text{ V-A}/103.5 \text{ Vrms} = 1.59 \text{ Arms maximum}$$

Using these figures and the result of step 3, the primary and secondary wire sizes should have approximately 1121 CM and 5013 CM, respectively. Using tables for solid wires, 20 AWG and 13 AWG wires will provide approximately these CM for the primary and secondary windings, respectively.

5. Using Equation (52) from Appendix A gives for the AcAw product,

$$AcAw = \frac{(25) \times (164.88) \times (1 \times \ 2)}{\sqrt{2} \times (2.54)^2 \times 15000 \text{ G} \times .95 \times .3 \times 57 \text{ Hz}} \times \text{CM/A}$$

$$AcAw = 3.16 \text{ in}^4$$

6. Using this product and the "Rule of 2s", a core is selected from a catalog or preferably from the existing list in NAC drawing number 200AS152 (formerly drawing number 62A5A53). The tentative core selected from 200AS152 is numbered 200AS152-14, which has a commercial part number of either AA431-L, A431-L, or CA431 from references [C-1], [C-2], and [C-3], respectively. The core dimensions in inches and "Rule of 2s" ratios are,

D = 2.047max, 2.000 min F = .859 min

E = .750 ± .031 G = 2.297 min

$D/E = 2.667$ $G/F = 2.674$ $\dfrac{G \times F}{D \times E} = 1.315$

7. The primary turns required are found from,

$$N_p = \frac{\text{Erms}}{\pi\sqrt{2} \times B \times Ai \times f \times 10^{-8}}$$

$$= \frac{126.5 \text{ Vrms}}{\pi\sqrt{2} \times 15000 \text{ G} \times .75 \times 2 \times (2.54)^2 \times 57 \text{ Hz} \times 10^{-8}}$$

$$N_p = 344$$

8. For a 60 Hz transformer the number of turns for the secondary is found from,

$$N_s = N_p \times (E_s/E_p) \times 1.05$$

$$N_s = 344 \times (20.9/126.5) \times 1.05$$

$$N_s = 60$$

9. To maximize coupling, split coil construction will be used. Thus, one half of each winding will be on each of the two coils, and the resulting turns per winding are,

$$N_p = 172$$

$$N_s = 30$$

The margin is found from,

$$\text{Margin} = 0.8 \times (\text{Dia})^{.5}$$

where Dia = wire maximum diameter,
= .0353 in for 20 AWG
= .0765 in for 13 AWG

The resulting design margins should be about,

Primary = .150 in
Secondary = .221 in

The layer length, L, is then,

$$L = \text{Gmin} - 2(\text{margin})$$
$$= 2.297 - 2(.150) = 1.997 \text{ in, Primary}$$
$$= 2.297 - 2(.221) = 1.855 \text{ in, Secondary}$$

To estimate the number of layers required for the windings, the total length of a coil for the required number of turns is divided by the maximum layer length as follows:

for the Primary number of layers = 172(.0353)/1.997 = 3.04

and for the Secondary number of layers = 30(.0765)/1.855 = 1.24

Assuming that 3 layers of 57 turns each are used in the primary winding, and 2 layers of 15 turns each are used on the secondary winding, the resulting layer lengths are,

Primary layer length = (57 + 1) x (1.02) x (.0353 in) = 2.088 in

Secondary layer length = (15 + 1) x (1.02) x (.0765 in) = 1.248 in

10. The coil buildup is calculated to see if the windings, with coil form and insulation, will fit into the core window. The coil form to be used with this core is selected from drawing number 60A5A274 - 54. This coil form has a maximum wall thickness of .037 inches. Using an interlayer insulation 7 mils thick for the primary layers, 10 mils between the secondary layers, 10 mils between windings, and a 4-mil coil outer wrap gives a coil buildup of,

$$\begin{aligned} \text{Buildup} = {} & .037 \\ & + 3(.0353) + 2(.0765) \\ & + 2(.007) \times 1(.010) \\ & + 1(.010) \\ & + .004 \\ & + 10\% \end{aligned}$$

Buildup = .368 in

The winding buildup for the two coils fills about 86% of the core window in the F direction. However, in the G direction, the margin is .525 inches for the secondary winding, which is unnecessarily large. Using a bifilar wound secondary with 16 or maybe 15 AWG size wire would reduce the margin and in addition will improve coupling. Another option is to use a core with more optimal dimensions.

C.2 Design of a 60 Hz Primary Power Transformer Using the CODED-T Program

In this example the transformer for the circuit in Figure C.1 is redesigned using the identical requirements from C.1, but using the CODED-T program instead of the manual method of design. This example begins with a brief discussion of the transformer specification and the required inputs to the program. A sample session using the program is presented along with the program output.

The transformer specifications from C.1 are restated below:

Prima[illegible]quency = Single-Phase, 60 Hz $\pm$ 5%
voltage = 115 Vrms $\pm$ 10%

Secondary: voltage = 10.45 Vrms maximum, center-tapped
current = 10.0 Arms maximum (at rectifier output)

The minimum required inputs to the CODED-T program are the maximum primary RMS voltage, the minimum and maximum primary frequencies, the maximum secondary RMS voltage (across the secondary winding), and the maximum secondary RMS current. Other inputs are optional, such as regulation, maximum winding resistance, tap voltage in percent, and volume, which, if not specified, will be assigned default values. A complete list of the optional inputs, and their default values, is shown in Table C-1.

Table C-1. List of Optional Inputs and Default Values

Tolerance	= 5%	Ambient Temperature	= 100^{o}F
Regulation	= 10%	Temperature Rise	= 50^{o}F
Tap Voltage	= 0%	Volume	= 10^{5}in^{3}
Kilovolts to Core	= .3 kV	L Dimension	= 10000 in
Max DC Current	= 0 Amps	S Dimension	= 10000 in
Max Winding Rst	= 999999 ohm	Max Magnetizing Current	= 0 Amps
Max Phase Shift	= 90^{o}		

Since this design example has a center-tapped secondary, the percent tap voltage must be changed from its default value of 0% to 50%. Also, since the center-tapped secondary voltage is 10.45 Vrms, the required secondary winding voltage is 20.9 Vrms, and the secondary current is $10/\sqrt{2}$ Arms or 7.1 Arms. Thus, the required inputs to the CODED-T program for this example are:

Maximum primary voltage = 126 Vrms
Minimum primary frequency = 57 Hz
Maximum primary frequency = 63 Hz
Maximum secondary voltage = 20.9 Vrms
Maximum secondary current = 7.1 Arms
Secondary tap voltage = 50%

These inputs are used in an example session using the CODED-T transformer design program. In the following session, the messages are shown as they appear on the screen; responses which the user must enter are underlined. The CODED-T program is accessed from any user-ID on the Honeywell central computer by typing the system command XFRMER. Upon entering this command, the following help message will appear on the terminal screen:

```
TRANSFORMER DESIGN PROGRAM (TDP) COMMANDS:
-----------------------------------------------------------
COMMAND   DESCRIPTION                                  VALID
-----------------------------------------------------------
  BACK  -- START OVER FOR CURRENT LEVEL OF INPUT      -- IOSC
  CRT   -- ENTER INPUT DATA FROM CRT (INPUT)          -- I
        -- PRINT DESIGN RESULTS AT CRT (OUTPUT)       --  O
  DISP  -- DISPLAY CURRENT (DEFAULT) INPUT DATA       -- I  C
  EXEC  -- EXECUTE PROGRAM USING CURRENT INPUT DATA   -- I  C
  HELP  -- DISPLAY HELP SCREEN (THIS SCREEN)          -- IOSC
  INIT  -- INITIALIZE CURRENT INPUT DATA (SET TO 0)   --    C
  MISC  -- ENTER MISCELLANEOUS INPUT DATA             --    C
  NEXT  -- ENTER INPUT DATA FOR NEXT WINDING          --    C
  QUIT  -- EXIT PROGRAM (RETURN TO TSS)               -- IOSC
  REDO  -- START OVER (RETURN TO INPUT FILE PROMPT)   -- IOSC
  SAVE  -- SAVE CURRENT INPUT DATA ON A FILE          -- I  C
-----------------------------------------------------------
ONLY THE FIRST LETTER OF A COMMAND IS REQUIRED
A NULL RESPONSE (RETURN ONLY) USES CURRENT [DEFAULT] VALUE
VALID:  I = INPUT FILE PROMPT,   O = OUTPUT FILE PROMPT,
        S = SAVE FILE PROMPT,    C = CRT IPUT DATA PROMPTS
-----------------------------------------------------------
PRESS RETURN TO PROCEED:
```

After pressing return the following messages are displayed:

```
    DEFAULT DATA INPUT FILE IS:
    DEFAULT DESIGN OUTPUT FILE:  CRT

ENTER INPUT FILE NAME (OR BACK, CRT, DISP, EXEC, HELP, QUIT, REDO, SAVE):
    FILE NAME (OR COMMAND)? C
```

which indicates that initially, no input file is specified and that the program output results will be displayed on the terminal. Entering CRT or "C" at the FILE NAME (OR COMMAND)? prompt instructs the program to accept input from the terminal. If a file had been previously created, its name could have been entered here, and the program would then read in the data from this file. However, since the data for this example have not been previously saved in a file, these data must be entered as illustrated below. A null response (a carriage return) means that the default value is acceptable.

```
ENTER THE REQUESTED DATA (OR BACK, DISP, EXEC, HELP, INIT, MISC, NEXT,
QUIT, REDO, SAVE).  CURRENT (DEFAULT) VALUES ARE IN [BRACKETS].

    TRANSFORMER/INDUCTOR (0/1)       [0]?  [_]
    NUMBER OF SECONDARIES (1 - 8)    [1]?  [1]
    SERIAL NUMBER OF DATA SET    [     0]? [1      ]

    PRIMARY
      MAX RMS VOLTS          [    0.00000]? [126.5    ]  (MUST BE > 0)
      TAP VOLTAGE (PCT RMS)       [ 0.00]? [N      ]

    SECONDARY NO. 1
      MAX RMS VOLTS          [    0.00000]? [20.9     ]  (MUST BE > 0)
      MAX AC CURRENT (MA)    [    0.00000]? [7100.0   ]  (MUST BE > 0)
      TOLERANCE (PCT)             [ 0.00]? [2.5    ]   (IF 0, 5 IS USED)
      REGULATION (PCT)            [ 0.00]? [5      ]   (IF 0, 10 IS USED)
      TAP VOLTAGE (PCT RMS)       [ 0.00]? [50     ]
      KILOVOLTS TO CORE           [ 0.00]? [N      ]   (IF 0, 0.3 IS USED)
```

```
MISCELLANEOUS DATA
  MIN FREQUENCY (CPS)  [     0.00000]?  [57        ]  (MUST BE > 0)
  MAX FREQUENCY (CPS)  [     0.00000]?  [63        ]  (MUST BE > 0)
  SINGLE/THREE-PHASE (0/1)         [0]?  [_]
  SINGLE/SPLIT COIL (0/1)          [0]?  [1]
  AMBIENT TEMPERATURE(DEG C) [  0.0]?  [N         ]  (IF 0, 100 IS USED)

EXECUTE PROGRAM USING CURRENT INPUT DATA (Y OR N) [N]? Y
```

After entering the data, the program asks the user whether to execute using the data as entered. If "N" is entered, the program starts over from the point where input data is requested. After a "Y" response, the program asks for the name of an output file, which can be either a file in the user's ID or the CRT. The design output data will be routed to whichever is chosen. In this example, the file "XFRMROUT" was chosen, into which the CODED-T program wrote the design output data. This file is shown on pages C-10 through C-13. Page C-14 shows the input data and default options.

```
ENTER DESIGN OUPUT FILE NAME (OR BACK, CRT, HELP, QUIT, REDO):
    FILE NAME (OR COMMAND)? XFRMROUT
    THIS FILE EXISTS.  OVERWRITE (Y OR N) [N]? Y


DEFAULT DATA INPUT FILE IS:
DEFAULT DESIGN OUTPUT FILE:  XFRMROUT

  ENTER INPUT FILE NAME (OR BACK, CRT, DISP, EXEC, HELP, QUIT, REDO, SAVE):
      FILE NAME (OR COMMAND)? Q
```

POWER TRANSFORMER DESIGN READOUT

PART 1 - CONSTRUCTION PARAMETERS

A. IDENTIFICATION

CODED SERIAL NUMBER - 1

B. MATERIAL

1.	TEMPERATURE CLASS - F		
2.	CORE - 12 MIL TAPEWOUND C-TYPE:		
	CATALOGUE NO. A179	SINGLE-PHASE	BUWEPS 62A5A53 - 15 LIST NO. 15
3.	COIL FORM:		
	2.062 BY 0.781 BY 2.500 BY 0.032 (IN)		BUWEPS 60A5A274 - 54
4.	WIRE	POLYTHERMALEZE	BUWEPS 61A5A9
5.	INSULATION	SILICONE VARNISHED GLASS	BUWEPS 61A5A169
6.	IMPREGNANT	EPOXY	BUWEPS 64A5A9-1
7.	ENCAPSULANT	THIXOTROPIC EPOXY	BUWEPS 62A5A194
8.	SOLDER	SP&B SEAL, 5% TIN ALLOY	QQ-S-571 TYPE SN5WS
9.	ADHESIVE TAPE	SILICONE VARNISHED GLASS	BUWEPS 61A5A157 - 3
10.	FLUX		MIL-F-14256 TYPE A
11.	BRACKETS		
12.	BAND NO.		BUWEPS 57A5A256
13.	SEAL NO.		BUWEPS 57A5A258

C. CONSTRUCTION

1. COIL (START ALL WINDINGS AT SAME END OF COIL UNLESS OTHERWISE SPECIFIED)

WINDING TYPE - SINGLE-PHASE SPLIT COIL

NO. OF COILS REQUIRED 2

WINDING NUMBER....	1	2
2. WIRE		
GAGE	13	20
TURNS/WNDG/COIL ..	31	179
TAP AT TURNS	31	0

TURNS/LAYER	16	60
LAYERS REQUIRED ..	2	3
MARGIN (IN)	0.635	0.139
WIRE LENGTH (FT) .	33	223

3. INSULATION

INTERLAYER (MIL) .	10	7
LAYERS REQUIRED ..		
INTERWNDG (MIL) ..	7	0
LAYERS REQUIRED ..		

COIL TO COIL FORM ...... NONE REQUIRED

OUTER WRAP OVER COIL .... 4 MIL BUWEPS 61A5A157

SHIELDING REQUIRED ..... NONE REQUIRED

4. COIL BUILD

MANDREL 2.010 BY 0.760 BY 2.500 IN

MAX BUILDUP (OVER LEADS) .... IN

MAX BUILDUP (NO LEADS) ...... 0.352 IN

5. CORE

AIR GAP REQUIRED . NONE REQUIRED

WEIGHT (IRON) .... 3.55 LB

TOTAL WEIGHT ..... 6.13 LB

POWER TRANSFORMER DESIGN READOUT

PART 2 - ELECTRICAL PARAMETERS (CODED SERIAL NUMBER - 1)

A. WINDING RATINGS

WINDING NUMBER	1	2
WINDING DESIGNATION ..	SEC. 1	PRIM
POWER (VOLT-AMPS)	148.8	157.8
VOLTAGE RATINGS		
FULL LOAD VOLTAGE	20.9	126.5
NO LOAD VOLTAGE	21.9	126.5
TAP AT VOLTS	10.5	0.0
VOLTAGE REG. (PCT) ...	4.7	
TOLERANCE (PCT)	2.5	
CURRENT RATINGS (MA)		
RMS CURRENT	7100.0	1221.7
D-C CURRENT	0.0	0.0
IMPEDANCE (OHMS)		
WINDING RESISTANCE ...	6.69E-02	2.30E+00
RESISTIVE LOAD	2.9	
CAP. REACT. TO CORE ..	20312.8	27691.3
PARAMETERS OF SECONDARIES REFERRED TO PRIMARY		
PHASE SHIFT (DEG)	-0.4	
LEAK. IND. (MILHYS) ..	0.9	
IND. REACT. (OHMS) ...	0.3	
INTERWNDG CAP. (MMF) .	359.9	
CAP. REACT. (KOHMS) ..	7304.5	

B. SIZE (NOMINAL)

DOES NOT INCLUDE ENCAPSULATION, BAND SEAL, BRACKETS, OR TERMINAL BULGE.

VOLUME	39.855	CUBIC IN
OVERALL DIMENSIONS		
HEIGHT (F DIR.)	3.375	IN
WIDTH (D DIR.)	2.938	IN
LENGTH (G DIR.)	4.020	IN

C. GENERAL RATINGS

EFFICIENCY	94.3	PCT
FLUX DENSITY	15158.	GAUSS
MAX FREQUENCY	63.0	CPS
MIN FREQUENCY	57.0	CPS
RESONANT FREQUENCY ...	170650.	CPS
AMBIENT TEMPERATURE ..	100.	DEG C
TEMPERATURE RISE	28.	DEG C
TEMPERATURE CLASS		F
OPERATING LIFE	20000.	HRS

D. GENERAL PARAMETERS

CORE LOSS	2.2	WATTS
COPPER LOSS	6.8	WATTS
EXCITATION LOSS	1.8	VOLT-AMPS
MAGNETIZING LOSS	9.9	VOLT-AMPS
CORE LOSS CUR (NOM) ..	0.018	AMPS
MAGNETIZE CUR (MAX) ..	0.079	AMPS
IRON SHUNT RESIST	6895.	OHMS
IRON SHUNT REACT	1613.	OHMS

LISTING OF DATA INPUT FILE:

DATA SET SERIAL NUMBER: 1

WINDING DATA:

CRD NUM	MAX RMS VOLTS	MAX AC CUR (MA)	TOL (PCT)	REG (PCT)	TAP V (PCT)	KV TO CORE	MAX DC CUR (MA)	MAX WNDG RST (OHM)	MAX PS (DEG)
0	126.50000	0.00000	0.00	0.00	0.00	0.00	0.000	0.000	0.00
1	20.90000	7100.00000	2.50	5.00	50.00	0.00	0.000	0.000	0.00

MISCELLANEOUS DATA:

CRD NUM	MIN FREQ (CPS)	MAX FREQ (CPS)	S C	3 P	C T	AMB T (DEG)	T RIS (DEG)	VOL (IN3)	L DIM (IN)	S DIM (IN)	MAX MAG CUR (MA)	N S	S H
9	57.00000	63.00000	1	0	0	0.0	0.0	0.0	0.00	0.00	0.000	0	0

THE PROGRAM WILL REPLACE ZERO VALUES WITH DEFAULT VALUES.
DEFAULT VALUES:

TOL (PCT) =	5.0	AMB T (DEG) =	100.0
REG (PCT) =	10.0	T RIS (DEG =	50.0
TAP V (PCT) =	0.0	VOL (IN3) =	100000.0
KV TO CORE =	0.3	L DIM (IN) =	10000.0
MAX DC CUR =	0.0	S DIM (IN) =	10000.0
MAX WNDG RST =	999999.0	MAX MAG CUR =	0.0
MAX PS (DEG) =	90.0		

C.3 Design of a 2-Pulse Rectified, 400 Hz Primary Power Inductor Using the Manual Approach

In this example an inductor is designed for a two pulse rectified 400 Hz power line, using the method outlined in 6.1.1. The electrical performance specifications are outlined below:

Inductance = .002 henries, H, at 5 Adc
= .000765 H minimum at 30 Adc
Erms = 20 Vrms at 800 Hz

This application will use a 4-mil tape wound C-core, using split coil construction. Following the procedure in section 6.1.1., the inductor design is as follows:

1. Since the inductance is already given as a specification, the first step of calculating the critical inductance is not required.

2. A tentative core size is chosen by using the equation for the AcAw product (Equation (63), Appendix A),

$$AcAW = \frac{(10^2) \times (L) \times \hat{I} \times (Irms) \times (CM/A)}{(B \times Kc \times Kw) \times (2.54^2)}$$

where,

Ac = gross core cross-sectional area in in^2
Aw = core window area in in^2
L = inductance in henries
= .002 H at 5 Adc
= .000765 H at 30 Adc

$\hat{I}ac$ = inductor peak ripple current in amperes = $\frac{\Delta I}{2} = \frac{\hat{E}}{2\pi \times f \times Lmin}$

Irms = inductor RMS current in Arms
$\hat{I}$ = inductor peak current
B = peak flux density in gauss = 16000 G

Kc = core stacking factor = .90

Kw = core window fill factor = .60

CM/A = 600 circular mils per ampere

where $\hat{E}$ = peak inductor Ac voltage = Erms x $\sqrt{2}$

f = frequency = 800 Hz

Lmin = minimum inductance

$$\hat{I}ac = \frac{20\sqrt{2}}{2\pi \times 800 \times (.000765)}$$

$\hat{I}ac$ = 7.356 A peak

Irms = $[30^2 + (7.356/\sqrt{3})^2]^{1/2}$ (Equation (38), Appendix A)

Irms = 30.299 Arms

$\hat{I}$ = 30 A + 7.356 A = 37.356 A (Equation (29), Appendix A)

Because of range of DC current, design for 20% over the minimum inductance (.000765)(1.20) = .000918 H.

$$AcAw = \frac{(10^2)(.000918)(37.356)(30.299)(600)}{(16000)(0.9)(0.6)(2.54^2)}$$

$$= 1.118 \text{ in}^4$$

As a check, also calculate AcAw for the low DC current condition. At Idc = 5A the inductance required is .002 H,

$$\text{then } \hat{I}ac = \frac{20\sqrt{2}}{(2\pi)(800)(.002)} = 2.813 \text{ A}$$

$$\text{Irms} = \left[5^2 + \left(\frac{2.813}{\sqrt{3}}\right)^2\right]^{1/2}$$

$$= 5.257 \text{ A}$$

$$\hat{I} = 5 + 2.813$$

$$= 7.813 \text{ A}$$

The inductor wires must be sized for the high current condition, and thus the required CM/A will be multiplied by the RMS current for the high current condition. Also, at the high current condition the flux density will be a maximum, Bmax, which is designed to be 16000 gauss. For this calculation, since the flux density is proportional to the current, the flux density, Bmin, for the low current condition is,

$$Bmin = Bmax\left(\frac{7.813\ A}{37.356A}\right)$$

$$= 16000\left(\frac{7.813}{37.356}\right) \text{gauss}$$

$$Bmin = 3346 \text{ gauss}$$

The AcAw product is then,

$$AcAw = \frac{(10^2)(.002)(7.813)(30.299)(600)}{(3346)(0.9)(0.6)(2.54^2)}$$

or,

$$AcAw = 2.437 \text{ in}^4$$

Since this figure is larger than that found using the high current (low inductance) condition, the AcAw product for this example is based on the low current, high inductance requirement using the high current condition for determining flux density and wire size.

3. The tentative core selected is from NAC drawing number 200AS138-25 or commercial number H188. The dimensions are:

AcAw product = 3.375 inches^4
D = 1.500 inches

E = .750 inches

F = 1.000 inches

G = 3.000 inches

Ac = DE = 1.125 inches^2

Aw = FG = 3.000 inches^2

Kc = 0.9

4. Rearranging Equation (12) from Appendix A and using the relationship between Ai and Ac, the number of turns is found from,

$$N = \frac{L\hat{I} \times 10^8}{B\ Ac\ Kc}$$

Inserting values, and multiplying Ac by 2.54^2 to convert to cm^2 gives,

$$N = \frac{(.002)(7.813)(10^8)}{(16000)\left(\dfrac{7.813}{37.356}\right)(1.125)(.9)(2.54^2)}$$

= 71.5 turns

Use 72 turns

5. Calculate the wire size to be used. At 600 CM/A the conductor cross-sectional area will be

(600)(30.448) = 18269 CM which is between AWG 7 and AWG 8

It is advisable to use 2 strands of wire connected in parallel. For this design use 2 strands of AWG 11 which is closely equivalent to AWG 8, but will bulge less in the winding.

6. Trial margin = $(.0957)^{1/2}$ = .309 but the margin need not be larger than about .250 inches.

Length of each layer = 3.000 - (2 x .250)
= 2.500 inches

$$\text{Diameter/layer} = \frac{2.5}{(.0957)(1.02)} = 25.6$$

Try using 24 turns/layer, which results in

$$\frac{72}{24} = 3 \text{ layers required.}$$

If 3 layers can be accommodated, the flux density can be reduced and/or the inductance increased. The winding buildup is calculated below.

Buildup = .025 coil form
.0957 AWG 11
.007 2 layers of 3.5 mil Kapton
.0957
.007
.0957
+ .007
.3331
+ .0333 bulge
= .3664 total buildup per coil or
.7328 total for windings.

1.000 - .733 = .267 inch clearance between windings.

Since there can be 3 layers on each leg, the turns per leg (and total turns when windings are paralleled) will be (24 turns/layer)(3 layers) = 72 turns.

7. Since the buildup is less than 75% of the F dimension, another iteration, using perhaps a smaller core or maybe larger wire, might result in a more optimum design.

8. Calculate the required air gap using the expression,

$$\ell g + \frac{\ell c}{\mu \Delta} = \frac{.4\pi \, N \, \hat{I}}{B} \left(\frac{1}{2.54} \right) \text{ inches}$$

$$= \frac{(.4\pi)(72)(37.356)}{16000} \left(\frac{1}{2.54} \right)$$

$$= .083 \text{ inch assuming no fringing.}$$

Since there will be fringing, the gap per leg used is .070 to give the required inductance values.

C.4 Design of a High Frequency Switchmode Transformer

This section presents a design example for a switchmode power supply transformer. The transformer configuration is push-pull, and there are three secondaries. The design equations are discussed in Section 5.2, Design of Switchmode Power Transformers, and restated in Appendix A.

Definitions

E_p	Primary voltage
E_s	Peak secondary voltage for a given primary voltage
N_p	Number of turns on one-half of the push-pull transformer primary
N_s	Number of active turns on the secondary
B	Flux density in gauss
A_c	Core gross area = Ai (core net area)
f	Switching frequency
CM/A	Circular mils per ampere rms
t_{on}	on time = duty cycle x (1/f)

Specifications

Transformer Configuration: Push-pull

$f = 70$ kHz

$E_p = 130$ V to 172 V

Outputs

5 Vdc +/-10% at 900 mA max

+/-15 Vdc +/-10% at 200 mA max

RMS Transformer Currents

Primary 0.124 Arms

+5 V Secondary 0.636 Arms

+/-15 V Secondary 0.200 Arms

1. From Section 5.2.1, the required output power is calculated as,

$$[(5\ \mathrm{V}) \times (0.9\ \mathrm{A}) \times (1.2)] + 2 \times [(15\ \mathrm{V}) \times (0.2\ \mathrm{A}) \times (1.1)] = 12.00\ \mathrm{W}$$

2. A P-42616-UG, pot core number 2616, is selected using the ferrite core selection table in the Magnetics Inc. catalog [ref. C-4]. This selection is based on the output power calculation and the available space. The minimum cross-sectional area for this core is 0.7610 cm. (See Section 3.3.3, Figure 4.)

3. The number of primary turns is calculated using

$$E_p \times t_{on} = 2B \times Ac \times Np \times 10^{-8}$$

E_p will be less than the input voltage due to rectifier drops, switch losses, etc. E_p is also dependent on the input filter; voltage ripple will decrease as the value of capacitance in the input filter increases.

E_p and t_{on} are interrelated; t_{on} will reach its maximum value when E_p is a minimum. If a maximum duty cycle of 0.85 is desired $t_{on} = 0.85/f = 12.14 \times 10^{-6}$ seconds.

Then

$$N_p = \frac{E_p \times t_{on} \times 10^8}{2 \times B \times Ac \times f} = \frac{130 \times 12.14 \times 10^{-6} \times 10^8}{2 \times 2000 \times 0.761} = 51.85 \text{ turns}$$

The number of turns must be an integer; N_p is 52 turns. Because this transformer will be used in a push-pull configuration, two separate primary windings of equal turns are needed. The minimum number of turns per primary is 52.

4. The secondary voltage is dependent on the output voltage, the duty cycle, and any losses between the secondary and the output, including rectifier losses and the drop across the output inductor. In cases with more than one output voltage, one output is used to determine the optimum N_p:N_s ratio. The turns for the remaining secondaries are determined as a function of the ratio of the secondary voltages.

$$E_s = \frac{\text{Vout} + \text{Vloss}}{\text{duty cycle}} = \frac{5 + 1}{.85} = 7.06$$

5. For this configuration, N_s corresponds to the number of turns in each of the +5 V secondary windings. The number of secondary turns is found from

$$N_s = \frac{E_s \times N_p \times 1.02}{E_p} = 2.87$$

The number of secondary turns must also be an integer. Rounding N_s to 3 will increase E_s and decrease the duty cycle. Other alternatives are to change N_p to reach the optimum N_p:N_s ratio or to choose another core.

The +/-15 V outputs are both derived from the 15 V secondary windings. For this calculation, Vout is 2 x 15 = 30 V. Vloss is 2 V for the two diode drops and the two inductor drops. The number of turns are determined as follows:

$$\frac{(\text{Vout} + \text{Vloss}) \times 3}{(\text{Vout} + \text{Vloss})} = \frac{(30 + 2) \times 3}{(5 + 1)} = 16$$

For every 3 turns on the 5 V secondary, the +/-15 V secondary will have 16 turns, 8 in each half.

6. The next step would be to calculate the RMS current for the primary and each secondary. However, since these currents are given in the specifications, this step is not required.

7. The RMS current is used to determine the minimum wire size. The equation is:

desired CM/A x Irms = CM required

The desired CM/A is in the range of 500-600. The minimum wire size is then selected. The wire size may be increased to better fill the available space. If space is tight, the desired CM/A may be decreased as low as 400 CM/A. This results in smaller wire, more turns in a given space, and more heat. The minimum wire sizes for 500 CM/A or more in each winding are given below. The CM/A does not have to be the same for every winding on the core.

Primary	AWG #32
5 V Secondary	AWG #25
+/-15 V Secondary	AWG #30

8. Fitting the windings into the available space can require several iterations. Since this transformer will be wound on a pot core, Table 2 (section 5.2.1) can be used to help fill the layers.

The primary will be placed on the bobbin first. 104 bifilar turns (two parallel strands of wire wound together) will require 6 layers. 104 is twice the minimum turns required per primary winding. The primary is covered with two layers of insulation (2.5 mil Kapton), necessary due to the difference in voltage between the primary and the secondaries. The space remaining on the bobbin is

bobbin height	.166 in
less 6 layers AWG 31 6 x .0108	-.0648 in
less 2 layers insulation 2 x .0025	-.0050 in
available height	.0962 in

Because insulation will be placed between the 5 V secondary and the +/-15 V secondary and on top of the last winding, 0.01 inches must be subtracted from the available height, leaving .0862 in.

The 5 V secondary will have only 6 bifilar turns and will probably not fill a layer. Note that the minimum of 3 turns on each half of the +5 V secondary has been doubled to maintain the proper turns ratio with the primary. Therefore, the +/-15 secondary will be placed on the bobbin on top of the primary. For every 3 turns in each half of the +5 V secondary, each winding of the +/-15 V secondary will have 8 turns.

number of turns per winding = (8/3) x 6 = 16 turns. This becomes 16 bifilar turns.

number of layers of AWG #25 = 2

available space = .0862 - (2 x .0206) = .0450 in

The space remaining for the 5 V secondary is more than adequate. Usually this would indicate that the number of turns should be increased, the turns ratio(s) changed, and/or the wire sizes changed. However, in this particular case, the transformer design is restricted by the coupled inductors that follow. The turns ratios for the secondaries on the transformer must be the same as the turns ratios on the coupled inductor. Calculations will show that not enough room remains to increase the number of turns while maintaining the same turns ratios. Increasing the wire sizes will cause partially filled layers, possibly resulting in errors or difficulty in winding. The wire selected for the +5 V secondary is AWG #23 in order to fill the layer.

C.5 Design of a Coupled Inductor

This section presents an example of a coupled inductor design. The coupled inductor will be used in the same circuit as the push-pull transformer designed in Appendix C.4.

Definitions

Lc	Critical inductance
Vout	Output voltage
f	Switching frequency
toff	Maximum off time = (1 - min duty cycle)/f
I_n	DC current @Output n
I1	DC current @ +5 V = 0.5 to 0.9 A
I2	DC current @ +15 V = 50 to 200 mA
I3	DC current @ -15 V = 50 to 200 mA
n1	Turns on 5 V transformer winding
n2	Turns on 15 V transformer winding
n3	Turns on -15 V transformer winding
N_1	Turns on 5 V inductor winding
N_2	Turns on +15 V inductor winding
N_3	Turns on -15 V inductor winding
k	Desired value of design inductance/critical inductance. This example will use k = 2.
Ac	Gross core area in cm^2
B	Flux density, in gauss. This example will use 2400 gauss.

Specifications (from Appendix C.4),

n1 = 6

n2 = n3 = 16

toff = 5.71×10^{-6} sec

1. The equation for critical inductance for one winding of a coupled inductor is, from Equation (45), Appendix A,

$$Lc = \frac{Vout \times toff}{2 \times (I1 + I2 \times n2/n1 + I3 \times n3/n1)}$$

where In is the minimum DC current for winding n.

which gives for the +5 V output inductor a value of 20.5×10^{-6} H for Lc. The design inductance is $L = 2 \times 1.1 \times Lc$ or $L = 45.1 \times 10^{-6}$H.

2. The maximum peak inductor currents, $\hat{I}n,max$, are calculated using Equation (47), Appendix A, as follows:

$$\hat{I}n,max = Indc,max + \frac{\Delta In,max}{2}$$

or using Equation (35), Appendix A $\Delta In,max = \dfrac{2Indc,min}{Kn}$,

$$\hat{I}n,max = Indc,max + \frac{Indc,min}{Kn}$$

which gives,

$$\hat{I}_1,max = 1.15 \text{ A}$$
$$\hat{I}_2,max = .225 \text{ A}$$
$$\hat{I}_3,max = .225 \text{ A}$$

3. The minimum number of turns, N_1, required for the +5 V inductor is calculated from Equation (48), Appendix A, which is rearranged as,

$$N_1 = \frac{L \times \left[\hat{I}_1,\text{max} + \frac{n_2}{n_1} \hat{I}_2,\text{max} + \frac{n_3}{n_1} \hat{I}_3,\text{max} \right]}{B \times Ac \times 10^{-8}}$$

inserting values give,

$$N_1 = \frac{(45.1 \times 10^{-6}) \left[1.15 + \frac{16}{6} (.225) + \frac{16}{6} (.225) \right]}{(2400) \times (.761) \times (10^{-8})}$$

or

$$N_1 = 5.8 \text{ or } 6 \text{ turns minimum.}$$

The Ac value of .761 is for a 2616 pot core since it is common to start the coupled inductor design with the same core as used in the transformer.

The number of turns for the +15 V and the -15 V inductor are determined by the transformer turns ratio: 8 turns each for every 3 turns on the +5 V inductor.

4. As pointed out, the core size is chosen tentatively to be a 2616 pot core.

5. Determination of minimum wire size is discussed in Appendix A. For this design, AWG #23 is selected for the +5 V inductor, and AWG #28 is selected for the +15 V and -15 V inductors.

6. The next step is to wind the bobbin. It is best not to use bifilar windings in a coupled inductor so there will be 3 separate windings separated by 2 layers of 2.5 mil Kapton tape.

The maximum number of turns are put on each winding to increase the inductance. If the +5 V inductor has 27 turns, each of the two remaining inductors will have 72 turns. The bobbin will be filled as follows:

bobbin height		.1660 in
less +5 V inductor: 2 layers AWG #23	2 x .0255	-.0510 in
less +15 V inductor: 3 layers AWG #28	3 x .0148	-.0444 in
less -15 V inductor: 3 layers AWG #28	3 x .0148	-.0444 in
2 layers insulation between each winding	4 x .0025	-.0100 in
2 layers insulation on last winding	2 x .0025	-.0050 in
remaining height		.0112 in

As with the transformer design in Appendix C.4, the bobbin is not completely full. However, the constraint that the inductor turns ratios must equal the transformer secondary turns ratios makes it difficult to fill the bobbin.

7. The critical inductance for the +15 V and -15 V inductors should be calculated to confirm that the design also meets this requirement.

$$Lc = \frac{Vout \times toff}{2 \times (I2 + I1 \times n1/n2 + I3 \times n3/n2)}$$

Lc for the +15 V inductor and for the -15 V is 0.149 mH.

Using the same equation used to determine the minimum number of turns for the +5V inductor, it can be shown the 72 turns each gives .393 mH, which is more than twice Lc.

8. The gap, ℓg, is calculated using Equation (41) from Appendix A assuming that $\ell c/\mu\Delta$ is very small compared to ℓg, thus,

$$\ell g = \frac{.4\pi}{B}\left[N_1\hat{I}_1,max + N_2\hat{I}_2,max, + N_3\hat{I}_3,max\right]$$

using the results from the above steps,

$$\ell g = \frac{.4\pi}{2400}\left[(27)(1.15) + (72)(.225) + (72)(.225)\right]$$

or,

$$\ell g = 33 \text{ mils.}$$

REFERENCES

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C.2 Hipersil Core Design Engineer's Handbook, Westinghouse Corp., 1965.

C.3 Tape Wound Magnetic Products, National Magnetics Corp., Cerritos, CA, 1985.

C.4 Ferrite Cores for Power and Filter Applications, Magnetics Inc., FC-405-12S, 1985.

APPENDIX D

TRANSFORMER LOAD COMPARISONS

Transformer Load Comparisons

Constant Resistance, e.g., fixed resistive load.

Primary current, secondary current, and capacitor ripple current are highest at high line voltage. Secondary current and load component of primary current are inversely proportional to load resistance.

Constant Current Load, e.g., a linear regulator.

Secondary current and load component of primary current are essentially independent of the instantaneous and RMS line voltages. Capacitor ripple current is proportional to line voltage.

Constant Power Load, e.g., a switching power supply.

Load components of primary and secondary currents are highest at low line voltage and decrease as the instantaneous and RMS line voltages increase. Capacitor ripple current increases as the line voltage increases.

Calculation of Primary Currents in Transformers for Constant Resistance Loads, Constant Current Loads and Constant Power Loads.

For <u>Constant Resistance Load</u>,

$$Ip = \frac{1}{n}\left[\frac{E_p(h)}{E_p(\ell)} \times Is_1 \times \frac{Es_1(h)}{E_p(h)} + Is_2 \times \frac{Es_2(h)}{E_p(h)} + \ldots \text{etc.}\right]$$

$$= \frac{1}{n}\left(\frac{1}{E_p(\ell)} \times Is_1 \times Es_1(h) + Is_2 \times Es_2(h) + \ldots \text{etc.}\right)$$

Use (VA) computed at high line voltage to get Idc and ripple.

For Constant Current Load (e.g., linear regulator load),

$$Ip = \frac{1}{n}\left(Is_1 \times \frac{Es_1(h)}{E_p(h)} + Is_2 \times \frac{Es_2(h)}{E_p(h)} + \ldots \text{etc.}\right)$$

For Constant Power Load (e.g., switchmode power supply load),

$$Ip = \frac{1}{n}\,\frac{1}{E_p(\ell)}\left[(VA)_1 + (VA)_2 + \ldots \text{etc.}\right]$$

Note: Idc is greatest at low line whereas ripple is greatest at high line.

Ip = Load component of primary current (i.e., exciting current is not included).

n = Efficiency = $\dfrac{\text{output power}}{\text{input power}}$

$E_p(h)$ = Input voltage at high line voltage.

$E_p(\ell)$ = Input voltage at low line voltage.

$Es_1(h)$ = Output voltage of sec no. 1 at high line voltage.

$Es_2(h)$ = Output voltage of sec no. 2 at high line voltage.

Is_1 = RMS value of Idc and ripple current combined for sec no. 1.

Is_2 = RMS value of Idc and ripple current combined for sec no. 2.

$(V\text{-}A)_1$ = Product of RMS voltage and true RMS current (Idc and ripple combined) for sec no. 1.

$(V\text{-}A)_2$ = Product of RMS voltage and true RMS current (Idc and ripple combined) for sec no. 2.

END

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