Switch Mode Power Supply Transformer Design

Switch mode power supply (SMPS) transformers are commonly used in a regulated power supply and function to step up or step down voltage or current, and/or provide isolation between the input and output side of a power supply. On the primary side of a switch mode transformer, the duty cycle (on-time) of the input voltage waveform is varied (switched) to deliver a constant output voltage under varying load conditions. SMPS transformers are designed to operate at a specific frequency, typically between 10 kHz to 500 kHz. SMPS transformer power levels now extend into the 50 kW range.

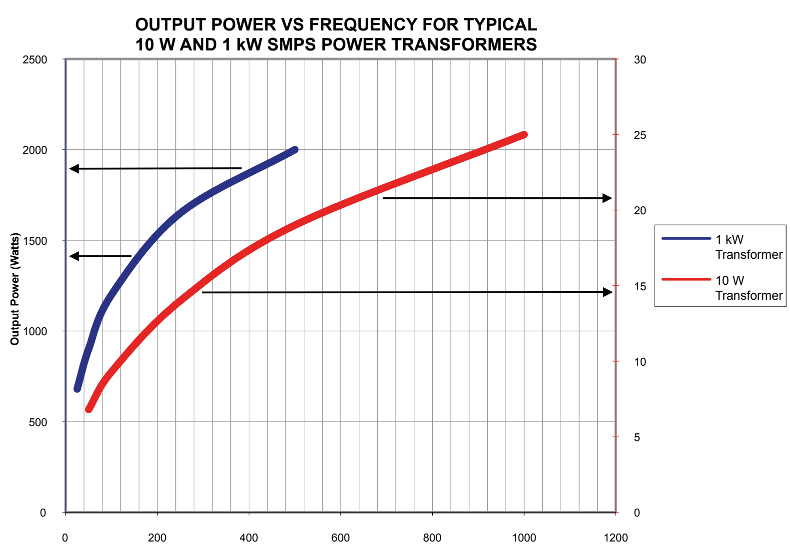
Choosing a Core and Bobbin Geometry

The first step in designing an SMPS transformer is to choose a core and bobbin geometry. Most practical SMPS transformer designs are derived from standard geometries offered by Ferroxcube and other manufacturers. The decision of which geometry to choose is a function of many interrelated variables, and the discussion that follows will assist in this process.

The amount of power deliverable through a given SMPS transformer is a function of output power and frequency. As the frequency increases, greater power can be accommodated with a given transformer core and bobbin, or possibly a smaller geometry core can be used. Figure A shows the typical RMS power available from a 10 W and a 1 kW SMPS transformer assuming the temperature rise of each transformer is fixed at 40° C.

The scaling effect shown in Figure A is driven by the fact that the gauss level in a given winding is inversely proportional to the volt microsecond product applied to the primary side of the transformer. As the frequency of operation increases the gauss level decreases. This permits fewer turns and less magnetic core cross section. An increase in volumetric cores losses and AC copper losses at higher frequencies act to mitigate the benefits of a lower gauss level and this accounts for the flattening of the curves seen in Figure A.

FIGURE A



For the high currents typical of low voltage applications above 2.5 kW, it becomes increasingly difficult to manage the AC winding losses. Compounding this problem, scaling effects dictate lower power densities for both the core and the bobbin, to prevent excessive temperature rise. Therefore higher power SMPS applications typically operate at lower frequencies. To some degree the problem of managing high frequency high current density power has been mitigated for the transformer designer because silicone devices such as IGBT's have current and frequency limits below what transformers can accommodate today.

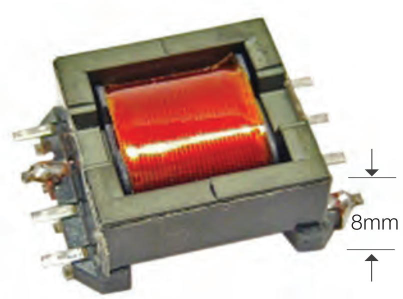
There are many successful designs being done today in the 500 kHz to 1 MHz range, but the designer must be aware that there are a few more hurdles to this type of design. First, the most complex part of the design (AC copper losses) becomes a very important factor at these higher frequencies. Second, capacitive effects and winding parasitics begin to play a very significant role, and conventional design rules may no longer apply. Finally, while the ferrite manufacturers have developed excellent materials for this frequency range, the high frequency cores are not always available through distribution.

Surface Mount Device (SMD) vs Printed Circuit Board (PCB) vs Chassis Mount

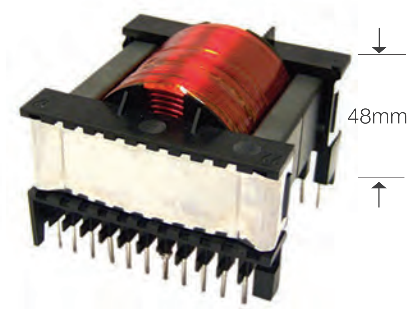
Figure B shows an SMD, a PCB and a chassis mount transformer. Each of these mounting styles has some unique advantages and disadvantages which will determine their suitability for a given application.

FIGURE B

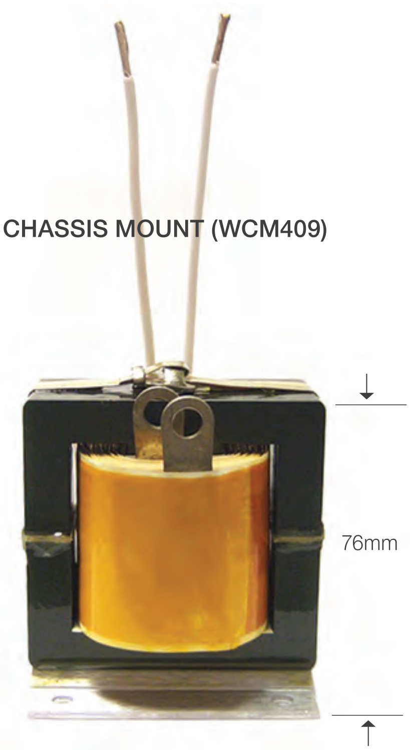
SMD (WCM402)



PCB (WCM408)



CHASSIS MOUNT (WCM409)



TRANSFORMER TABLE

Product Code	*Description*	*Weight*
WCM 402 EFD 15	10 W Transformer	5.50 g
WCM 408 ETD 54	1000 W Transformer	0.35 kg
WCM 409 E 73	5000 W Transformer	1.90 kg

Core and bobbin geometries used for SMD transformers were developed originally for low voltage applications, and are typically chosen in applications of 100 kHz and up with power levels up to approximately 250 W. The window for SMD transformers is typically low profile, and there is not much clearance to the core, so 750 V of Vout is a practical upper limit. SMD cores are not typically suitable for isolation transformers requiring safety agency approval because input to output creepage and clearance values are typically limited to 4 mm and it is difficult to achieve hypot values greater than 2500 Vac.

continued →

PCB mount geometries provide many more choices than SMD geometries. There are PCB mount geometries which will handle up to 2.5 kW of power. The diverse selection of PCB mount cores and bobbins makes this style of transformer suitable for a wide range of output voltages (up to 3000 V). Isolation values up to 4000 Vac hypot and 10 mm creepage and clearance are achievable. But each core geometry is different with some being more suitable for isolation than others.

Once the power level goes above 2.5 kW the designer is typically driven into a chassis mount style of transformer. With a chassis mount transformer, very high isolation is achievable. Also, it is possible to accommodate high output voltages of up to 5000 V. Unlike conventional PCB mount transformers based on available core and bobbin geometries, with a chassis mount transformer the limit on high voltage output is typically not the creepage and clearance inherent in the geometry, but rather the capacitance of the high voltage winding at SMPS frequencies.

TABLE 1.0

COMPARISON OF SMD, PCB AND CHASSIS MOUNT TRANSFORMER GEOMETRIES

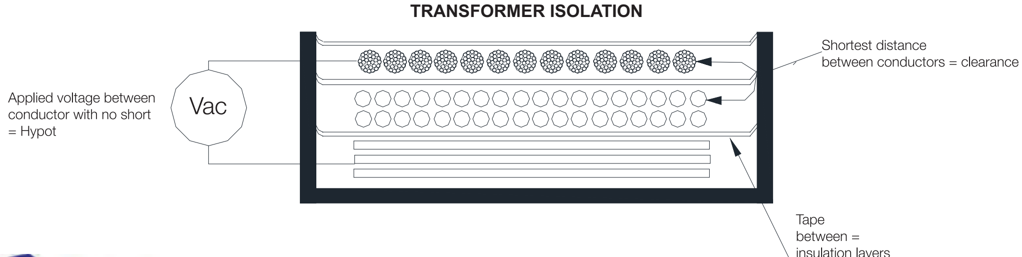
Product	*Power*	*Voltage*	*Frequency*	*Safety Isolation*
SMD	1 W to 200 W	up to 750 V	100 kHz to 1 MH z	Low voltage, low power only
PCB	1 W to 2500 W	up to 3000 V	10 kHz to 500 kHz	Yes
Chassis Mount	1000 W and higher	Any	5 kHz to 500 kHz	Any

Degree of Isolation/Safety Agency Approval

The isolation between the input and output of a transformer is determined by three factors. These three factors are: the creepage and clearance between the windings, a dielectric test typically in the form of AC or DC hypot, and the class of insulation between non-insulated conductors. They are illustrated in Figure C below.

FIGURE C

TRANSFORMER ISOLATION



The amount of isolation required depends on the power level, the specific use of the transformer, and the voltage levels on each winding. Temperature rise and frequency of operation are variables that are completely independent of the three safety factors noted above, and are typically not part of an agency approval process, provided that the temperature rise of the transformer remains within the acceptable use limits of the insulating materials. There is no trade-off between these three variables, and along with higher creepage and clearance values, the designer is faced with a higher minimum dielectric test and a more robust insulation requirement.

The lowest level of isolation is typically defined as operational. At this level, the only concern is that the transformer operates with minimal leakage current between the input and output and with no failures. A 5 W transformer stepping down 12 V to 3.3 V would typically only require sufficient insulation to keep a short from forming under normal operation, and a 250 Vac hypot would be adequate to verify the integrity of the inter winding insulation. At the other extreme, say for example a transformer delivering power to a surgical application in which patient isolation is required, the designer may be faced with 10 mm minimum creepage and clearance, a 4000 Vac hypot, and reinforced insulation.

When high levels of isolation are required, it is important to de-rate the power ratings typically supported by a given geometry to allow for more room in the winding window to support the required insulation and creepage/clearance. Table 2.0 shows the power available from the WCM 408 ETD 54, 1 kW core geometry with and without safety isolation.

TABLE 2.0

TYPICAL RMS OUTPUT POWER (W) vs FREQUENCY WCM 408 ETD 54 TRANSFORMER

Frequency	25 kHz	50 kHz	100 kHz	250 kHz	500 kHz
Operational Isolation	680	900	1200	1650	2000
Safety Grade Isolation	410	540	720	990	1200

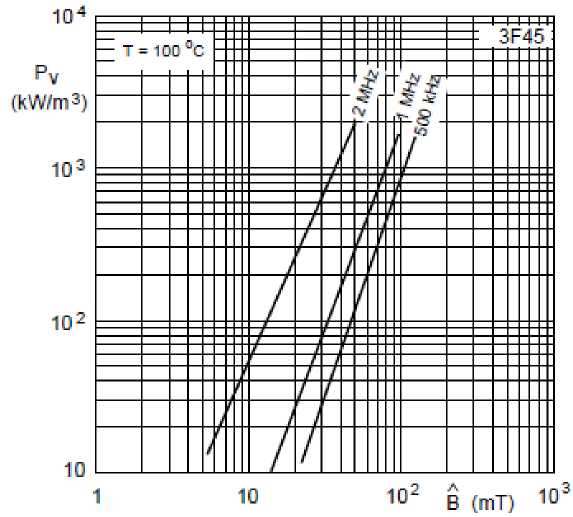
Choice of a Core Material

SMPS transformers are typically not saturation limited, but they are loss limited. Most all SMPS transformers use manganese zinc ferrite cores because this family of magnetic materials has the lowest losses at SMPS frequencies, and is relatively low cost. Manganese zinc cores typically have a saturation flux density of 4000 to 5000 gauss, and SMPS transformers are typically designed to operate between 500 and 2500 gauss. Even if most SMPS transformer designs use a manganese zinc ferrite, within this family of materials there is considerable choice available to the designer. Manganese zinc ferrites are simply not all the same and it is necessary to understand their properties in order to make an informed choice for any SMPS design.

Figure D, on the next page, is a plot of core loss density as a function of frequency and gauss level for Ferroxcube's 3F45 material. This exciting new material has excellent loss characteristics for operating frequencies above 500 kHz.

FIGURE D

SPECIFIC POWER LOSS
AS A FUNCTION OF PEAK FLUX DENSITY
WITH FREQUENCY AS A PARAMETER

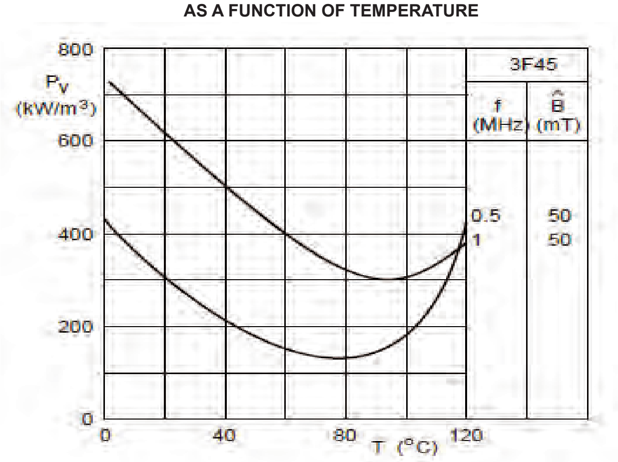


Every successful SMPS transformer design will use this figure, or one similar, as a basis for estimating core losses under continuous operating conditions. This figure relating core loss density to flux density and to frequency should be used to determine cores losses with every design.

It is quite possible to complete a design using only the information shown in the core loss curve above, but it is important to pay attention to some other important properties of the core. One important property is the variation in core losses with temperature. Figure E illustrates how losses vary with temperature for 3F45 material.

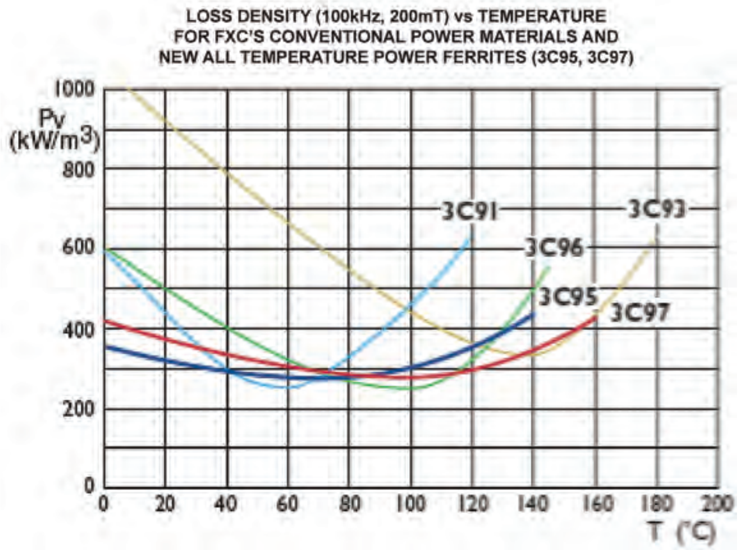
FIGURE E

SPECIFIC POWER LOSS
FOR SEVERAL FREQUENCY/FLUX DENSITY COMBINATIONS
AS A FUNCTION OF TEMPERATURE



Historically most transformers have been designed to run at a temperature of 40 to 60° C above ambient and therefore it was useful to minimize losses in the 60 to 100° C operating temperature range. Indeed this is the case for 3F45 material as it is apparent that the losses are the lowest at all frequencies in the 80 to 100° C range. New materials such as Ferroxcube's 3C95 and 3C97 material are designed to have lower losses over a much wider temperature range and therefore are suitable for applications where transformer efficiency is important. Figure F shows loss vs temperature for Ferroxcube's new 3C95 and 3C97 materials, compared to more convential materials.

FIGURE F



Another family of properties that can be important relates to variations in permeability for the core material. For all ferrite cores the permeability of the core varies measurably with changes in core excitation and with changes in temperature. The variation with temperature is shown in Figure G. The variation with peak flux density is shown in Figure H, on the next page. For most SMPS transformers this is not a concern, however it can be a factor in some designs.

FIGURE G

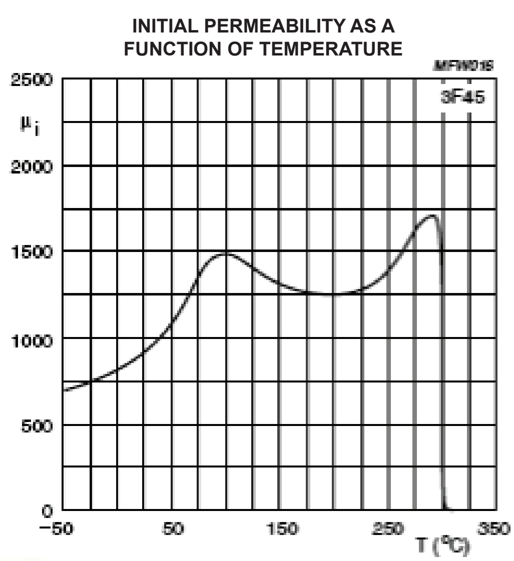


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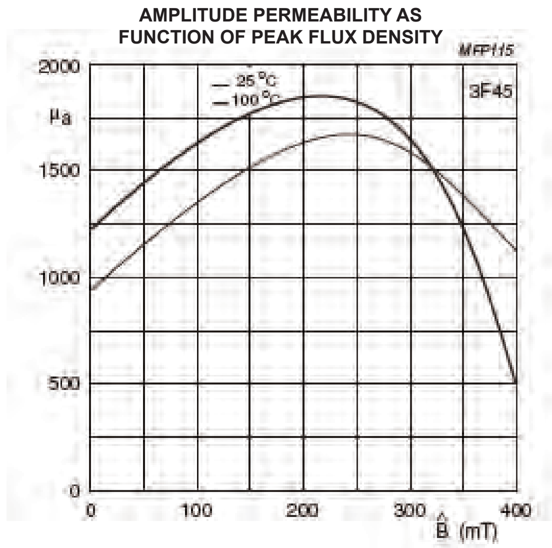
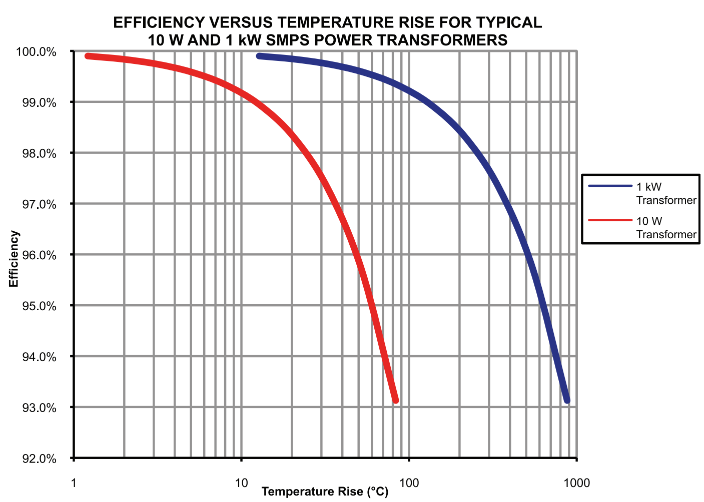
FIGURE H



Efficiency and Allowable Temperature Rise

Modern SMPS transformers are highly efficient. With a very high power to size ratio, they have to be efficient or they would simply get too hot. Figure I illustrates overall transformer temperature efficiency vs temperature rise for our 10 W (WCM 402 EFD 15) and 1 kW (WCM 409 ETD 54) transformers. From this figure it is obvious that a practical SMPS transformer design will result in an efficient device. SMPS transformers can normally be designed to operate at efficiencies exceeding 99%, however, at this very high efficiency level, there is typically a cost and sometimes a size penalty.

FIGURE I



Many transformer designs done today are still based on a 40° C temperature rise above ambient with no forced air cooling. At a 40° C temperature rise the transformer can generally be counted on to generate a hot spot temperature well within the operating limits of the materials of construction and to run at 95% efficiency or greater. Forced air cooling, done correctly, will allow an SMPS transformer to run at significantly higher loss levels (and reduced efficiency) without overheating.

Determining the Coefficient of Thermal Resistivity

The coefficient of thermal resistivity for an electronic device is the temperature rise per unit of loss, expressed in °C per W. For large magnetics this expression has typically been expressed as a function relating the losses and the core surface area to the temperature rise. While this method of generating temperature rise is useful, and reasonably accurate, it is less useful for SMPS designs since the surface area of core and bobbin combinations is not published.

A simpler method requires only the core volume, and this parameter is widely available in all the core manufacturer's data sheets. Ferroxcube conducted an empirical study comparing the hot spot temperature rise of many different core and bobbin geometries to the total core and winding losses. The results of this study indicate that the core volume is a reasonable proxy for determining the coefficient of thermal resistivity. Ferroxcube's curve fit their data to come up with the following very useful expression, which is graphed in Figure J.

Equation 1:

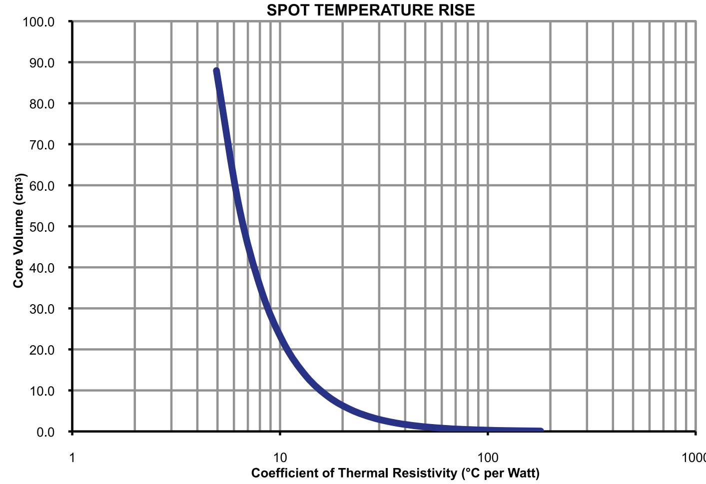
$Rth = 53*(Ve)^{-0.53}$

Where Rth = coefficient of thermal resistivity in °C per W

Ve = core volume in cubic centimeters

FIGURE J

RELATIONSHIP OF FERRITE CORE VOLUME TO TRANSFORMER HOT SPOT TEMPERATURE RISE



It should be noted that this is an approximation only, and it does sidestep some important factors including air flow conditions around the transformer, varying rates of heat transfer from the transformer to its surroundings at different ambient temperature and differing ratios of transformer surface area to core volume. But the author can verify that it is adequate as a first level approximation for most SMPS designs.

Specifying Primary Turns

Once a core geometry and core material has been chosen the next step is to determine the primary turns. The number of turns is determined from the following expression which relates the core flux density to the number of primary turns, the ET (volt microsecond) product, and the core cross sectional area. This expression, when used with the units specified, is also directly comparable to the loss curves published by the core manufacturers.

Equation 2:

$$B = \frac{V(ton)(10^8)}{2AeN}$$

Where

B = peak AC flux density (gauss)

V = primary voltage (volts)

ton = primary switch on time (seconds)

Ae = core area (cm^2)

N = number of primary turns

While it is possible to simply choose a core and the number of turns for the primary and back into the losses using the core manufacturer's loss curves, there is a faster method than this iterative process. With the coefficient of thermal resistivity known for the core, the designer will first choose an acceptable temperature rise and then calculate the allowable transformer losses. If you are designing to an efficiency specification, then you will already know the acceptable transformer losses. Budget half of your total acceptable losses as core losses, and from this loss number and the core volume you can determine the core loss density in mW/cm^3. Once the core loss density is known, go directly to the core manufacturer's loss curves (Figure D or Figure K) to determine the gauss level, and from this determine the primary turns.

Typically the minimum input voltage, maximum power and continuous duty condition will yield the highest overall transformer losses. Therefore, this is the design condition that is recommended for completing the first pass of an SMPS transformer design. However, it is important to check the design under the high input voltage condition because this case can yield higher overall losses depending on the circuit topology and the balance chosen by the designer between core and copper losses.

Determination of Primary Turns

Design Example:

- Offline 48 V, 100 W Power Supply
- Core and Bobbin: WCM 402 EFD 30
- Vin = 85 Vdc to 170 Vdc
- Max on-time = 70%
- Frequency = 100 KHz

FIGURE K

SPECIFIC POWER LOSS AS FUNCTION OF PEAK FLUX DENSITY WITH FREQUENCY AS A PARAMETER

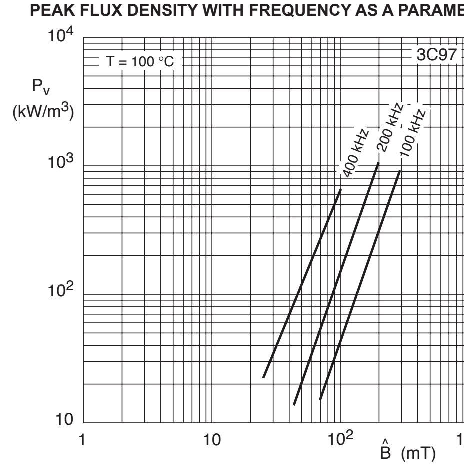


TABLE 3.0

WCM 402 EFD 30

CORE AND BOBBIN PARAMETERS

CORE MATERIAL: 3C97

Product Code	*Core Volume (cm^3)*	*Core Area (cm^2)*	*Magnetic path length (cm)*	*Bobbin window width (in)*	*Bobbin window height (in)*	*Mean length per turn (mm)*
EFD 30	4.70	0.69	6.8	0.80	0.10	52

Equation 3:

- Determine Rth for the core from Equation 1:

$$Rth = 53\,(Ve)^{-.53}$$
$$= 53\,(4.7)^{-.53} = 23.3°\text{ C per W}$$

- Allow 40° C T rise and determine total (core and copper) acceptable losses:

$$\text{Total losses} = \frac{40°\text{ C}}{23.3°\text{C/W}} = 1.72\text{ W}$$

- Budget half of the total for core losses and determine core loss density:

$$\text{Core loss density} = \frac{\text{Core losses (mW)}}{\text{Core Volume (cm}^3)} = \frac{1720/2}{4.70} = 183\text{ mW/cm}^3$$

- Go to manufacturer's loss curves to determine B at 100 kHz. (See Figure K)

183 mW/cm^3 @ 100 kHz = 1700 gauss

- Set the number of primary turns from Equation 2:

$$N = \frac{V\,(ton)(10^8)}{2(Ae)(B)} = \frac{85(\frac{0.7}{100{,}000})(10^8)}{2(0.69)(1700)} = 25\text{ Turns}$$

Secondary Turns

Unlike linear transformers, an SMPS transformer is duty cycle regulated, and as a result the turns ratio is determined from the input-output transfer function for the indicated topology. Many transformer designers will simply divide Vout by Vin, add 10% and use this for a turns ratio and this works quite well. However a little more information will enable the designer to make a more informed and superior choice for the turns ratio. Each transformer topology has a unique Vin to Vout transfer function. All of these transfer functions relate the input voltage, the output voltage, the turns ratio, and the primary side voltage on time.

Most topologies are duty cycle limited to 50% maximum on-time, although the forward converter and the flyback can accommodate duty cycles in excess of 50%.

Regulation will only apply to one secondary output so, if multiple outputs are specified, there are some important design factors to keep in mind. First, a topology should be chosen that permits good cross regulation between windings. Second, the turns ratio must be chosen carefully to insure that each output voltage comes as close as possible to the target output.

Appendix A, page 15, shows a number of different topologies and the input-output transfer function for each.

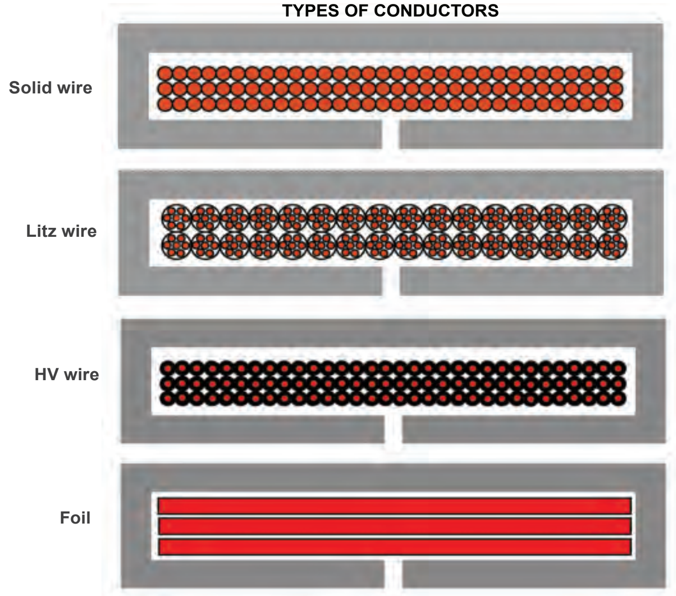
continued →

Windings

Once the core geometry has been chosen and the number of primary and secondary turns have been determined, a conductor must be chosen for each winding, and winding losses must be determined. There are four basic classes of windings, shown in Figure L, to choose from and the final choice will depend on the specific requirements of each design.

FIGURE L

TYPES OF CONDUCTORS



Solid wire…is typically used for transformers operating at 100 kHz and less, and RMS current levels under 20 amps. The main advantage of solid wire is low cost, both in material cost, and time to wind. The main disadvantage is losses. Solid wire can generate excessive AC winding resistance at SMPS frequencies. Solid wire windings in an SMPS transformer should always be kept to a single layer if possible.

*Litz wire…*consists of multiple fine strands of insulated solid wire twisted together to form a single conducting bundle. It is available in most all gauges with individual strands from 24 awg all the way up to 50 awg. At SMPS frequencies most designers use 36 to 44 awg stranding, with 36 to 40 awg being used at frequencies under 100 kHz and 42 and 44 awg used at higher frequencies.

HV wire…safety isolation style wire. This wire is characterized by 2 or 3 very thin insulating layers of teflon or similar polymer. Manufacturers include Tex-E and Rubadue. The main advantage of this wire is that it is possible to achieve a 4000 Vac hypot and reinforced insulation rating with a wire that has a relatively modest increase in overall diameter compared to conventional magnet wire. This type of wire vastly simplifies the build of the transformer as no additional insulation or creepage and clearance needs to be built into the winding. The disadvantages are the same disadvantages as litz, namely higher cost and lower winding window utilization.

continued →

*Copper foil…*is a very good choice for low voltage windings. A copper foil winding typically has far greater window utilization and typically has the lowest DCR of any other alternative, and copper foil can be the lowest loss alternative. The inherently low profile cross section of a copper foil winding helps to minimize copper losses as the magnetic field inherent in a transformer winding tends to draw AC current to the top and bottom of the winding. As a disadvantage, each turn is a single layer, so with multiple turns winding capacitance and AC copper losses start to add up quickly as turns increase.

Determining Winding Resistance

Winding losses in an SMPS transformer consist of both AC and DC losses. A first order approximation of copper losses can be determined by calculating the DC resistance of each winding and multiplying this by Irms squared, again for each winding. The sum of the losses for each winding yields a loss approximation which can be reasonably accurate for some designs. Equation 4 is used to determine DC resistance for each winding in a transformer.

Equation 4:

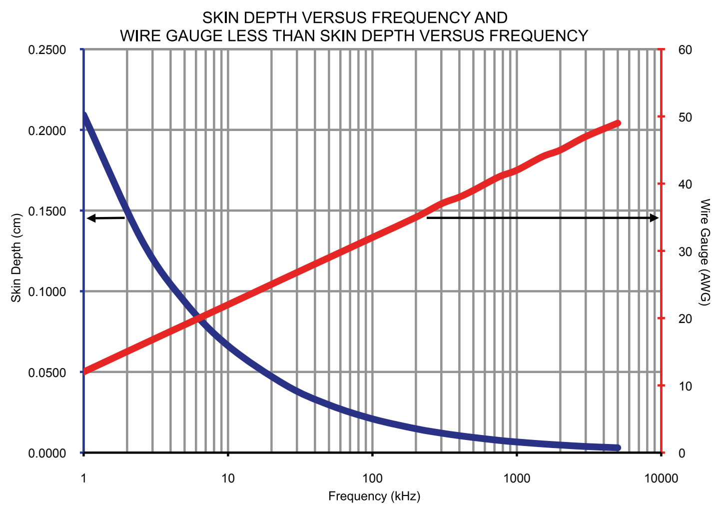
DCR = N* (mtl)* (Conductor resistance)
Where:
DCR = winding resistance (Ohms)
N = number of turns
mtl = mean turn length for the bobbin being used (ft)
Conductor resistance = (Ohms/1000 ft)*

Using the DC resistance (calculated from Equation 4) and the RMS current to predict copper losses results in one minor and one major source of error:

The minor source of error is that the mean length per turn for the bobbin is not the same as the average turn length for each winding. Depending on the style of bobbin, the turn length can vary considerably between windings. That said, when the windings have similar current densities this source of inaccuracy tends to balance out because windings placed first on the bobbin have lower resistance and windings placed last have higher resistance than the resistance predicted by Equation 4. In effect, the total DC copper resistance for the transformer turns out about right.

The major source of inaccuracy in estimating copper losses using DC resistance and RMS current is that this method fails to account for AC winding losses due to proximity effects. AC winding losses are caused by the tendency of current to concentrate on the surface of a conductor in the presence of the strong MMF field inside every transformer. Unfortunately the degree to which this field is concentrated on the surface depends on the frequency, with greater current concentration occurring at higher frequencies The skin depth of a conductor has been arbitrarily defined as the depth at which the current density has fallen to 37% of its value on the surface. Skin depth vs frequency is plotted in Figure M, on the next page.

FIGURE M



It is clear from this chart that at SMPS frequencies, AC copper resistance must be managed carefully in any SMPS transformer design. By way of example, at 250 kHz, with a skin depth of 0.0148 cm, a designer would need to chose 35 awg wire to keep the copper cross section at one skin depth or lower. Clearly a single strand of 35 awg will not carry much current, so litz wire is required in most cases.

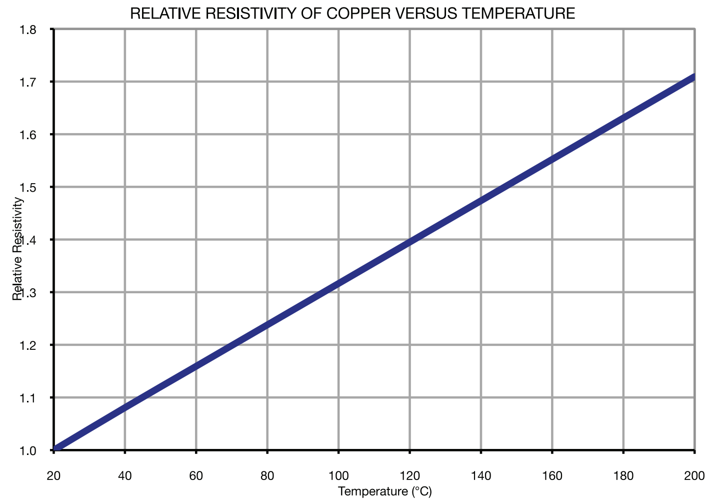
AC losses frequently exceed DC copper losses and can overwhelm the design if they are not managed properly. It is beyond the scope of the application note to present techniques for predicting AC copper losses since they are quite complex. Some rules of thumb to minimize AC copper losses include:

1. Minimize the number of layers in each winding (one is great, but not always possible.)
2. Interleave primaries and secondaries if possible
3. Keep the single layer copper conductor diameter to one skin depth or less, windings with more than one layer need even smaller diameter conductors
4. Use litz or copper foil for windings of two or more layers or at 100 kHz and higher.

The Thayer School of Engineering at Dartmouth maintains an excellent on-line freeware tool for predicting copper losses in SMPS transformers with litz wire windings.
It can be found at: *http://engineering.dartmouth.edu/inductor/programs.shtml*

*Note: Conductor resistance does vary significantly with temperature. If you look at Figure M, next page, it is apparent that copper resistance will increase by 8% from 20 to 40° C, by 16% from 20 to 60° C and by 24% from 20 to 80° C. This increase in resistivity corresponds directly to the same percentage increase in AC and DC copper losses. Fortunately, unlike AC copper losses, it is relatively easy to predict this effect by using Figure N. For temperature rise limited designs a separate calculation of core and copper losses and the associated device temperature rise should be performed at the predicted worst case operating condition.

FIGURE N



West Coast Magnetics Data Sheets

In order to expedite the choice of a core, WCM has data sheets available on our website which show the power handling capability of many different cores as a function of frequency. This information is based on a temperature rise of 40° C, for a conventional push pull transformer with minimal parasitic losses. If your application demands high isolation, very high efficiency, or has an output or input voltage of more than 200 V, you may have to choose a bigger size. *Click here to connect to the WCM switch mode transformer page online.*

Design Help

There is a sound option to designing your own transformer. WCM is set up to complete a design quickly and easily at minimal cost to you. All you need to do is go to our website and you can choose your own core and bobbin geometry and schematic. We will take care of the rest. WCM has completed hundreds of designs for many satisfied customers. We will deliver a specification sheet to you for certifying the key properties of the transformer(s) you require. *Click here to contact WCM for design help.*

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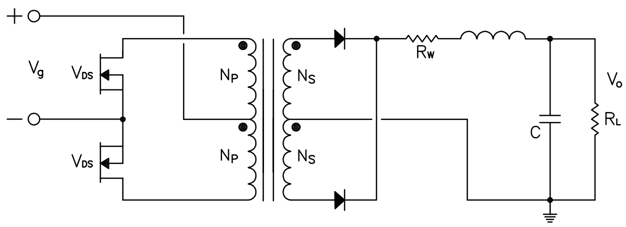
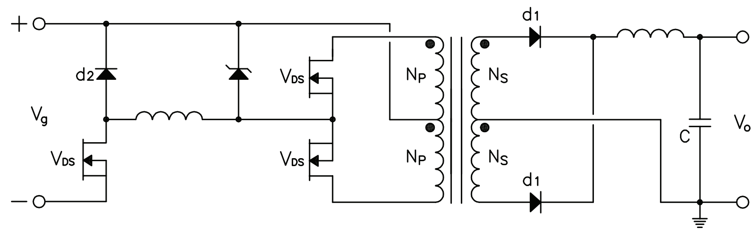
Appendix A… CONVERTER SCHEMATICS

PUSH PULL CONVERTER



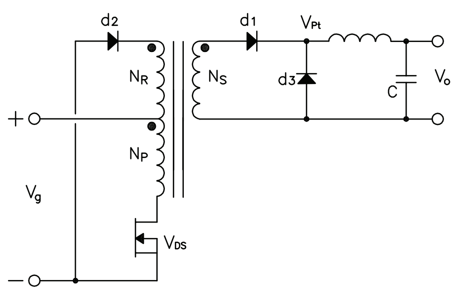
TURNS RATIO: $\frac{N_S}{N_P} = \left[V_o\left(\frac{R_L + R_W}{R_L}\right) + V_{d1}\right]\left(V_g - V_{DS}\right)D$

CURRENT FED PUSH PULL CONVERTER



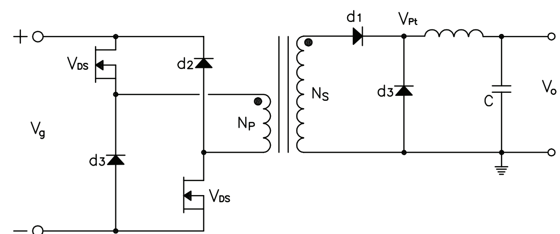
TURNS RATIO: $\frac{N_S}{N_P} = \frac{\left(V_o + V_{d1}\right)\left(T - T_d\right)}{\left(V_g - 2V_{DS}\right)T_{on} - \left(V_{DS} + V_{d1}\right)\left(T - T_{on}\right)}$

FORWARD CONVERTER



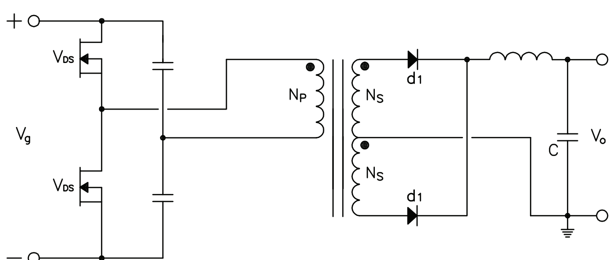
TURNS RATIO: $\frac{N_S}{N_P} = \left(\frac{V_o + V_{d1}}{V_g - V_{DS}}\right)\frac{T}{T_{on}}$

DUAL SWITCH FORWARD CONVERTER



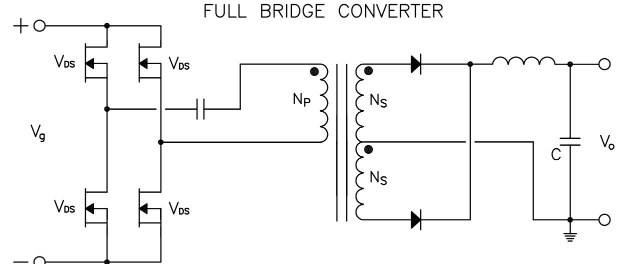
TURNS RATIO: $\frac{N_S}{N_P} = \left(\frac{V_o + V_{d1}}{V_g - 2V_{DS}}\right)\frac{T}{T_{on}}$

HALF BRIDGE CONVERTER



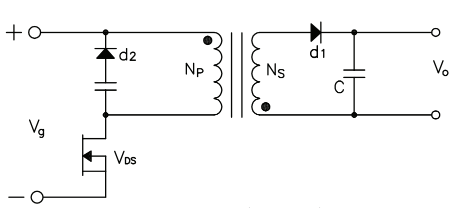
TURNS RATIO: $\frac{N_S}{N_P} = \left(\frac{V_o + V_{d1}}{V_g/2 - V_{DS}}\right)\frac{T}{T_{on}}$

FULL BRIDGE CONVERTER



TURNS RATIO: $\frac{N_S}{N_P} = \left(\frac{V_o + V_{d1}}{V_g - 2V_{DS}}\right)\frac{T}{T_{on}}$

FLYBACK CONVERTER



TURNS RATIO: $\frac{N_S}{N_P} = \left(\frac{V_o + V_{d1}}{V_g - V_{DS}}\right)\frac{T - T_{on}}{T_{on}}$

Code	*Description*	*Code*	*Description*
V_g	*input voltage (Volts)*	D	*duty cycle =T_{on}/T (seconds/seconds)*
V_o	*output voltage (Volts)*	R_L	*load impedence (Ohms)*
V_{DS}	*drain to source voltage voltage drop across the transistor (Volts)*	R_W	*winding resistance (Ohms)*
V_{d1}	*voltage across output diode d_1 (Volts)*	N_S	*number of turns on the secondary winding*
T	*period of the cycle (seconds*	N_P	*number of turns on the primary winding*
T_{on}	*on time of the cycle (seconds)*	N_R	*number of turns on the reset winding*
T_d	*delay time (seconds)*		