Discontinuous-Mode Flyback Transformer

Journey from Specification to Finished Design

Introduction:

The first transformer that I ever designed was back in 1968. It was a 6.3 V, 60 Hz filament transformer for a vacuum tube amplifier. I designed it on a slide rule along with several paper catalogs. Remember those days?
Voltages were low, the size was not critical; the insulation scheme consisted of kraft paper. It was wound on a paper tube. The potting was paraffin wax. It was enclosed in a metal can.

Today, I use an HP calculator, a computer (two displays) and the internet. Transformers must operate at kilohertz ranges. The voltages are high. Sizes have shrunk.

The tools have changed throughout the years, but basic magnetic theory? Not so much.

What has changed is the environment: The invention of the switched-mode power supply in its many iterations has necessitated the push for higher frequency core materials, higher voltage insulation, higher power densities, smaller envelopes and lower cost designs.

Today, Magnetic Engineers must take into consideration factors that were unheard of back in 1968: Corona? Proximity and skin effects? Science fiction?

When looking for a subject to write about I thought of several candidates and ended up settling on this one as it embraces almost all of those factors that must be considered in order to develop a design that is efficient and low cost while meeting all of the customer's specifications. This subject is an actual design that is presently in production.

By Robin Benas, Engineering Manager -- ICE Components, Inc.

The Challenge:

Design a power transformer for a discontinuous-mode (DCM) power supply that requires a high input voltage, moderately high switching frequency, maximum efficiency, excellent coupling between windings (minimum leakage inductance), strict temperature rise limits under high ambient temperatures, and high isolation voltages; designed for minimum size and cost.

Basic Topology:

The transformer is used in a dual-switch flyback converter utilizing 1.5 KV MOSFETs to supply multiple 28 V regulated outputs and one 24 V output for cooling fan control. The converter design is based on a Texas Instruments application note: SNVA716 – July 2014.

Principle of Power Conversion:

Basic operation of the two-switch flyback powerstage is similar to traditional flyback topology. In the beginning of the switching period, both MOSFETs are closed and the primary of the transformer is connected between the input voltage and ground. Current starts to flow through the primary and the diode on the secondary side is reverse biased due to the transformer polarity. Therefore, all of the energy is stored in the transformer while the load current is supplied by the output capacitor. The time frame when the switches are closed is called the magnetizing period. When the switches are opened, stored energy is transferred to the output, supplying the load and the charging output capacitor. At the same time, the reflected voltage is applied over the primary. The time frame when the switches are open is called the demagnetizing period.

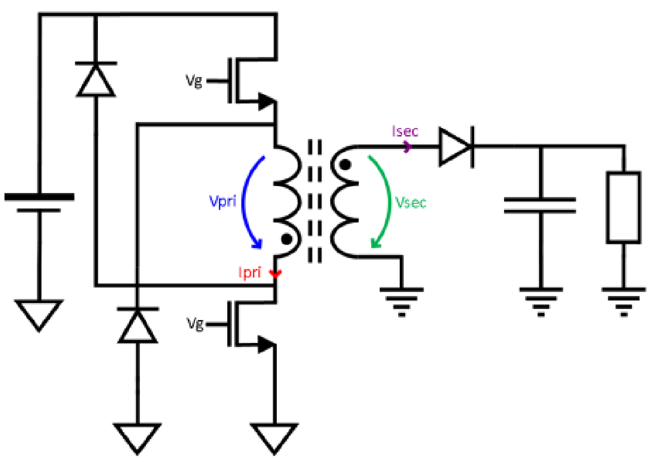


Figure 1. Two-Switch Flyback Power Stage

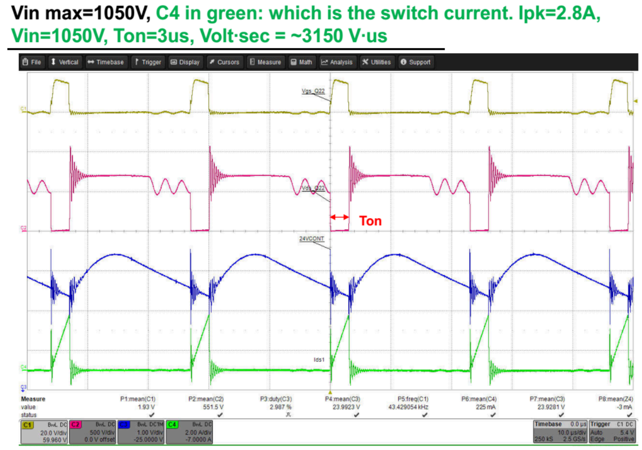


Figure 2. Example Waveforms

Specifications:

Primary Input Voltage: 1050 V_{PK} , square wave

Switching Frequency: 40 – 46 KHz

Duty Cycle (customer supplied): 26.4% nominal. See Figure 2 above.

Primary Inductance: 1100 µH +/- 5% @ 25°C

Δ Primary Inductance: < 2% from 25°C to -40°C

Nominal Primary Current = 1.45 A_{RMS}

Leakage Inductance:

- 40 µH maximum from primary to secondary 2 (S2)
- 3.0 µH maximum from S2 to all other secondaries shorted together.

Secondary Windings: 5 outputs (includes an auxiliary winding for voltage feedback)

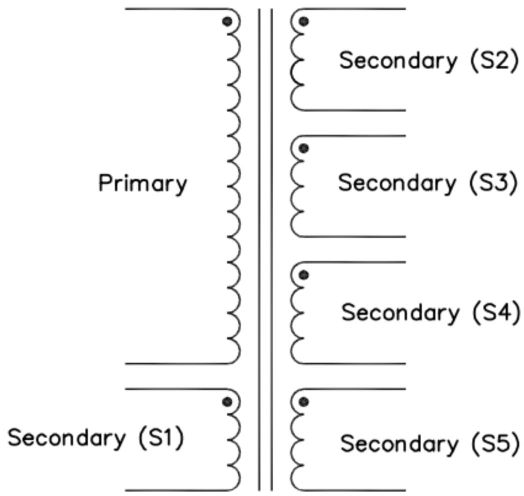


Figure 3. Schematic

Turns Ratios and Full-Load Currents:

Primary to S1 = 14:1, load current = 0.25 A_{RMS}

Primary to S2 = 10:1, load current = 8.45 A_{RMS}

Primary to S3 = 8.3:1, load current = 3.25 A_{RMS}

Primary to S4 = 8.3:1, load current = 1.10 A_{RMS}

Primary to S5 = 8.3:1, load current = 1.70 A_{RMS}

Dielectric strength (Hi-Pot): 3750 V_{RMS}, 60 Hz from primary plus S1 to all other secondaries. 1500 V_{DC} between all other secondaries.

Maximum ambient temperature: 80°C

Maximum operating temperature: 125°C

Maximum temperature rise: 45°C

Precautions:

1. High primary input voltage could cause dielectric breakdown within the primary coil due to a high turn to turn and layer to layer voltage gradients.
2. High primary input voltage suggests the tendency to develop corona (partial discharge) which could degrade the insulation over time.
3. High switching frequency could accelerate the effects of corona discharge resulting in damaged insulation over a *very* short period of time.
4. High ambient temperature requires that we keep the temperature rise to a minimum in order to stay within the temperature rating of the insulation. High-temperature insulating materials are expensive. Keeping the temperature rise low will allow us to utilize lower-cost insulation.
5. The flux density should be low enough as to avoid core saturation at the elevated temperatures. Note: It is typical for a ferrite core to exhibit a saturation flux density (βsat) at 100°C of 20 – 30% lower than the 25°C B_{SAT} rating.
6. Coupling between windings is critical to ensure low leakage inductance.
7. Considering the operating frequency and power level, care must be taken to select the appropriate wires sizes and stranding in order to minimize the copper losses resulting from skin and proximity effects.
8. Litz wires and other stranded conductor configurations are not efficient in their use of space. Whereas common film-coated magnet wire, randomly wound in a bobbin, may take up about 50% of the available volume, stranded wire bundles can only utilize about 30% of the available bobbin "window". These factors must be considered when selecting a core/bobbin size. I will use the 30% factor when designing the coil.
9. EAU is not high and price point is critical so standard-catalog materials (core and bobbin) are suggested.

10. Low primary inductance requires a large core gap. *Note:* A ferrite core is suggested in order to minimize the core losses. The permeability of ferrite is high, so a large gap is required to reduce the effective permeability. A large core gap could introduce a high level of fringing flux around the gap area which would cause excessive copper losses in the windings located adjacent the core gap.

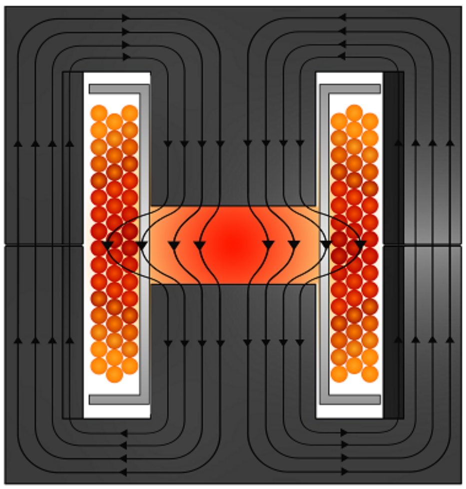


Figure 4. Fringing flux around core gap.

11. By definition, fringing flux is the magnetic field that tends to spread out into the surrounding space adjacent to the core's gap. In this case, the center leg of the core. The concentration of magnetic flux around the gap tends to produce eddy currents within the coil. Those eddy currents are often significant and can create hot spots which can burn through wire insulation. Please refer to the following photos: Figure 5a and 5b. The transformers in the photos are examples of the subject design. They are identical except one is wound directly over the bobbin coil form, and the other spaced approximately 4 mm outboard of the coil form. They were built to demonstrate the effect of fringing flux on the coil temperature rise. You can see a significant difference in their respective hot-spot temperatures.

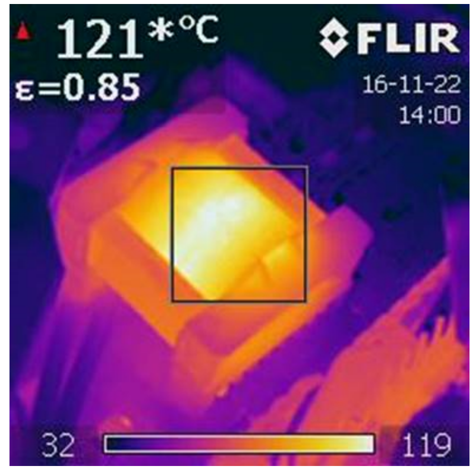


Figure 5a. Coil wound directly on the bobbin surface, 1.0 mm above the core gap.

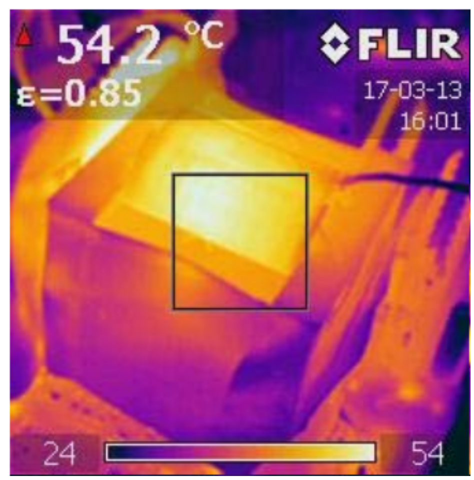


Figure 5b. Coil spaced 3.77 mm above the bobbin surface, 4.77 mm above the core gap.

Design Approach:

Previous experience suggests that an EC or ETD core with a printed circuit bobbin would be the best approach. The round center leg of the core provides a smooth, uniform winding surface. The bobbin flanges provide a barrier between the coil and core. The printed circuit pins provide an easy way to integrate the design with the customer's printed circuit board.

In addition to the mechanical advantages, the round winding surface has proven to be conducive to good coupling between windings which minimizes power losses due to leakage inductance.

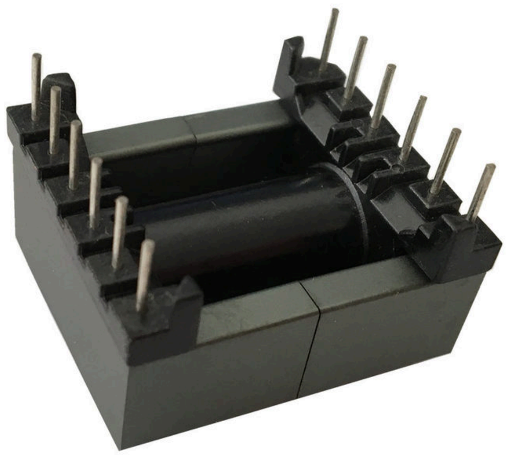


Figure 6.

Core Selection:

There are several ways of determining the minimum core size: Area product vs. total power throughput, minimum regulation specifications, etc. Considering the high ambient temperature, and high voltage specifications, I have chosen a possible core size based on the maximum total losses.

If I assume that the efficiency is about 90%, which is common for this topology, the power dissipation (Pt) = 150/0.90-150=16.7 W = 16700 mW. The trick is to find a core that will dissipate the losses and result in a maximum temperature rise of 45°C.

Heat Dissipation and Surface Area:

I will assume that the heat dissipation, due to the temperature rise, will be 50% radiated through air and 50% conduction through the printed circuit board. It is assumed that no forced air will be supplied (worst case). The surface area is calculated for the whole assembly including the core and coil. For a typical core/bobbin configuration, the surface area can be calculated using the following guideline:

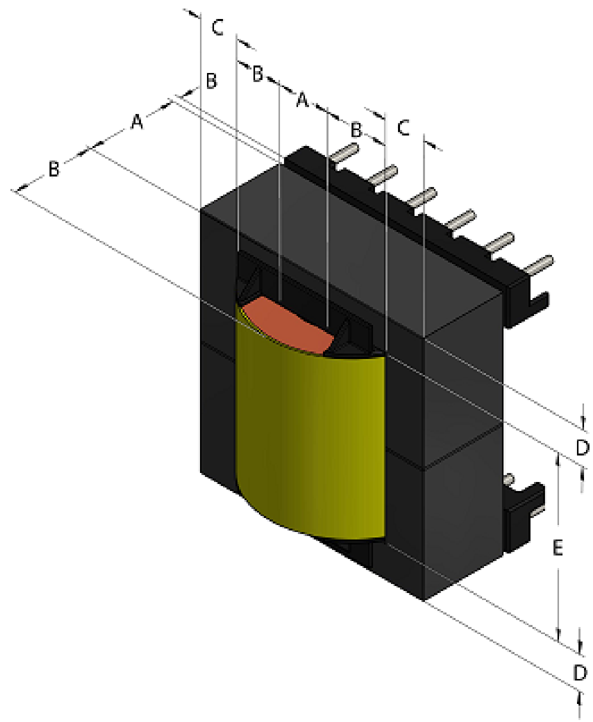


Figure 7.

SA = 2[A + 2B)(A + 2B + 2C) - 4BC +
(E + 2D)((A + 2B)- 4BD +
(E + 2D)(A + 2B + 2C)]

The minimum surface area can determined to insure that the temperature rise does not exceed 45°C with a previously estimated power loss of 16.7 W:

$$(SA) = Pt\,[mW]/\sqrt[.833]{\Delta T}$$

$$SA = 16700/\sqrt[.833]{45} = 173\ cm^2 \text{(core and coil)}$$

If I compare another, more conventional method to determine the minimum core size, the results will be quite different. For example, consider the area product (Ap) method:

$$Ap\ [cm^4] = (Po/(0.014 * \Delta\beta * f))^{1.33}$$

- Po = Output power
- 0.014 = Standard factor for this topology
- Δβ = Flux density (B_{SAT}) in teslas at 25°C
- f = Frequency in Hz (minimum specification)

$$Ap = (150/(0.014 * 0.19 * 40,000))^{1.33} = 1.58\ cm^4$$

(For comparison, the Ap of the EC60 core = 4.00 cm^4)

It is assumed that a flux density of half of the Bsat [of the core material] at maximum temperature is a reasonable starting point for the operating flux density. I will start with 0.38 teslas/2 = 0.19 teslas.

You can see from the result that the AP method would yield a core that is much smaller than the core determined by the surface area method. If I start off with the smaller core, the temperature rise would have been much higher than expected and I would have needed to switch to a larger core in the end. The minimum surface area method will save me much time in the long run.
Given the minimum surface area, I can choose an appropriate core size.

In this example, I have determined that the EC-60 core would be ideal. The core is a "standard catalog" and available from several sources. The core and printed circuit bobbin are available from stock.

Manufacturer's Data, Core and Bobbin:

- Core Cross-Sectional Area = 3.60 cm^2 minimum
- The Effective Bobbin Winding Area = 0.9 cm x 4.10 cm * 0.3 [fill factor] = 1.11 cm^2
- Ae = 3.60 cm^2 minimum
- Le = 14.2 cm
- Ve = 51.1 cm^3
- Weight = 264.5 grams
- Surface Area = 200 cm^2
- Material = Acme P4

Published saturation flux density (B_{SAT}):
- 25°C = 4800 gauss = 0.48 teslas
- 100°C = 3800 gauss = 0.38 teslas

Knowing the core parameters, I can calculate the core loss from the manufacturer's data.

Core Loss Calculation:

$$[mW/cm^3] = 1.32e - 7 * f^{1.82} * B^{2.29}$$

f = Frequency in KHz • B = mT

$$Core\ loss = 1.32e - 7 * 40^{1.82} * 190^{2.29} = 91 mW/cm^3 * 51.1\ cm^3 = 4.65\ W$$

Core Gap Considerations:

The permeability of Acme P4 material (Ui): 2500

The primary inductance with an ungapped core:

$$(Lp) = 0.4 * \Pi * N^2 * Ae/Le * Ui * 1e - 8$$
$$Lp = 0.4 * \Pi * 96^2 * 3.6/14.2 * 2500 * 1e - 8$$
$$Lp = 0.0734\ Hys$$

The required permeability @ 1100 µH (*Lp*):

$$(Ud) = Lp * Ui/Lp \, [no\ gap]$$
$$Ud = 1100e - 6 * 2500/.0734 = 37.47$$

The length of the gap:

$$(Lg) = Le * (Ui - Ud)/Ui * Ud$$
$$Lg = 14.2 * (2500 - 37.47)/2500 * 37.47$$
$$Lg = 0.37\ cm$$

Note: If the gap is divided between the outside legs of the core, the spacer would be 0.185cm thick. In this design, the gap will be ground into the center leg of the core. The core can be ground by the manufacturer to a specific inductance factor (AL). The AL value can be determined once the primary turns are determined. $AL = nH/Np^2$

Now that I know the actual surface area of the core, I can calculate the maximum power loss that can be accommodated and still keep the temperature rise below 45°C.

The total [allowable] power loss (Pt), core + copper losses:

$$(Pt) = SA * \sqrt[.833]{\Delta T}$$
$$Pt = 200 * \sqrt[.833]{45} = 16.7\ \mathrm{W}$$
$$Copper\ Losses = Pt - Core\ Loss = 12.05\ W$$

Primary Coil Design:

$$(Np) = Ep * 1e8/4.0 * f * Ae * \beta$$
$$Np = 1050e8/4.0 * 40e3 * 3.60 * 1.9e3$$
$$Np = 96\ turns$$

In order to maximize the coupling efficiency (minimize the leakage inductance), I will split the primary winding into two parts [P1 and P2]; one to be wound first and the other to be wound over the highest current secondary winding. Each primary winding will be 48 turns, to be connected in series later. This is where I need to be careful of the input voltage. At 1050 V_{PK}, 46 KHz (max), the voltage potential is high between turns and between layers. At 48 turns per winding, the voltage from turn to turn = 22 Vpk, which should not be a concern. But the voltage potential from the start (first turn) to the finish (48th turn) will be 1050 $V_{PK}/2$ = 525 V_{PK}. Also, if the winding were to occupy more than one layer, there could be high voltage potential between layers.

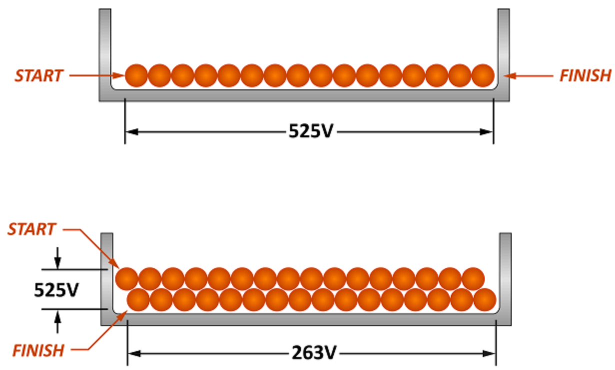


Figure 8.

Corona (partial discharge) could be a problem as the voltage is well above the typical corona inception voltage of about 300 V_{PK}.

Coil Winding Space:

The next step is to determine the maximum winding space available in the bobbin.

The total bobbin space available for winding and insulation:

$$(Aw) = 5.5 * 41.0 * 30\% = 68\ mm^2$$

I will assume that the winding space will be equally divided between the primary windings (P1 + P2) and the secondaries.

Area (P1 + P2) = 70/2 = 34 mm^2

The primary will be divided into two windings, so the cross-sectional area of one primary = 34/2 = 17 mm^2, 17 mm^2/48 turns = 0.35 mm^2/turn ~ 24 AWG equivalent wire size. I will opt for #25 AWG as I will probably need extra space for sleeving and extra insulation. I can always increase the wire size if it turns out I have extra space after the first prototype is wound.

AC vs DC Copper Losses (Rac/Rdc):

The maximum operating frequency is 46 KHz. According to the Litz wire *Technical Tips* from Kerrigan Lewis Wire Products, the ideal wire strand for 50 KHz is #38 AWG. (Rac/Rdc = 1.0) Therefore, I will strive to use #38 AWG stranded wire throughout the design. I will, however, try to stay away from actual Litz wire as it is expensive and small quantities are difficult to acquire for prototyping. Our manufacturing facility has the capacity to manufacture stranded wire.

For the primary windings, the cross-sectional area of #25 AWG = 0.25 mm^2; #38 AWG = 0.014 mm^2.

Equivalent number of strands of 38 AWG to equal 25 AWG ~ 0.25 mm^2/.014 mm^2 = 18 strands.

With the exception of winding S1, which is a low-current winding; I will use #38 AWG stranding for the primary and all secondary windings. The number of strands will be adjusted for the appropriate load current for each winding.

The following are the strands for each winding based on the rms current and the same circular-mil rating as the primary windings (200 circular-mils/amp):

S1 = Single strand of #32 AWG, triple-insulated magnet wire, diameter = 0.42 mm
S2 = Bifilar wound 60 strands of 2 x #38 AWG = 120 strands, diameter = 1.10 x 2.20 mm

S3 = Bifilar wound 25 strands of 2 #38 AWG = 50 strands, diameter = 0.60 mm

S4 = 25 strands of #38 AWG, diameter = 0.60 mm
Note: S3 and S4 will be wound on the same layer as a trifilar winding.

S5 = 25 strands of #38 AWG, diameter = 0.60 mm

Each winding will occupy only one layer. The total build up of the windings is 5.25 mm.

Core Gap, Fringing Flux and Winding Geometry:

The next step is to determine how we can best keep the winding away from the core gap. The easiest solution is to wrap multiple layers of insulation around the bobbin in order to space the windings away from the center-leg of the core. See Figures 5a and 5b on page 4.

Considering I know the space required for the winding, I can provide a maximum of 4mm of insulation under the windings.

In order to maintain the dielectric integrity between windings, 2 layers of insulation will be placed between windings, with the exception of S3 and S4 as they are trifilar wound on the same layer.

Another consideration is the creepage path along the outside edges of the windings where they encounter the bobbin flange. It is possible to experience a dielectric breakdown between an adjacent winding where they also end at the bobbin flange. To prevent the problem, I will specify tape barriers between the start and finish wires of the coil and the edge of the insulation where it meets the bobbin flange. The barriers are indicated by the orange features in Figure 10 on page 10.

Inter-Winding Insulation:

Nomex 410 was selected as insulation between winding for the following reasons: It has excellent cut-through resistance, high dielectric strength, a high temperature rating and it accepts varnish impregnation well.

Nomex 410 exhibits a stiffness that provides a flat, uniform winding surface. Its surface is slightly rough which helps secure the windings in place, and prevents the turns from sliding around. Although the material is more expensive than some other materials, the positive characteristics make it ideal for the current application.

Separation of Windings from the Core Gap:

We will wrap the coil form with 28 layers of 0.127 mm (5-mil) thick Nomex 410 before applying the first winding.
Build up of material = 3.77 mm.

Winding Order, Resistance and Copper Losses:

Note: It is good practice to wind the auxiliary (feedback) winding first to optimize the coupling between the auxiliary and the primary winding. In order to determine the resistance of the windings, I need to calculate the length of the magnet wire around the coil. To do so, I calculate the mean turn (Mt), and then factor in the number of turns and the resistance of the wire.

Winding Mean Turn:

$$(Mt) = \Pi * Coil\,for\,OD + 2 * winding\,build$$

Winding Resistance (DCR):

$$(DCR) = (Mt * N + Lead\,wire\,length) * \Omega/mm$$

Secondary (S1):

7 turns of #32 Furukawa TEX-E, triple-insulated magnet wire, wound on one layer
$Mt = \Pi * 25 + 2 * 3.77 + 0.42 = 103.5\,mm$
$DCR = (103.5 * 7 + 60) * 6.076e - 4 = 0.47\,\Omega$
$Copper\,Loss\,(I^2 DCR) = 0.25^2 * 0.47 = 0.03\,W$
Winding covered with one layer Nomex 410

Primary (P1):

48 turns or 18 x #38 AWG MW80C magnet wire, wound in one layer
Mean Turn = Π ** (33.64 + 0.55) = 108 mm*
$DCRp1 = (108 * 48 + 60) * 2.12e - 3/18 = 0.62\,\Omega$
$Copper\,Loss\,(I^2 DCRp1) = 1.45^2 * 0.62 = 1.30\,W$
Winding covered with two layers of Nomex 410

Secondary (S2):

10 turns of 2 x 60 strands of #38 AWG MW80C magnet wire, bifilar wound on one layer
Mean Turn = Π ** (36) = 113 mm*
$DCR = (113 * 10 + 60) * 2.12e - 3/2 * 60 = 0.021\,\Omega$
$Copper\,Loss\,(I^2 DCR) = 8.55^2 * 0.021 = 1.54\,W$
Winding covered with two layers of Nomex 410

Primary (P2):

48 turns of 18 x #38 AWG MW80C magnet wire, wound on one layer

Mean Turn = Π ** (38.17) = 120 mm*

$DCRp2 = (12 * 48 + 60) * 2.12e - 3/18 = 0.69\,\Omega + 0.62\,\Omega\,[DCRp1] = 1.31\,\Omega$

$Copper\ Loss\ (I^2 DCRp2) = 1.45^2 * 1.31 = 2.75\,W$

Winding covered with two layers of Nomex 410

Secondary (S5):

11 turns or 25 x #38 AWG MW80C magnet wire, wound on one layer

Mean Turn = Π ** (39.8) = 125 mm*

$DCR = (125 * 11 + 60) * 2.12e - 3/25 = 0.122\,\Omega$

$Copper\ Loss\ (I^2 DCR) = 1.70^2 * 0.122\,\Omega = 0.35\,W$

Winding covered with two layers of Nomex 410

Secondary (S3):

11 turns of 2 x 25 strands of #38 AWG MW80C magnet wire, wound on one layer, trifilar wound with S4

Mean Turn = Π ** (41.5) = 130 mm*

$DCR = (130 * 11 + 60) * 2.12e - 3/2 * 25 = 0.063\,\Omega$

$Copper\ Loss\ (I^2 DCR) = 3.25^2 * 0.063\,\Omega = 0.67\,W$

Secondary (S4):

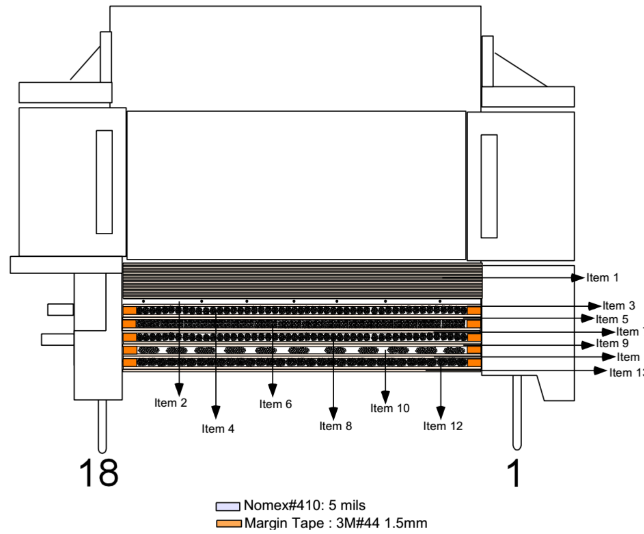
11 turns of 25 strands of #38 AWG MW80C magnet wire, wound on one layer, trifilar wound with S3

Mean Turn = 130 mm (Same layer as S3)

$DCR = (130 * 11 + 60) * 2.12e - 3/25 = 0.13\,\Omega$

$Copper\ Loss\ (I^2 DCR) = 1.10^2 * 0.13\,\Omega = 0.16\,W$

Winding covered with two layers of Nomex 410



Item	Winding (S - F)	Description	Number
1.	Insulation	Nomex#410 5milsx41.5mm width	28 Layers
2.	5 – 6	Triple Wire AWG#32	7 Ts (Total)
3.	Insulation	Nomex#410 5milsx41.5mm width	1 Layer
4.	3 – 2	18/AWG#38	48Ts (Total)
5.	Insulation	Nomex#410 5milsx41.5mm width	2 Layers
6.	18,17 – 16,15	60/AWG#38x2 BIF	10 Ts (Total)
7.	Insulation	Nomex#410 5milsx41.5mm width	2 Layers
8.	2 – 1	18/AWG#38	48Ts (Total)
9.	Insulation	Nomex#410 5milsx41.5mm width	2 Layers
10.	9 – 8	25/AWG#38	11 Ts (Total)
11.	Insulation	Nomex#410 5milsx41.5mm width	2 Layers
12.	14 – 13 12 – 11	25/AWG#38x2 25/AWG#38 Trifilar	11 Ts (Total)
13.	Insulation	Nomex#410 5milsx41.8mm width	2 Layers

Figure 9. Coil Winding Geometry and Insulation Layout

Power Losses (The Sum of Core and Copper Losses):

Copper Loss, Primary and All Secondaries = 5.50 W

Core Loss = 4.65 W

Total Losses (Pt) = 10.15 W

Knowing the core and total copper losses, I can calculate the temperature rise.

Calculated Temperature Rise:

$$\Delta T = (Pt\,[mW]/SA\,)^{0.833}$$

$$\Delta T = (10.15 * 1000/200\,)^{0.833} = 26.3°C$$

Specification	Value	Tolerance	Measurements
Primary Inductance (Lp):	1100 µH	5%	1121 µH
Lp Shift from 25 to -40°C	2%	Max	1.25%
Leakage Inductance (Primary to S2):	40 µH	Max	25.7 µH
Leakage Inductance (S2 to S3 & S5 shorted):	3 µH	Max	1.12 µH
Dielectric strength (Hi-Pot):			
Primary and S1 to All Secondaries	3750 V_{AC}	< 1 mA Leakage	< 1.0 mA
Primary to S1	1500 V_{DC}	< 0.5 mA Leakage	< 0.5 mA
Between All Secondaries	1500 V_{DC}	< 0.5 mA Leakage	< 0.5 mA
Turns Ratio:			
Primary to S1	13.7:1	2%	13.70
Primary to S2	9.6:1	2%	9.60
Primary to S3	8.73:1	2%	8.73
Primary to S4	8.73:1	2%	8.73
Primary to S5	8.73:1	2%	8.73
Winding Resistance (DCR):			
Primary	1.60 Ω	Max	1.400 Ω
Secondary (S1)	0.550 Ω	Max	0.440 Ω
Secondary (S2)	0.027 Ω	Max	0.023 Ω
Secondary (S3)	0.082 Ω	Max	0.070 Ω
Secondary (S4)	0.162 Ω	Max	0.140 Ω
Secondary (S5)	0.160 Ω	Max	0.134 Ω
Temperature Rise:			
400 V Input, 80°C Ambient	45°C	Max	25.4°C
1100 V Input, 80°C Ambient	45°C	Max	29.3°C

Figure 10. Test Results from First Prototype Build

Partial Discharge (Corona) and the Importance of Vacuum Varnish Impregnation:

Corona is an electrical discharge accompanied by ionization of the air surrounding a conductor. Corona discharge acts like small electrical sparks. Each spark, or discharge, is the result of an electrical breakdown of an air pocket within the insulation. These discharges erode insulation and could eventually result in insulation failure.

Electrodes may consist of two wires, separated by either air space or an insulator; between wires and a metal terminal and/or wires and a magnetic core.

Corona can be present within the voids (air pockets) in an insulator such as epoxy, varnish or other porous material.

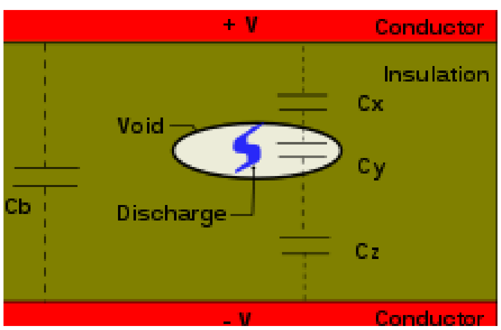


Figure 11. Corona Discharge in Air Pocket.

Corona inception voltage is the lowest voltage at which corona occurs. It can start as low as 300 V_{PK}. The inception voltage does not vary with frequency, but the life expectancy of materials under corona discharge is inversely proportional to frequency. This effect becomes more critical as the operating frequencies of power supplies increase as the technology advances.

To put the relationship between corona, frequency and the life of the product into proper perspective, please consider the following: A transformer that exhibits a slight corona may function 50 years at 60 Hz before the insulation breaks down enough to fail under the voltage stress. That same transformer will last only about five weeks when operated at 50 KHz and about three weeks at 100 KHz.

Corona effects can show up later. A transformer can go through production testing, final inspection and the customer's incoming inspection and finally burn-in, and then fail in the field a month later -- if the transformer is not corona free.

The best practice to ensure that the transformer remains corona free is to ensure that the varnish impregnation is performed under vacuum.

If there are no voids, or air pockets, within the impregnation, corona will not be generated. It is recommended that the varnish be exposed to a minimum of 29 inches of mercury (982 millibars) of vacuum during impregnation.

References:

1. Ruben, L., and Stephens, D. *Gap Loss in Current-Limiting Transformer. Electromechanical Design*,

April 1973, pp. 24-126.

2. F. W. Peek, Jr., *Dielectric Phenomena in High Voltage Engineering*, April 1915

3. Kerrigan-Lewis Wire Products, *Litz Wire Technical Tips*

4. Colonel William T. McLyman, *Transformer and Inductor Design Handbook*, 1978

5. Mil-PRF-27G, *Performance Specification, Transformers and Inductors (Audio, Power and High-Power Pulse), General Specification for. PERFORMANCE SPECIFICATION*, May 19, 2014

6. Texas Instruments, Application Report SNVA716, Improving the Performance of Traditional Flyback Topology With Two-Switch Approach, July 2014; http://www.ti.com/lit/an/snva716/snva716.pdf