

SMPS Components and their effects on System Design

The choice of components used in a switch mode power supply can greatly affect the performance and reliability of a design. Transistors are not ideal switches, the capacitors behave like resistors, and the inductors and transformers can transform themselves into “shorts”.

The following slides will provide some insight into how transistor, diode, inductor, transformer, and capacitor characteristics can affect the overall design of SMPS applications.

Session Agenda

- **MOSFET characteristics**
- **Diode characteristics**
- **Capacitor issues**
- **Magnetic components**
- **Printed circuit board recommendations**

This is the agenda for this course.

The various loss mechanisms for MOSFETs will be discussed, as well as secondary issues such as the “Miller Effect”.

The loss mechanisms for diodes will be described, and recommendations for diode selection will be made.

Capacitors, which never function like ideal devices, will be discussed. Their characteristics are closely intertwined with SMPS circuit design.

The behavior of magnetic components, such as transformers and inductors, can vary widely. We will cover some of the major issues.

And last but not least, are recommendations for printed circuit board (PCB) design to enhance SMPS system performance.

MOSFET Losses

simplified

$$P_{\text{loss}} = P_{\text{conduction}} + P_{\text{switching}}$$

$$P_{\text{conduction}} = (I_D)^2 \cdot R_{\text{ds}(\text{hot})} \cdot D$$

$$P_{\text{switching}} = (V \cdot I_D / 2) \cdot (T_{\text{on}} + T_{\text{off}}) \cdot F_{\text{sw}} + (C_{\text{oss}} \cdot V^2 \cdot F_{\text{sw}})$$

Where: D = Duty cycle, F_{sw} = Switching Frequency

C_{oss} = Transistor Output Capacitance

The MOSFET is a common choice as the “switch” for smps applications, they are easy to drive and have fast switching times.

The power dissipated in a MOSFET is the conduction losses and the switching losses. The conduction loss is the I^2R loss. When reading MOSFET device specifications, use the $R_{\text{DS(on)}}$ specification at the expected operating junction temperature. The R_{ds} value is highly temperature dependent.

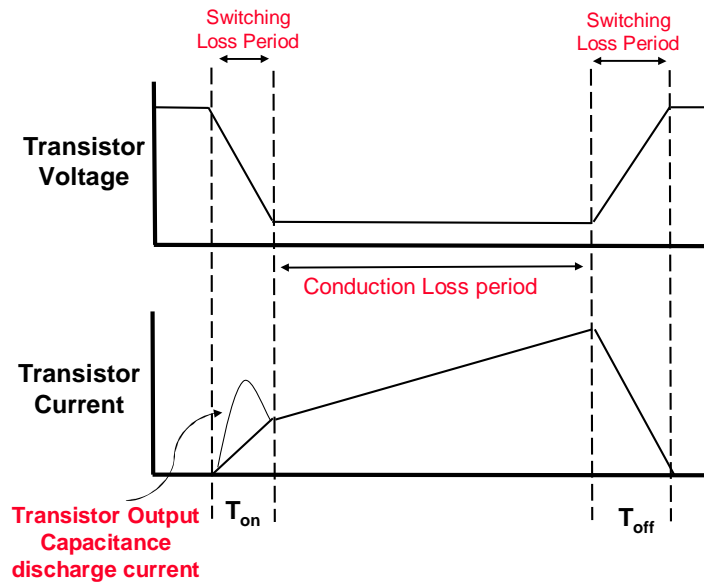
The switching losses are primarily due to the “Cross Over” losses that occur because the switching time for a transistor is non zero. As a transistor is turning on or off, the current is flowing through the transistor, and the voltage across the transistor is transitioning between the supply voltage and the device’s saturation voltage. The energy dissipated is equal to the average applied voltage times the average current times the time spent in the switching process. Reducing the switching time reduces switching losses.

Additionally, there are losses due to the stored charge in the transistor’s output capacitance (C_{oss}) that is discharged when the transistor is turned on. Usually, the losses due to the output capacitance are small.

There is another loss not shown on this slide associated with switching, the gate drive circuit. Power is dissipated in the FET gate driver to charge and discharge the MOSFET gate capacitance. In most applications, the gate driver loss is not significant.

Switching losses are directly proportional to switching frequency.

MOSFET Waveforms



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SMPS Components and their Affects on System Design

Slide 4

This slide shows the simplified voltage and current waveforms of a MOSFET switching between the ON and OFF states.

The conduction losses are lower in magnitude but typically persist for a longer period of time. For MOSFETs, the conduction losses are I^2R , while a Bipolar transistor, or a IGBT (Insulated Gate Bipolar Transistor) have conduction losses in the form of $(I \bullet V_{sat})$ where V_{sat} is the transistor's saturation voltage.

The primary switching loss is the voltage and current "Cross Over" points during the turn ON and OFF process. The $(I \bullet V)$ product is large but if the switching times and switching frequencies are kept low enough then the switching losses will be tolerable.

This slide also shows the Output Capacitance Losses (drawn in red) that are experienced during the transistor turn on.

MOSFET Losses

example assumptions

Assume:

I_D	= 10 amps
R_{ds}	= 0.15 ohm
V	= 120 volt
T_{on}	= T_{off} = 50 nsec
F_{sw}	= 250 KHz
D	= 0.5
C_{oss}	= 185 pf

This slide and the next show an example MOSFET switching loss calculation. The numbers are chosen from real transistors with voltages and currents of reasonable magnitudes.

MOSFET Losses

example calculations

$$P_{\text{conduction}} = (10)^2 \cdot 0.15 \cdot 0.5 = 7.5 \text{ watts}$$

$$P_{\text{switching}} = (600 \cdot 100\text{ns} \cdot 250\text{Khz}) + \\ (120^2 \cdot 185 \text{ pf} \cdot 250\text{Khz})$$

$$P_{\text{switching}} = 15 \text{ watts} + 0.66 \text{ watts}$$

$$P_{\text{loss(total)}} = 7.5 + 15 + 0.66 = 23.16 \text{ watts}$$

The conduction loss is 7.5 watts, and the switching losses are 15.66 watts. The switching losses associated with the transistor's output capacitance is a small percentage of the total switching loss, about 4%.

The total MOSFET dissipation is 23.16 watts. This device will obviously require significant heat sinking to extract the heat out of the device to keep the MOSFET junction temperature within allowable limits.

Duty Cycle Effects MOSFET Losses

$$P_{\text{conduction}} = (I_D)^2 \cdot R_{\text{ds}} \cdot D$$

Assume: $I_L = 1$ amp, $R_{\text{ds}} = 1$ ohm, $I_L = I_D / D$

$$\text{If } D = 1.0, \text{ then } P = (1)^2 \cdot 1 \cdot 1.0 = 1 \text{ Watt}$$

$$\text{If } D = 0.5, \text{ then } P = (2)^2 \cdot 1 \cdot 0.5 = 2 \text{ Watt}$$

$$\text{If } D = 0.25, \text{ then } P = (4)^2 \cdot 1 \cdot 0.25 = 4 \text{ Watt}$$

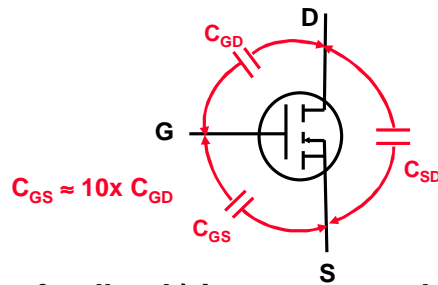
In a power supply circuit, assuming a constant average output power, if the duty cycle is halved, then the current must be doubled. If the duty cycle is reduced to one quarter, then the current must be increased 4x. The product of ($I \cdot D$) remains constant.

For MOSFETs, the conduction loss is proportional to the current squared. As the duty cycle is reduced, the power dissipation is rising with the square of the current but the power dissipation is only being reduced at a linear rate by the decreasing duty cycle. The net result is that MOSFET power dissipation will rise linearly as the duty cycle decreases.

In Buck or Boost converters, there is little choice in the duty cycle values. In these applications the duty cycle is fixed by the input to output voltage ratio.

In transformer coupled converter topologies (Push-Pull, Forward, Bridge, etc) the transformer turns ratio can provide the needed input to output voltage conversion. If the transformer ratios are selected carefully, then the duty cycle can be chosen to be higher rather than lower. If the dutycycle is near 100% (1.0) then the lowest possible conduction losses can be obtained.

The Miller Effect and MOSFETs



- The Miller Effect (capacitive feedback) is represented by C_{GD} and it opposes the gate driver circuit.
- At low voltages (<50V), it is not usually a concern.
- At high voltages (>150V), it becomes a serious issue. The miller effect can overwhelm the FET driver.
- Bridge transistor configurations are susceptible of having the “OFF” transistor turn “ON” with catastrophic current “Shoot-Thru”.

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Slide 8

The “Miller Effect” is one those things you probably learned about in college, but then quickly forgot ! The Miller Effect describes the feedback capacitance that exists in most active devices such a Transistors, MOSFETs, vacuum tubes, etc..

In MOSFETs, the feedback capacitance is relatively small compared to the input capacitance. In low voltage applications, the miller effect is usually not an issue. But at higher voltages, the stored charge on the feedback capacitance may become greater that the stored charge on the Gate-Source capacitance.

A Classic Scenario: Someone debugs a new power supply at lowered voltages to check out the MOSFET gate waveforms. They all look great. Then the supply voltage is raised upward to normal levels..

... BAM .. In an instant the transistors are smoked...

The solution is to use strong enough FET gate drivers with good wiring practices. Make sure the FET drivers have adequate voltage supply bypassing, and the connection between the driver's common (ground) connection to the MOSFET's source terminal are short and wide. Minimize PCB trace resistance and inductance.

MOSFET Recommendations

- **Select FETs with lowest $R_{ds(on)}$ to reduce resistive losses**
- **Provide ample FET drive to minimize FET switching time to reduce switching losses and Miller effects**

Select the MOSFETs with the lowest $R_{ds(on)}$ specification that your budget allows to minimize heat sink costs and size.

Make sure that the FET drivers can provide the current needed to drive the MOSFET gate capacitance to insure short switching times. If the application involves switching higher voltages, increase the FET driver capability to insure that the miller effect does not overdrive your gate driver and turn on or off your transistor at a bad time...

Diode Losses simplified

$$P_{\text{loss}} = P_{\text{conduction}} + P_{\text{reverse}} + P_{\text{switching}}$$

$$P_{\text{conduction}} = (I_f \cdot V_f \cdot D)$$

$$P_{\text{reverse}} = (I_r \cdot V_r \cdot (1-D))$$

$$P_{\text{switching}} = (T_{rr} \cdot I_{rrm} \cdot V_r \cdot F_{sw}) / 2$$

Where:

D = Duty cycle, F_{sw} = Switching Frequency
 V_f = Forward voltage drop, I_f = Max forward current
 I_r = Reverse leakage current, V_r = Reverse voltage
 T_{rr} = Reverse recovery time, I_{rrm} = Max Recovery current

Diodes (Rectifiers) are not “ideal” devices when operated in SMPS applications. Diodes suffer from three sources of losses: Conduction loss is due to the forward voltage drop of the diode. Switching losses are due to the fact that a diode does not switch from conducting state to a non conducting state in zero time. Power diodes (Rectifiers) also conduct current when reversed biased. This reverse leakage is very temperature dependent

The conduction losses are proportional to current and the forward voltage drop of the diode. Choose diodes with low forward voltage drop specifications.

The reverse leakage losses tend to be small, but in high voltage application, the leakage current times the applied voltage product may become appreciable.

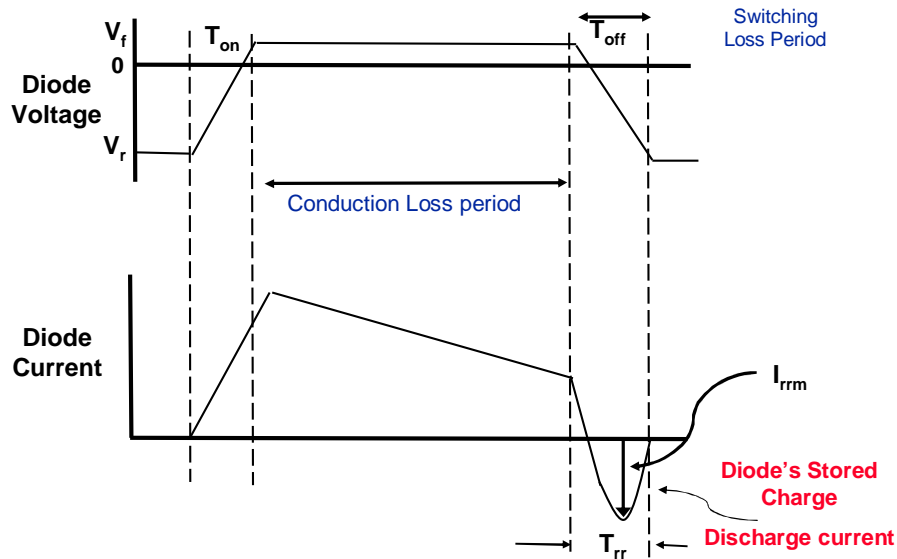
The switching losses become a real issue at high switching frequencies. The diodes’ stored charge must be swept out before the diode turns off.

For low voltage applications, Schottky diodes feature both a low forward voltage drop, and no switching losses. Schottky diodes do not suffer from a stored charge. Schottky diodes are usually limited to application below 150 volts.

For high voltage applications use “Ultra Fast” rectifiers, or Silicon Carbide rectifiers.

Never, ever use general purpose (60 Hz) rectifiers for SMPS.

Diode Waveforms



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Slide 11

This slide shows the diode's current and voltage waveforms in an SMPS application.

Usually, the turn on switching losses are not significant. But the turn off losses can become a serious issue. Different diode vendors specify the stored charge in different ways. Some vendors specify a reverse recovery current (I_{rrm}), while other vendors specify a Reverse Recovery charge (Q_{rr}).

The diode conduction losses can become difficult to handle at high current levels. And in low voltage applications, the diode's forward voltage drop can represent a large fraction of the total power supply losses.

In fact, most SMPS vendors now use a MOSFET as a synchronous rectifier in low voltage applications to minimize the conduction losses associated with most rectifiers.

Diode Losses

example #1 calculations
(Ultra Fast Diode)

$$P_{\text{loss}} = P_{\text{conduction}} + P_{\text{reverse}} + P_{\text{switching}} = 9.98 \text{ w}$$

$$P_{\text{conduction}} = (16 \cdot 0.8 \cdot 0.5) = 6.4 \text{ w}$$

$$P_{\text{reverse}} = (0.00025 \cdot 100 \cdot 0.5) = 0.0125 \text{ w}$$

$$P_{\text{switching}} = (34 \text{ ns} \cdot 4.2\text{A} \cdot 100\text{V} \cdot 500 \text{ KHz}) / 2 = 3.57 \text{ w}$$

Where:

D	= 0.5	F _{sw}	= 500 KHz
I _{load}	= 8 amps	V _f	= 0.80 v
I _f	= 16 amps	I _r	= 250 uA
V _r	= 100V	T _{rr}	= 34 nsec
I _{rrm}	= 4.2 amp		

This slide presents an example power loss calculation for an UltraFast diode. UltraFast diodes typically have switching times of approximately 35 nsec.

The conduction loss is the dominant switching loss for this example at 6.4 watts.

The reverse leakage loss is insignificant.

The switching loss is 35% of the total loss.

Diode Losses

example #2 calculations

Schottky Diode

$$P_{\text{loss}} = P_{\text{conduction}} + P_{\text{reverse}} + P_{\text{switching}} = 5.55 \text{ w}$$

$$P_{\text{conduction}} = (16 \cdot 0.69 \cdot 0.5) = 5.20 \text{ w}$$

$$P_{\text{reverse}} = (0.007 \cdot 100 \cdot 0.5) = 0.35 \text{ w}$$

$$P_{\text{switching}} = (0 \text{ ns} \cdot 0 \text{ A} \cdot 100 \text{ V} \cdot 500 \text{ KHz}) / 2 = 0 \text{ w}$$

Where:

D	= 0.5	F _{sw}	= 500 KHz
I _{load}	= 8 amps	V _f	= 0.69 v
I _f	= 16 amps	I _r	= 7 mA
V _r	= 100V	T _{rr}	= 0 nsec
I _{rrm}	= 0 amp		

This slide show an example power loss calculation for a Schottky diode.

The Schottky has lower conduction losses than an UltraFast diode, and it has no switching losses!

The total losses for the Schottky is about half of the loss for an UltraFast diode.

Schottky diodes typically cost more than UltraFast diodes.

Diode Losses

example #3 calculations
(a Good Fast Recovery Diode)

$$\begin{aligned}
 P_{\text{loss}} &= P_{\text{conduction}} + P_{\text{reverse}} + P_{\text{switching}} &= 49.6 \text{ W} \\
 P_{\text{conduction}} &= (16 \cdot 1.2 \cdot 0.5) &= 9.6 \text{ W} \\
 P_{\text{reverse}} &= (0.007 \cdot 100 \cdot 0.5) &= 0.0125 \text{ W} \\
 P_{\text{switching}} &= (160 \text{ ns} \cdot 10 \text{ A} \cdot 100 \text{ V} \cdot 500 \text{ kHz}) / 2 &= 40 \text{ W}
 \end{aligned}$$

Where:

$$\begin{aligned}
 D &= 0.5 & F_{\text{sw}} &= 500 \text{ kHz} \\
 I_{\text{load}} &= 8 \text{ amps} & V_f &= 1.2 \text{ v} \\
 I_f &= 16 \text{ amps} & I_r &= 5 \text{ mA} \\
 V_r &= 100 \text{ V} & T_{\text{rr}} &= 160 \text{ nsec} \\
 I_{\text{rrm}} &= 10 \text{ amp}
 \end{aligned}$$

This slide shows an example power loss calculation using a “Fast” rectifier. A Fast rectifier is much faster switching than a standard 60Hz rectifier, but much slower than an UltraFast diode. A Fast rectifier has a switching time of about 200 nsec.

The conduction loss is slightly higher than an UltraFast diode, but the switching losses (40 watts) are too much to handle.

At modern SMPS pwm switching frequencies, the turn off time for a Fast diode is a significant portion of the entire pwm cycle period !

Diode (Rectifier) Recommendations

- **Use Schottky rectifiers for low voltage (<100 V) applications to reduce conduction and switching losses.**
- **Use Ultra-Fast rectifiers for higher voltage applications.**
- **Consider Silicon Carbide rectifiers for high voltage applications. (higher cost, higher performance).**
- **DO NOT use general purpose or “fast” recovery rectifiers (1N4002, 1N493x, etc).**
“smoke em if you got em”

For applications below 100 volts, it is recommended to use Schottky rectifiers.

For higher voltage applications, the primary choice are UltraFast rectifiers. Also consider using Silicon Carbide rectifiers for high voltage applications.

Never use general purpose or fast diodes except in very low frequency SMPS applications (< 10 KHz). At higher frequencies, they cease to function as rectifiers, and effectively become low ohmage resistors.

Capacitor Losses simplified

$$P_{\text{esr}} = (I_{\text{ripple}})^2 \cdot \text{ESR}$$

Where: ESR = Effective Series Resistance

- **Select low ESR Capacitors to promote long capacitor life by minimizing internal heating.**
- **Every 10°C reduction in capacitor temperature doubles capacitor life.**

The dominant loss for capacitors used in SMPS applications is due to the ESR (Effective Series Resistance). The ESR is due to the physical construction of a capacitor including the resistance of the internal interconnect, and the behavior of the electrolyte or other material used as the insulator between the capacitor plates. These internal processes that impede the flow of current into and out of the capacitor are modeled as a resistor in series with an ideal capacitor.

Just like a real resistor, the ESR will dissipate heat in the presence of a current flow into or out of a capacitor. The power dissipated is proportional to the square of the current flow.

Many capacitors are very sensitive to high temperature. Electrolytic capacitors will dry out at an accelerated rate at high temperatures. Film capacitors may experience reduced voltage tolerance at high temperatures, and ceramic capacitors may crack with repeated temperature cycling when soldered to a PCB with a different thermal coefficient of expansion.

Output Capacitor ESR Affects

$$V_{\text{ripple}} = (I_{\text{ripple}} \bullet \text{ESR}) + (I_{\text{load}} / F_{\text{sw}} \bullet C)$$

Where: ESR = Effective Series Resistance

- **Select low ESR Capacitors to minimize output voltage ripple.**

The ESR of a capacitor is often the dominant specification when selecting capacitors for SMPS applications, even more important than the Capacitor's capacitance !

In virtually all SMPS designs, the output capacitor is responsible for supplying the power to the load during a portion of the pwm cycle. The capacitance value must be larger enough to keep the voltage "Droop" during a pwm cycle to acceptable limits. This voltage "droop" or "sag" is one source of the output voltage ripple.

The other source for output voltage ripple is the capacitor's ESR. As the capacitor is charged and discharged, the capacitor current times the ESR yields a voltage ripple that is added to the ripple caused by the capacitor's "droop".

Capacitor Recommendations

- **Select capacitors with low ESR at switching frequency**
- **Select capacitors rated for high ripple current. (SMPS rated)**
- **DO NOT use 50/60 Hz filter or audio application capacitors.**

Select capacitors for SMPS applications based on their ESR specification to minimize output ripple.

Select capacitors with low ESR that are rated for the expected ripple current. The ripple current will heat the capacitor.

Capacitors designed for general bypass or audio applications are typically not designed to handle large ripple currents. These capacitors will not provide good performance or reliability.

Select capacitors rated for elevated temperatures (105 °C) to promote longer capacitor life.

Capacitor Reliability

- **Electrolytic Capacitors: typically rated for 1 – 5K hours at rated temperature (< 1 year).**
- **Derate capacitor voltages to extend life, especially at high temperature (electrolytic and Film). A 15% reduction increases life by 50%.**
- **Select low ESR to reduce internal temperature rise. Life doubles for each 10 °C reduction.**

Electrolytic capacitors are typically rated for 1,000 to 5,000 hours of life at rated voltages and temperatures. That is less than one year !

The operating temperature must be derated to account for internal capacitor heating due to ripple currents. High temperatures cause the electrolyte to dry out. Then capacitor's ESR will rise, and the heating process accelerates.

High voltages can also cause a capacitor to fail more quickly. Operating the capacitor at lower temperatures and/or selecting capacitors with low ESR to reduce self heating will greatly extend the life of electrolytic capacitors. Reducing the operating voltage relative to the specified voltage rating for the capacitor also increases capacitor life.

Polymer film capacitors lose much of their voltage capability at high temperatures.

Ceramic capacitors can suffer cracking during large temperature excursions when mounted to PCBs due to differences in thermal coefficients of expansion.

Magnetic Properties

$$H = (0.4 \cdot \pi \cdot N \cdot I) / L$$

$$\beta = E \cdot 10^8 / (4 A \cdot N \cdot F_{sw})$$

Where:

hysteresis loop

β = Flux density (Gauss),

H = Magnetization force (Oersted),

N = Number of turns,

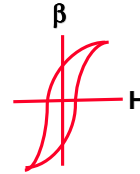
F_{sw} = Switching Frequency

I = current (amps),

L = magnetic path length (cm)

E = Applied voltage (rms),

A = Cross sectional area (cm²)



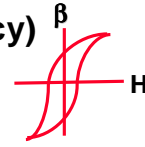
The magnetization force (h) is the magnetic field generated in response to a current flowing through a conductor. The magnetization force is proportional to the current flow and the number of turns of wire and inversely proportional to the length of the magnetic path.

The flux density (β) is the number of magnetic lines of force through a given area. The flux density increase with applied voltage to the windings, decreases with the area through which the flux passes, and the flux decreases with increasing frequency.

The β H curve is called the "Hysteresis Loop" and it traces the flux density versus the applied magnetic force. The surface of the β H plane represents energy. The area inside the hysteresis loop represents energy loss.

Magnetic Properties

(size, power, and frequency)



- Magnetic materials have limited flux capacity
- Exceeding flux capability: Magnetic Saturation \Rightarrow Catastrophic Failure
- For a given frequency, Higher power requires larger cross sectional area magnetic cores
- Higher frequency reduces flux density – smaller cores possible

Notice the hysteresis curve flattens out as the magnetization force H gets larger in magnitude. This flattening is called Saturation. As the magnetic material saturates, the device incorporating the material, such as a transformer or inductor ceases to function.

To prevent the flux density β from saturating, either the magnetic core can be made larger in area, or the frequency can be increased.

As the operating frequency is increased, the energy losses tend to increase. To operate at higher frequencies requires different magnetic materials.

Magnetic Component Losses

simplified

$$P_{\text{magnetic}} = P_{\text{hysteresis}} + P_{\text{eddy current}} + P_{\text{winding}}$$

$$P_{\text{hysteresis}} \approx k_h \cdot V_c \cdot F_{\text{sw}} \cdot (\beta_{\text{max}})^{2.5}$$

$$P_{\text{eddy current}} \approx k_e \cdot V_c \cdot (F_{\text{sw}})^2 \cdot (\beta_{\text{max}})^2$$

$$P_{\text{winding}} \approx \underbrace{I^2 \cdot R_{\text{DC}}}_{\text{DC winding resistance losses}} + \underbrace{I^2 \cdot R_{\text{DC}} \cdot \sqrt{k \cdot F_{\text{sw}}}}_{\text{AC winding resistance losses (Skin Effect)}}$$

The losses in magnetic components are primarily due to three issues: hysteresis losses, eddy current losses, and losses in the windings that surround the magnetic core.

The hysteresis loss is due to the resistance of the flipping of the magnetic domains in the material.

The magnetic core is structurally similar to a shorted conductor loop passing through the magnetic field. Think of a iron toroidal transformer: the iron core is conductive, and a conductor placed in an alternating magnetic field will have an induced current flow. The induced current in the core will dissipate energy. To reduce eddy current losses, core materials have been developed that suspend magnetic particle in a binder with high resistivity such as epoxy. For SMPS applications, the most common magnetic materials are Ferrites.

The windings used in magnetic components for SMPS applications experience losses due to winding resistance (I^2R losses), and the skin effect.

The term “skin effect” refers to the process where a conductor’s current flow becomes restricted to the conductor’s surface at increasing frequencies. This current crowding near the surface increases the effective resistance of the winding.

To counter the skin effect, single conductor windings are often replaced with parallel windings of smaller gauge. Often “Litz” wire is used.

Magnetic Component Loss Issues

- Losses increase rapidly with Frequency
- Losses increase with Flux density
- Hysteresis losses dominate below 200 kHz
- Eddy current losses dominant at higher frequencies
- Eddy current losses increase with (voltage)²
- Magnetic losses often dominate system losses

For a given magnetic core material, core losses rapidly increase with frequency.

Core losses increase with flux density, but increasing frequency reduces flux density. At higher frequencies, core saturation becomes less of a problem.

Below 200 KHz, hysteresis losses are the dominant loss mechanism.

Above 200 KHz, eddy current losses begin to dominate. Eddy current losses increase with the square of the applied voltage.

In SMPS applications, the volts per turn applied to a transformer or inductor is also related to the dutycycle required to achieve the desired output voltage. When calculating needed turns ratio, optimize for higher pwm duty cycles.

The choice and design of the magnetic components can define the success or failure of a SMPS application because the losses associated with transformers and inductors can be high at high frequencies.

Magnetic Component Issues

- **Magnetic Saturation – Flux density exceeds material capability – Inductors and Transformers cease to function.**
- **Flux Imbalance (Flux Walking) – small DC magnetic offsets add up over time to create magnetic saturation.**

Magnetic saturation is an important consideration in the design and use of smps applications.

Consider both the average (DC) current flow, the ripple current, and potential fault currents when sizing the inductor.

In transformers, the same considerations apply as in inductors, plus the issue of flux imbalance must be considered. Flux imbalance or flux walking is the gradual accumulation of a DC bias.

For example: In a Push-Pull or Bridge converter, the current flow through one set of transistor(s) versus the other transistor(s) may match to 0.00001 %. But with an operating frequency of hundreds of Kilohertz, the flux imbalance will accumulate very quickly.

To prevent flux imbalance, circuit techniques such as adding a small series capacitor with the transformer will block any DC component.

Alternatively, control techniques, such as current mode control, can be used to monitor the current flow through the transformer, and the pwm signal can be modified to balance the average currents to prevent any DC bias.

Magnetic Component Recommendations

- **Insure inductors are rated for maximum peak currents (including faults) to prevent saturation.**
- **Design circuits to prevent issues with Flux Imbalance (Flux Walking) to prevent saturation.**
- **Select magnetic materials designed for your switching frequency.**

Make sure the inductors and transformers are sized to carry the expected peak currents to prevent core saturation.

When designing SMPS applications, begin with the idea of preventing flux walking by applying current control techniques, or adding capacitors to block DC components.

Select inductors and transformer materials that are designed to operate at your chosen operating frequency.

Printed Circuit Board Recommendations

- **Design PCB with ample ground planes under control logic and power circuitry.**
- **Minimize current path “Loop area” on PCB.**
- **Use short, wide traces to minimize resistance and inductance.**
- **Use Heavy gauge copper plating (2 oz, 3 oz, ...) on PCBs.**
- **Use multiple vias for high current traces (5 per amp).**

The printed circuit board (PCB) is not just a framework to hold your components. The high currents and high frequencies encountered in modern SMPS circuits require that a lot of thought be used in designing the PCB.

SMPS PCB design is not like designing a typical digital circuit PCB that is auto routed. The copper layers that carry power must be thick enough to carry the current. All traces that connect to the power transistors, transformers, inductors, and capacitors must be designed to minimize trace inductance.

If the high power current flows between different PCB layers, make sure that multiple vias are used to stitch the traces on the layers together.

Avoid creating traces and component layouts that create large area current loops among the power components.

Place gate drivers close to the MOSFETs, making sure the trace from driver return (GND) terminal to the MOSFET source is short, direct, and wide.

Printed Circuit Board More Recommendations

- **SMPS switching frequencies are now in the “AM” radio band (500 – 1500 kHz). The harmonics extend well into the shortwave band.**
- **Treat SMPS PCB design like “RF”. The FCC doesn’t like unlicensed “Transmitters”.**

It is strongly suggested that multilayer PCBs be used with ground planes to reduce radiated EMI, and to reduce transients that may affect the operation of the MOSFETs.

Many modern high performance DC/DC converters use six to eight layer PCBs !

Often, the PCB is becoming a component in the design of the Transformer ! The traces on the PCB layers are stitched and stacked to create windings, and then a two piece magnetic core is clamped around the PCB to create the transformer. This approach saves a lot of labor winding transformers.

The pwm frequencies are now commonly in the AM broadcast band. With the sharp waveforms, harmonics now extend well into the Shortwave broadcast bands. Meeting FCC radiated emission requirements more difficult. Multilayer PCBs with ground planes help the issue a lot.

Prototyping SMPS Recommendations

- **Never prototype with wire-wrap boards.**
- **Never prototype with “plug-in” boards.**
- **Do not use sockets for MOSFETs.**

It is recommended that printed circuit boards (PCBs) be developed for prototyping.

An alternate method called “Dead-Bug” construction can be used for investigative efforts in a lab. The dead-bug approach uses a ground plane such as an unetched pcb, and the component's ground leads are soldered to the ground plane, and other signals are wired point to point (pin to pin). This approach is not practical anything more complicated than a Buck converter.

Wire wrap boards and “proto” plug in boards will not work because of the high inductance and high resistance of the connections. Also, the conductors on the proto boards can not carry any significant current.

Key Support Documents

<u>Device Selection Reference</u>	<u>Document #</u>
General Purpose and Sensor Family Data Sheet	DS70083
Motor Control and Power Conv. Data Sheet	DS70082
dsPIC30F Family Overview	DS70043
Base Design Reference	Document #
dsPIC30F Family Reference Manual	DS70046
dsPIC30F Programmer's Reference Manual	DS70030
MPLAB® C30 C Compiler User's Guide	DS51284
MPLAB ASM30, LINK30 & Utilities User's Guide	DS51317
dsPIC® Language Tools Libraries	DS51456

For more information, here are references to some important documents that contain a lot of information about the dsPIC30F family of devices.

The Family Reference Manual contains detailed information about the architecture and peripherals, whereas the Programmer's Reference Manual contains a thorough description of the instruction set.



Key Support Documents

Microchip Web Sites: www.microchip.com/smps
www.microchip.com/16-bit

For device data sheets, Family Reference Manuals, and other related documents please visit the following Microchip websites.

Thank You

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