

# TOPSwitch® Power Supply Design

## Techniques for EMI and Safety

### Application Note AN-15



Offline switching power supplies have high voltage and high current switching waveforms that generate Electromagnetic Interference (EMI) in the form of both conducted and radiated emissions. Consequently, all off-line power supplies must be designed to attenuate or suppress EMI emissions below commonly acceptable limits.

This application note presents design techniques that reduce conducted EMI emissions in TOPSwitch power supplies below normally specified limits. Properly designed transformers, PC boards, and EMI filters not only reduce conducted EMI emissions but also suppress radiated EMI emissions and improve EMI susceptibility. These techniques can also be used in applications with DC input voltages such as Telecom and Television Cable Communication (or Cablecom). Refer to AN-14 and AN-20 for additional information. The following topics will be presented:

- EMI Specifications for North America, European Community, and Germany
- Measuring Conducted Emissions with a LISN
- Peak, Quasi-Peak, and Average Detection Methods
- Safety Principles
- EMI Filter Components
- Flyback Power Supply EMI Signature Waveforms
- Filter Analysis
- Power Cord Resonances
- Transformer Construction Techniques
- Suppression Techniques
- General Purpose TOPSwitch EMI Filters
- EMI Filter PC Layout Issues
- Practical Considerations

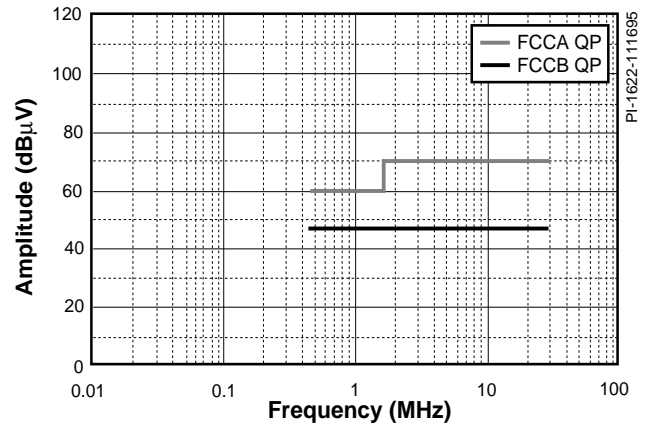


Figure 1. FCC Class A and B Limits (Quasi Peak).

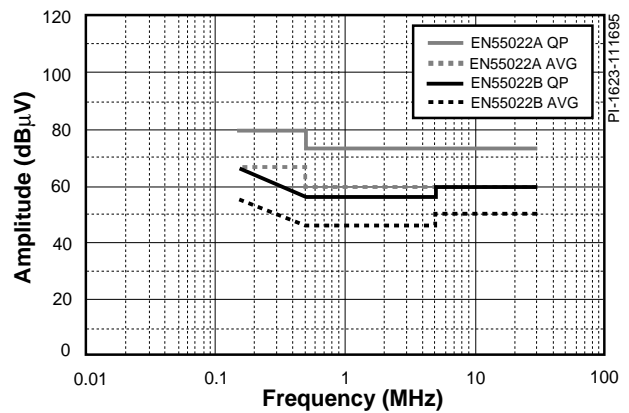


Figure 2. EN55022 Class A and B Limits (Average and Quasi Peak).

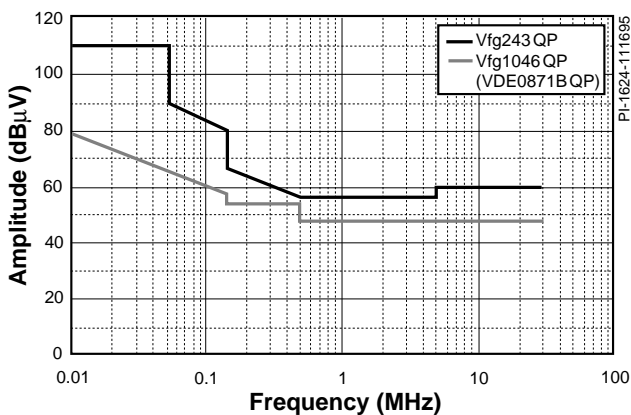


Figure 3. Vfg1046 and Vfg243 Class B Limits (Quasi Peak).

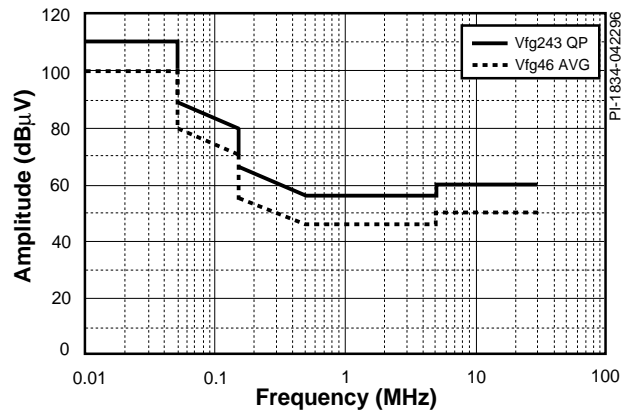


Figure 4. Vfg243 (Quasi Peak) and Vfg46 (Average) Class B Limits.

Safety is a vital issue which determines EMI filter component selection, the transformer reinforced insulation system, and PC board primary to secondary spacing. In fact, safety is an integral part of the power supply/EMI filter design and is difficult to discuss as a separate issue. Throughout this application note, design guidance will also be presented for meeting safety requirements in *TOPSwitch* power supplies.

## EMI Specifications

The applicable EMI specification must be identified for the intended product family and target market. In the United States, the Federal Communications Commission (FCC) regulates EMI specifications. Canadian specifications are similar to FCC specifications. Figure 1 shows the conducted emissions limits governed by FCC rules, Part 15, subpart J. Note that specification limits are given only for quasi-peak detection methods. A recent part 15 amendment allows manufacturers to use the limits contained in C.I.S.P.R. Publication 22 as an alternative when testing devices for compliance<sup>(1)</sup>.

The member countries of the European Community (EC) have established a harmonized program for electromagnetic compatibility. EN55022 for Information Technology Equipment is one of the first harmonized documents. EN55022 together with companion measurement document C.I.S.P.R. Publication 22 set the conducted emission limits shown in Figure 2 for information technology products marketed to the European Community. In fact, EN55022 limits are the same as C.I.S.P.R. Publication 22 limits. Note that class A and class B specification limits are given for both average and quasi-peak detection methods<sup>(2) (3)</sup>.

Figure 3 shows the well-known and most stringent VDE 0871 specification (narrow band limits) for German markets which has traditionally been the design target. German regulation Vfg 1046/1984 requires Information technology or Electronic Data Processing Equipment to meet the VDE 0871 class B narrow band limits from 10 kHz to 30 MHz. Note that specification limits are given only for quasi-peak detection methods. When marketing products only in Germany, there is a choice between meeting the regulation requirements of Vfg 1046/1984 or the new German regulation Vfg 243/1991 (as updated by Vfg 46/1992) which has relaxed limits from 10 kHz to 150 kHz and is harmonized with EN55022 from 150 kHz to 30 MHz. Vfg 243/1991 sets quasi-peak limits and Vfg 46/1992 adds mean or average limits as shown in Figure 4. Figure 3 also shows Vfg 243/1991 class B quasi-peak limits to compare with VDE0871<sup>(4) (5) (6)</sup>. The EMI filter designed to meet VDE 0871 (per Vfg 1046/1984) will generally be higher cost than the EMI

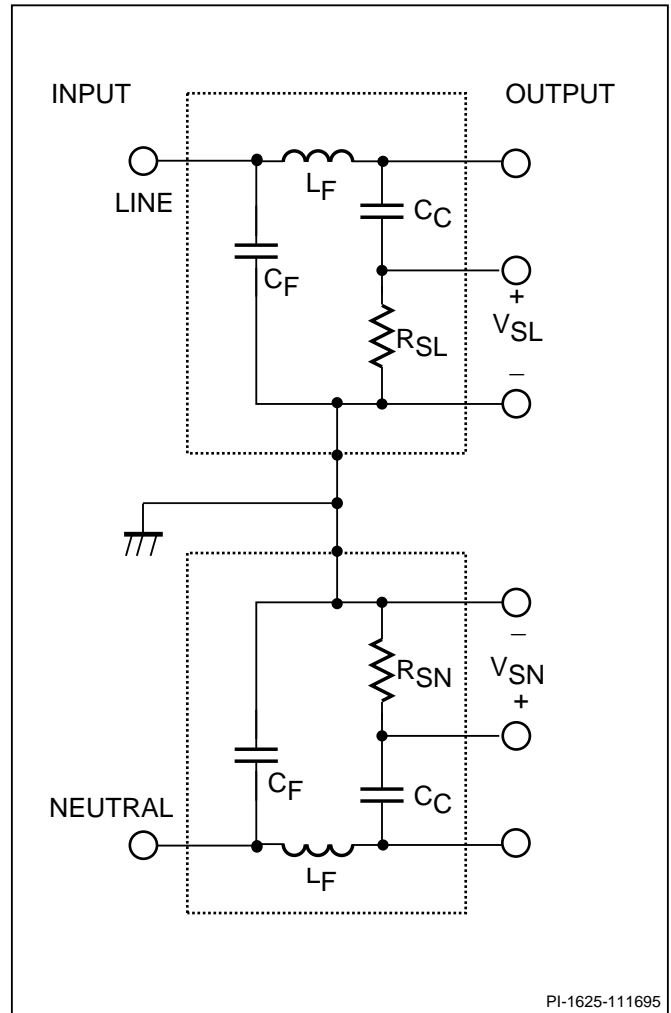


Figure 5. Line Impedance Stabilization Network (LISN).

filter designed to meet Vfg/243 regulation requirements.

## Measuring Conducted Emissions

Details of testing apparatus and methodology are governed by the various EMI regulations, but share the same general concept. Conducted emissions measurements are made with a Line Impedance Stabilization Network (LISN). Figure 5 shows the effective filter, represented by  $L_F$  and  $C_F$ , inside the LISN which passes line frequency currents but forces higher frequency power supply conducted emission currents to flow through coupling capacitor  $C_C$  and sense resistor  $R_S$ . A spectrum analyzer or EMI receiver reads the current emission signal magnitude as sensed voltages  $V_{SL}$  and  $V_{SN}$  across  $R_{SL}$  and  $R_{SN}$  in dB $\mu$ V.

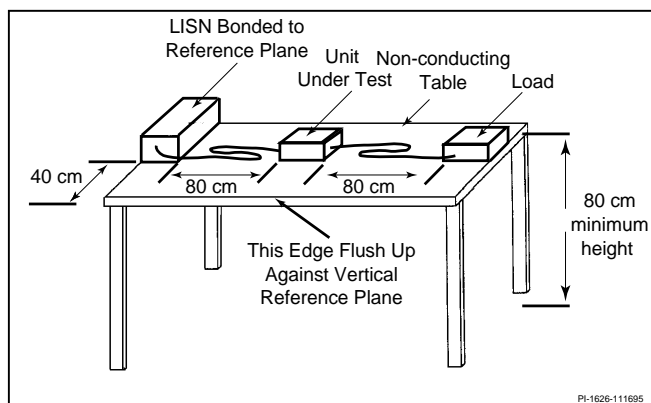


Figure 6. Typical Conducted Emissions Precompliance Test Set Up.

Figure 6 shows a typical conducted emissions pre-compliance test setup on a wooden table at least 80 cm high constructed with non-metallic fasteners<sup>(7)</sup>. The unit under test, LISNs, and load are all placed 40 cm from the edge of the table as shown. Six foot cables are used between the unit under test and both the LISN on the AC input and the load on the DC output. The LISN and load are each located 80 cm from the unit under test with excess cable bundled non-inductively. The edge of the table is placed flush against a vertical reference plane at least two meters square. The LISN is bonded to the reference plane with a low impedance, high frequency grounding strap or braided cable.

In applications where the power supply and load are located in the same physical package, the cable can be omitted between the unit under test and the load.

For design, investigation and precompliance testing, a spectrum analyzer is highly recommended compared to EMI receivers which are more expensive and more difficult to use. For conducted and radiated emissions testing, the spectrum analyzer should have a frequency range of 10 kHz to 1 GHz, wide range of resolution bandwidths (including C.I.S.P.R. specified bandwidths of 200 Hz, 9 kHz, 120 kHz), built in quasi-peak detector, video filter bandwidth adjustment capability down to 3 Hz or below for average measurements, maximum hold for peak measurements, and an accurate and temperature compensated local oscillator capable of centering a 100 kHz signal in the display with insignificant frequency drift. The HP 8591EM and Tektronix 2712 (option 12)<sup>(8)</sup> are two examples of lower cost spectrum analyzers sufficient for conducted emissions precompliance testing.

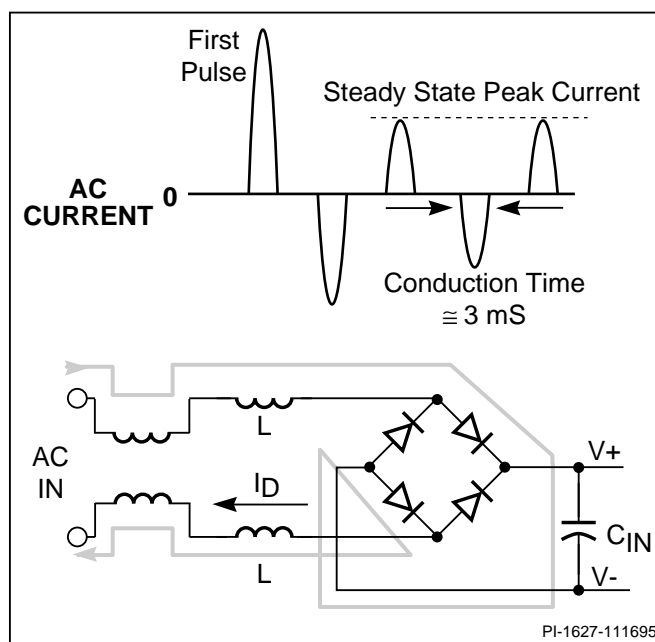


Figure 7. Differential Mode Currents Charging Input Capacitor  $C_{IN}$

## Peak, Quasi-Peak, and Average Detection

Power supplies operating from the 50 or 60 Hz AC mains use a bridge rectifier and large filter capacitor to create a high voltage DC bus from the AC input voltage as shown in Figure 7. The bridge rectifier conducts input current for only a short time near the peak of AC mains voltage. Actual conduction time is typically 3 mS out of effective line frequency periods of 8.3 to 10 mS which defines an effective “line frequency duty cycle” of 30% to 36%. Conducted emission currents can flow in the AC mains leads (and are sensed by the LISN) only during the bridge rectifier conduction time. The conducted emissions signal is actually applied to the spectrum analyzer or receiver detector input only during bridge diode conduction time which defines a “gating pulse” with pulse repetitive frequency (PRF)<sup>(8)(9)</sup> equal to the AC mains frequency (50 or 60 Hz) and “line frequency duty cycle” just defined. The “gating pulse” effect due to bridge rectifier conduction time causes the measured signal magnitude to change depending on whether peak, quasi-peak, or average detection methods are used.

A spectrum analyzer or EMI receiver displays the RMS value of the signal<sup>(9)</sup>. For example, a 100 kHz continuous sinusoidal

voltage when viewed on an oscilloscope may have a peak voltage of 1 Volt and hence an RMS voltage of 0.707 Volts. The spectrum analyzer (with 50  $\Omega$  input) will display a value for this 100 kHz signal of 0.707 volts (or 117 dB $\mu$ V or 10 dBmW) regardless of which detection method is used (peak, quasi-peak, or average) because the signal is continuous, narrow band, and not modulated or gated. If the signal was broadband, modulated, gated at a duty cycle, or in some other way not continuous, the displayed RMS value will change with the detection method. The measured display will then be the magnitude of an equivalent continuous sinusoidal signal with an RMS value equal to the RMS content of the LISN signal measured at the output of the detector stage.

Peak detection is the simplest and fastest method when measuring conducted emissions. Resolution bandwidth is set to 200 Hz for measurements from 10 kHz to 150 kHz and set to 9 kHz for measurements from 150 kHz to 30 MHz. Sweep times are relatively low. When displaying emissions in real time with no averaging, the peaks are not constant but change in magnitude with each measurement sweep due to the bridge conduction gating pulse effect described above. Most spectrum analyzers have a “maximum hold” feature which displays the highest peak occurring over many measurement sweeps. The peak detector measures the magnitude of the largest signal occurring during the bridge conduction gating pulse.

The average detector is simply a low pass filter with corner frequency sufficiently below the gating pulse repetitive frequency or PRF. In typical spectrum analyzers, the video filter bandwidth can be reduced to 30 Hz or below to average the signal but the sweep time must be increased for a calibrated measurement. For test purposes, the full conducted emissions range starting at 10 kHz (or 150 kHz or 450 kHz, depending on the regulation) up to 30 MHz should first be examined with a peak detection measurement. Peak detected emissions with insufficient margin compared to the regulation average limit should be centered on the spectrum analyzer display with the lowest possible frequency span per division setting before reducing video bandwidth and performing the average measurement sweep<sup>(10)</sup>. Figure 8 shows typical conducted emissions from 10 kHz to 500 kHz with both peak detection and average detection. Note that peak detection picked up an envelope of high order harmonics from line frequency rectification in addition to the fundamental and first three harmonics of the 100 kHz switching frequency.

The quasi-peak detector is designed to indicate the subjective annoyance level of interference. As an analogy, a soft noise that happens every second is much more annoying than a loud noise

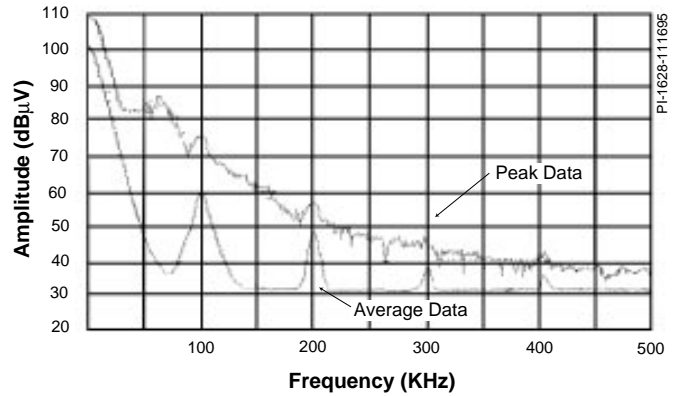


Figure 8. Peak Data vs Average Data.

that happens every hour. A quasi peak-detector (actually a calibrated, intermediate bandwidth video filter) behaves as a leaky peak detector that partially discharges between input signal pulses. The lower the pulse repetitive frequency (PRF), the greater the dB differential between the peak and quasi-peak measured response<sup>(8)(9)</sup>.

Quasi-peak and average detection methods will always give a lower measured value compared to peak detection. If a peak detector measurement meets the average or mean specification limit with sufficient margin, additional measurements using average detection are not necessary. When no average limit is specified, if the peak measurement meets the quasi-peak limit with sufficient margin, additional measurements using quasi-peak detection are not necessary. In general, when testing *TOPSwitch* power supplies to the C.I.S.P.R. Publication 22, EN55022, or Vfg 243/91 (and Vfg 46/92) limits, peak measured data usually meets the quasi-peak limit but, in some areas, may have insufficient margin when compared with the average limit. In this case, further measurement is necessary using average detection.

## Safety Principles

Safety principles must be examined before proceeding further with EMI filter concepts because safety requirements place several constraints on EMI filter design.

Virtually all equipment including computers, printers, televisions, television decoders, video games, battery chargers, etc., must be safety recognized by meeting the safety standard for the intended market and carrying the appropriate safety mark. Safety principles are very similar among the various standards. This application note will focus on the electric shock hazard requirements of one popular standard, IEC950<sup>(11)</sup>.

The European International Electrotechnical Commission Standard IEC950 entitled “Safety of Information Technology Equipment Including Electrical Business Equipment” provides detailed requirements for safe equipment design. Application of IEC950 is intended to prevent injury or damage due to hazards including electric shock, energy hazards, fire hazards, fire, mechanical and heat hazards, radiation hazards, and chemical hazards. IEC950 specifies the following definitions and requirements applicable to *TOPSwitch* power supplies. (This is only a partial list of the key requirements targeted specifically at typical *TOPSwitch* power supply implementations. The appropriate IEC950 section is identified by parentheses.)

**IEC950 Definitions (Applicable to *TOPSwitch* Power Supplies):**

(Introduction): Electric shock is due to current passing through the human body. Currents of approximately 1 mA can cause a reaction in persons of good health and may cause indirect danger due to involuntary reaction. Higher currents can have more damaging effects. Voltages up to about 40 V peak, or 60 VDC are not generally regarded as dangerous under dry conditions, but parts which have to be touched or handled should be at earth ground potential or properly insulated.

(1.2.4.1): Class I Equipment: equipment where protection against electric shock is achieved by:

- a) using basic insulation, and also
- b) providing a means of connecting to the protective earthing conductor in the building wiring those conductive parts that are otherwise capable of assuming hazardous voltages if the basic insulation fails.

(1.2.4.2): Class II Equipment: equipment in which protection against electric shock does not rely on basic insulation only, but in which additional safety precautions, such as double insulation or reinforced insulation, are provided, there being no provision for protective earthing or reliance upon installation conditions.

(1.2.8.1): Primary circuit: An internal circuit which is directly connected to the external supply mains or other equivalent source. In a *TOPSwitch* power supply, this includes the EMI filter, discrete or common mode chokes, bridge rectifier, transformer primary, *TOPSwitch*, and any components connected to *TOPSwitch* such as primary bias windings and optocoupler output transistors.

(1.2.8.2): Secondary circuit: A circuit which has no direct connection to primary power (except through properly selected Y-capacitors) and derives its power from a transformer.

(1.2.8.5): Safety extra-low voltage (SELV) circuit: A secondary circuit which is so designed and protected that under normal and single fault conditions, the voltage between any two accessible parts, or between one accessible part and the equipment protective earthing terminal for class I equipment, does not exceed a safe value.

(1.2.9.2): Basic Insulation: insulation to provide basic protection against electric shock.

(1.2.9.3): Supplementary Insulation: Independent insulation applied in addition to basic insulation in order to ensure protection against electric shock in the event of a failure of the basic insulation.

(1.2.9.4): Double Insulation: Insulation comprising both basic insulation and supplementary insulation.

(1.2.9.5): Reinforced Insulation: A single insulation system which provides a degree of protection against electric shock equivalent to double insulation.

(1.2.9.6): Working voltage: The highest voltage to which the insulation under consideration is, or can be, subjected when the equipment is operating at its rated voltage under conditions of normal use.

(1.2.9.7): Tracking: the progressive formation of conducting paths on the surface of a solid insulating material (such as PC board or transformer bobbin) due to the combined effects of electric stress and electrolytic contamination on this surface.

(1.2.10.1): Creepage distance: the shortest path between two conductive parts, or between a conductive part and the bounding surface of the equipment, measured along the surface of the insulation. In a *TOPSwitch* power supply, the most important creepage distance is between all primary circuits and all secondary circuits (typically 5mm to 6 mm).

(1.2.10.2): Clearance: the shortest distance between two conductive parts, or between a conductive part and the bounding surface of the equipment, measured through air.

(1.2.11.1): Safety Isolating Transformer: the power transformer in which windings supplying SELV circuits are isolated from other windings (such as primary and primary bias windings) such that an insulation breakdown either is unlikely or does not cause a hazardous condition on SELV windings.

**IEC950 Requirements (Applicable to TOPSwitch Power Supplies)**

(1.4.5): In determining the most unfavorable supply voltage for a test, the following variables shall be taken into account:

- multiple rated voltages
- extremes of rated voltage ranges
- tolerance on rated voltage as specified by the manufacturer.  
If a tolerance is not specified, it shall be taken as +6% and - 10%.

(1.6.5): Equipment intended to operate directly from the mains supply shall be designed for a minimum supply tolerance of +6%, -10%.

(2.1.10): Equipment shall be so designed that at an external point of disconnection of the mains supply, there is no risk of electric shock from stored charge on capacitors connected to the mains circuit. Equipment shall be considered to comply if any capacitor having a rated capacitance exceeding 0.1 uF and connected to the external mains circuit, has a means of discharge resulting in a time constant not exceeding 1 second for pluggable equipment type A (non-industrial plugs and socket-outlets). This requirement specifically applies to any EMI filter capacitor connected directly across the AC mains which could cause a shock hazard on the exposed prongs of an unplugged power cord.

(5.2.2): Earth Leakage Current: Maximum earth leakage current must not exceed the limits shown in the following table under the most unfavorable (highest) input voltage. For class II equipment when output is not connected to earth ground, the test shall be made on accessible conductive parts, and to metal foil with an area not exceeding 10 cm x 20 cm on accessible non-conductive parts.

Class	Type of Equipment	Maximum Leakage Current
II	All	0.25 mA
I	Hand-held	0.75 mA
I	Movable (other than hand-held)	3.50 mA

Table 1. Maximum Leakage Current.

(5.3.2): Electric Strength: The insulation shall be subjected for 1 minute either to a voltage of substantially sine-wave form having a frequency of 50 Hz or 60 Hz or to a DC voltage equal to the peak of the prescribed AC test voltage. Test voltage shall be as specified in the following table for the appropriate grade of insulation and the working voltage U across the insulation:

(5.4.1): Abnormal Operating and Fault Conditions: Equipment

Grade of insulation	U ≤ 130 VAC	130 ≤ U ≤ 250VAC
Basic, Supplementary Reinforced (Primary to Secondary)	1000 VAC 2000 VAC	1500 VAC 3000 VAC

Table 2. Insulation Electric Strength.

shall be so designed that the risk of fire or electric shock due to mechanical or electrical overload or failure, or due to abnormal operation or careless use, is limited as far as practicable.

(5.4.6): For components and circuits (other than motors, transformers, PC board creepage and clearance distances, or secondary circuit electromechanical components) compliance with the abnormal operating and fault condition requirement (5.4.1) is checked by simulating the following conditions:

- faults in any components in primary circuits (which includes EMI filter components, bridge rectifier, energy storage capacitor, TOPSwitch, and all TOPSwitch connected components);
- faults in any components where failure could adversely affect supplementary or reinforced insulation (specifically failure of Y2-capacitors connected between primary circuits and secondary circuits);
- additionally, for equipment that does not comply with the requirement of Sub-clauses 4.4.2 (Minimizing the risk of ignition) and 4.4.3 (Flammability of materials and components), faults in all components;
- faults arising from connection of the most unfavorable load impedance to terminals and connectors that deliver power or signal outputs from the equipment, other than mains power outlets (for example: connecting a class II equipment output terminal to earth ground will increase measured leakage current).

The equipment, circuit diagrams and component specifications shall be examined to determine those fault conditions that might reasonably be expected to occur.

(In general, components designed for use between primary and secondary circuits, rated for the full electric strength voltage, and carrying the appropriate safety agency approvals are not subject to the single component fault test because a short circuit fault is extremely unlikely. Two component examples are safety rated optocouplers and Y1-capacitors which can be applied directly between primary and secondary circuits operating from AC mains with rated voltages up to 250 VAC.)



## Typical AC Mains Input Voltage Configurations

*TOPSwitch* power supplies are typically connected to the AC mains in either 2-wire or 3-wire configurations. For the purposes of EMI design presented in this application note, 2-wire and 3-wire configurations are now defined.

### 2-Wire AC Input

The *TOPSwitch* power supply 2-wire AC mains connection may consist of one line wire and one neutral wire where the AC mains neutral is eventually connected back to earth ground at a local electrical panel. The 2-wire connection may also consist of two separately phased line wires where neither is connected directly to earth ground. The power supply SELV output may or may not be connected directly to earth ground.

In this application note, the neutral wire will be treated as an ungrounded AC mains or separately phased line conductor requiring the same safety considerations as any AC mains line conductor. In addition, the power supply SELV output return will be assumed to connect directly to earth ground which represents the worst case and most unfavorable connection for safety considerations.

### 3-Wire AC Input

In 3-wire connections, the third wire earth ground wire will be available for connection to EMI filter components, shields, chassis, and enclosures. The neutral wire will be treated as an ungrounded AC mains or separately phased line conductor requiring the same safety considerations as any AC mains line conductor. In addition, the power supply SELV output return will be assumed to connect directly to earth ground which represents the worst case and most unfavorable connection for safety considerations.

## EMI Filter Components

EMI filters are actually simple combinations of inductors or chokes and capacitors. Series resistors, which lead to undesirable power dissipation, are not normally used for reducing conducted emissions.

Single-section EMI filters (one stage of common mode and differential mode attenuation) take the least space and have the lowest cost but require careful attention to details such as circuit parasitics, component parasitics, and layout to meet the specifications with adequate margin. Multiple-section filters can also be used because one stage can be designed to overcome the deficiencies of the other. The two section design will reduce

current emissions and increase margin below the specification limit but may not address size or cost goals of the end product. Understanding the basics of EMI filter design and application allows the designer to implement small, low cost, single section EMI filters.

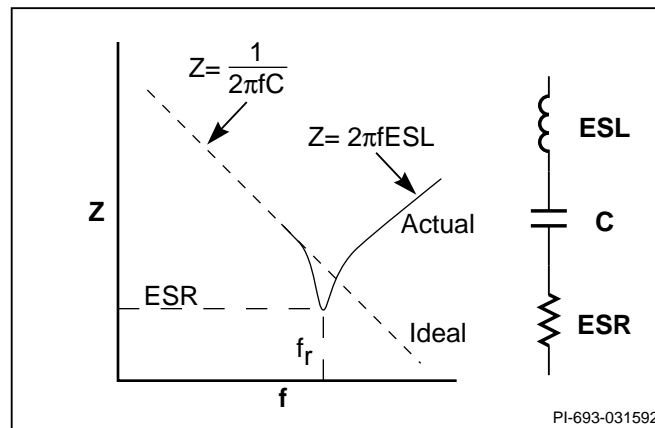


Figure 9. Comparison of Ideal and Real Capacitor Impedance.

### Capacitors

Proper capacitor selection for EMI filters requires attention to three key parameters: impedance characteristics, voltage ratings, and safety specifications.

Figure 9 shows impedance characteristics for ideal and non-ideal capacitor behavior. An ideal capacitor has an impedance characteristic that decreases linearly with frequency. A real capacitor has parasitic inductance and resistance elements which cause the impedance to behave quite differently from an ideal capacitor.

Equivalent series inductance (ESL) creates a capacitor self resonant frequency  $f_r$  as shown on the plot. The impedance of the capacitor at this self-resonant frequency is determined by equivalent series resistance (ESR). Beyond the self-resonant frequency ( $f_r$ ), the capacitor actually acts like an inductor. Capacitors with plastic film, combination plastic film/paper, or ceramic dielectrics usually have the highest self-resonant frequencies and are commonly used in EMI filters.

### Aluminum Electrolytic Energy Storage Capacitor

Switching power supplies always have a bridge rectifier and high voltage bulk energy storage aluminum electrolytic capacitor to convert AC mains input voltage to high DC bus voltage (typically 100 to 400 Volts DC) shown as  $C_{IN}$  in Figure 7.

The impedance of this capacitor, which must always be minimized, provides the first level of differential mode conducted emissions filtering.

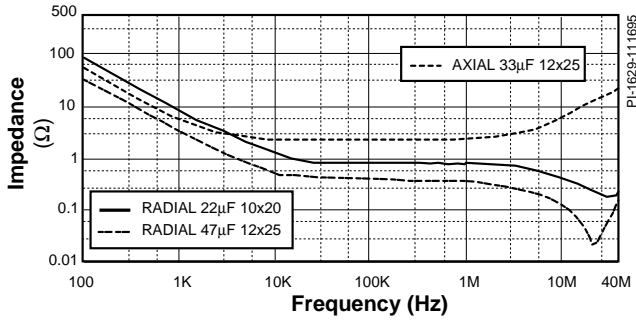


Figure 10. 200V Aluminum Electrolytic Capacitor Impedance.

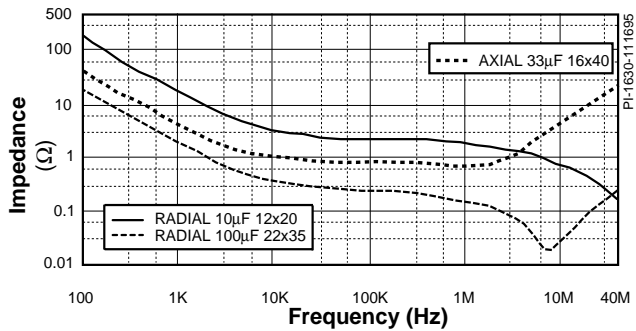


Figure 11. 400V Aluminum Electrolytic Capacitor Impedance.

Figures 10 and 11 show impedance of various 200V and 400V aluminum electrolytic capacitors with radial leads (both leads exiting one side of the capacitor can) compared with impedance of a similar capacitor with axial leads (one lead exiting each side of the capacitor can). Approximate dimensions are also shown (diameter by length in mm). Radial capacitors have an impedance characteristic that stays low up to 10 MHz while the axial capacitors become inductive at frequencies as low as 1 MHz. Radial capacitors should always be used and installed on end to minimize lead length and ESL. Axial leaded capacitors should never be used because the longer total lead length (equal to at least one can diameter) increases ESL which increases impedance. Note that above 1 MHz, the large axial capacitors actually have much higher impedance (and will generate higher conducted emission currents) than the smaller radial capacitors.

**EMI Filter Capacitors**

Capacitors used in EMI filters are identified by various companies as radio interference suppressors, suppression capacitors, or safety recognized capacitors. These capacitors must meet the European requirement EN 132400 for safety which defines two groups, X and Y<sup>(12) (13)</sup>.

X-capacitors are used only in positions where capacitor failure does not expose anybody to an electric shock hazard. X-capacitors are usually connected across the AC mains as part of

the differential mode portion of the EMI filter. X-capacitors are divided into three subclasses:

Subclass	Peak Pulse Voltage In Service	IEC-664 Installation Category	Application	Peak Impulse Voltage $V_p$ applied before Endurance Test
X1	> 2.5 kV ≤ 4.0 kV	III	High Pulse Application	$C \leq 1.0 \mu F$ $U_p = 4 \text{ kV}$
X2	≤ 2.5 kV	II	General Purpose	$C \leq 1.0 \mu F$ $U_p = 2.5 \text{ kV}$
X3	≤ 1.2 kV	-	General Purpose	None

Table 3. X-Capacitor Subclass.

X2-capacitors are most commonly used in TOPSwitch power supply EMI filters for differential mode suppression. X1-capacitors can also be used but cost is higher. X3-capacitors are not normally used.

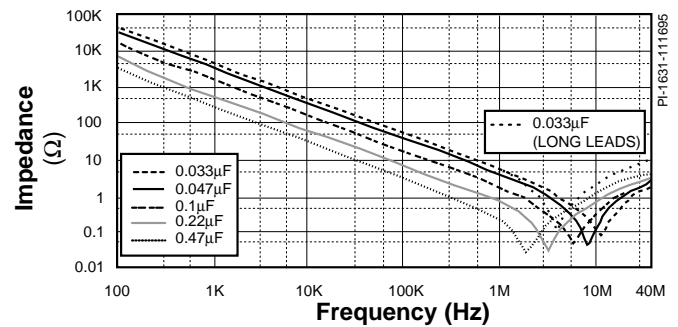


Figure 12. X2-Capacitor Impedance.

X2-capacitors are available from a variety of vendors including Murata, Roederstein, Panasonic, Rifa, and Siemens. Figure 12 shows impedance plots for various sizes of X2-capacitors with short leads and one plot for a small X2-capacitor with long leads. Short leads should always be used to minimize impedance and reduce high frequency conducted emission currents.

Y-capacitors are used where capacitor failure could expose somebody to an electric shock hazard. Y-capacitors are usually connected from the AC mains or bridge rectifier output to SELV secondaries, chassis, shields, or earth ground. The maximum Y-capacitor value is restricted because each application has an allowable maximum leakage current (which can range from 0.25 mA to 3.5 mA, depending on the AC mains connection). There are four EN 132400 specified subclasses of Y-capacitors:





Subclass	Type of Insulation Bridged	Rated Voltage (VAC)	Test Voltages for Quality Approval, Periodic and Lot-by-Lot Testing	Peak Impulse Voltage $V_p$ applied before endurance
Y1	Double Insulation or Reinforced Insulation	$\leq 250$ V	4000 VAC	8.0 kV
Y2	Basic Insulation or Supplementary Insulation	$\geq 150$ V $\leq 250$ V	1500 VAC	5.0kV
Y3	Basic Insulation or Supplementary Insulation	$\geq 150$ V $\leq 250$ V	1500 VAC	none
Y4	Basic Insulation or Supplementary Insulation	$< 150$ V	900 VAC	2.5 kV

Table 4. Y-Capacitor Subclass.

In two-wire 230 VAC or universal input applications, a single Y1-safety capacitor can be directly connected between the AC mains or bridge rectifier output to the SELV secondary. The single Y1-capacitor will also meet the electric strength voltage requirement (for 230 VAC mains connected power supplies, typically 3,000 VAC for one minute). Y1-capacitors with a value of 1000 pF are available from Murata<sup>(14)</sup> (ACT4K-KD series, DE1110 E 102M ACT4K-KD), Roederstein<sup>(15)</sup> (WKP series, WKP102MCPE.OK) and Rifa<sup>(12)</sup> (PME 294 series, PME 294RB4100M). In general, Y1-capacitors are not used in 3-wire applications.

Y2-capacitors do not meet reinforced insulation requirements. In a single component failure safety investigation, one Y2-capacitor may be replaced with a wire jumper to see if an electric shock or fire hazard condition will exist. In most 2-wire applications, a series combination of two 2200 pF Y2-capacitors are commonly used between primary and SELV outputs so that a short circuit failure of one Y2-capacitor creates no safety hazard. A series connection of two Y2-capacitors is also necessary to meet the electric strength requirement (for 230 VAC mains connected power supplies, typically 3,000 VAC for one minute). In 3-wire applications, the Y2-capacitor may be directly connected between AC mains or bridge rectifier output and earth ground because the earth ground wire will safely shunt the fault current created by a shorted Y2-capacitor. Y2-capacitors rated at 250 VAC are available from a variety of vendors including Murata, Roederstein, Panasonic, Rifa, and Siemens. Figure 13 shows impedance plots for various sizes of Y2-capacitors with short leads and one plot for a large Y2-capacitor with long leads. Y-capacitors perform most of the high frequency filtering from 10 MHz to 200 MHz. Note that capacitor resonant frequency is usually 40 MHz or higher unless artificially reduced with long leads or long PC traces. Long leads and long PC traces can also cause emission currents, though low enough to meet conducted emissions specifications, to radiate sufficient energy from the power cord to exceed

radiated emissions specifications. Using short leads and short PC traces for all Y-capacitor connections is critical to meet both conducted and radiated emissions specifications.

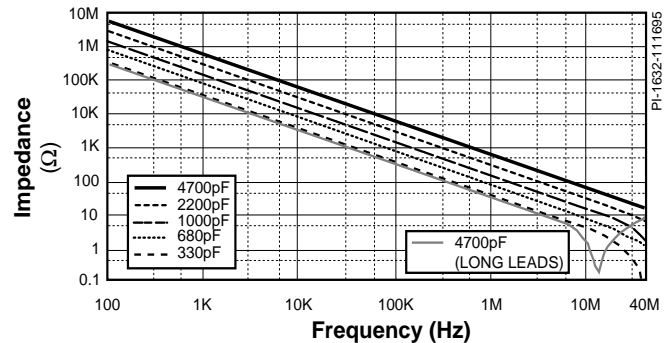


Figure 13. Y2-Capacitor Impedance.

In 115 VAC applications, a series combination of two Y2-safety or two Y4-safety capacitors can be directly connected between the AC mains or bridge rectifier to the SELV secondary. Y3-safety capacitors are not normally used.

Safety specifications such as UL1950, UL544, and IEC950 limit the amount of fault current that can flow when a safety ground connection has been opened or one component has failed (Y1-capacitors, because of their construction, are excluded from the failed component test). For example, UL1950 specifies that information technology equipment with Class I or three wire input (line, neutral, and earth ground), 240 VAC, 60 Hz input must have a leakage current no higher than 3.5 mA if earth ground is opened or one component has failed short which restricts Y-capacitor maximum value below 0.039  $\mu$ F (or 39 nF). For class II or two wire input (line, neutral, with no earth ground), leakage current must be less than 250  $\mu$ A with one failed component which restricts Y-capacitor size to under 0.0028  $\mu$ F (2.8 nF or 2800 pF) for 240 VAC, 60 Hz input. Capacitor and input voltage tolerance must also be taken into

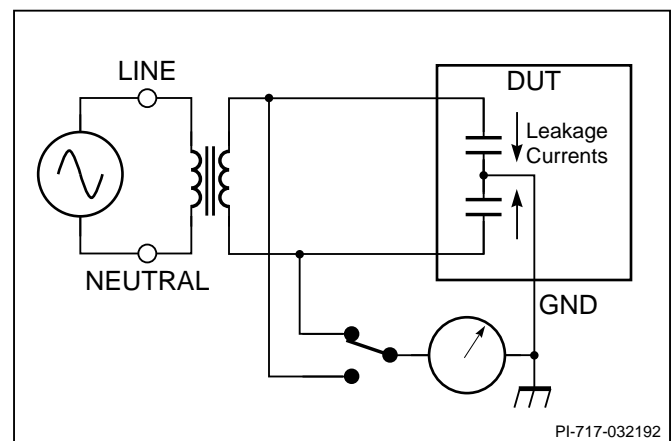


Figure 14. Typical Safety Measurement Setup.

account. Figure 14 shows a typical test setup for measuring leakage current.

**Inductors or chokes**

Proper inductor selection for EMI filters requires attention to three key parameters: effective impedance characteristic, current rating, and surge current capability.

Figure 15 shows impedance characteristics for ideal and non-ideal inductor behavior. Ideal inductors have an impedance characteristic that increases linearly with frequency. Real inductors have parasitic series resistance  $R_s$  and parallel interwinding capacitance ( $C_w$ ).  $C_w$  creates a resonant frequency as shown on the plot. Beyond the resonant frequency ( $f_r$ ), the inductor actually behaves like a capacitor.

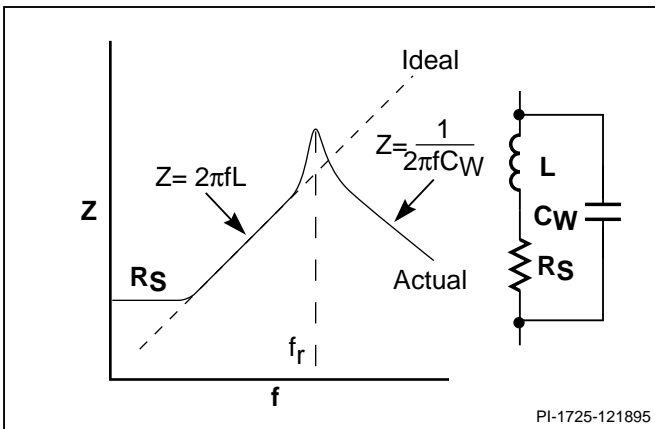


Figure 15. Comparison of Real and Ideal Inductor Impedance.

Power supplies have bridge rectifier input filters which draw line frequency currents with high peak values but relatively narrow widths as previously shown in Figure 7. A discrete filter choke usually has a minimal effect on the peak current but must pass the peak current without significant saturation (which reduces effective inductance). The discrete choke must also be rated to safely pass the higher peak value of the first surge of current occurring when AC power is initially applied with input capacitor  $C_{IN}$  completely discharged.

**Differential mode chokes**

Differential mode chokes are simply discrete inductors designed for EMI filters that pass line frequency or DC currents while blocking or filtering high frequency conducted emission currents. Differential mode chokes are usually wound on low cost solenoidal cores of either iron powder or ferrite material as shown in Figure 16. Toroids tend to be significantly higher in cost but can also be used. Chokes with single layer windings have the lowest capacitance and highest resonant frequency.

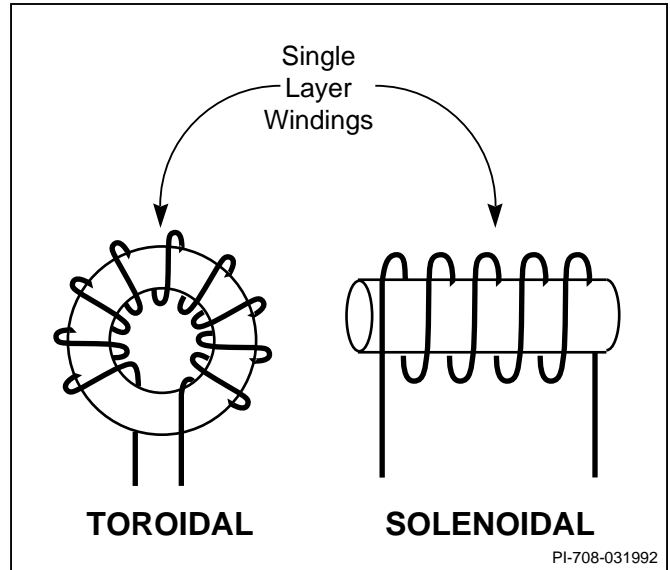


Figure 16. Differential Mode Chokes.

Effective inductance varies with peak differential mode choke current flow. Refer again to Figure 7 where the bridge rectifier and filter creates a high voltage DC bus from the AC line. AC input current flows only during a small conduction time as shown. Peak AC input current during normal operation is relatively high. Differential mode chokes are designed or selected to limit saturation at peak AC input current. Figure 17 shows how inductance for a powdered iron toroidal core varies with number of turns and peak current. To achieve the desired inductance under high peak AC input current, higher numbers of turns and/or larger choke cores are normally required. Typical impedance characteristics for two different differential mode chokes are shown in Figure 18. Note that the larger choke resonates at a lower frequency and becomes capacitive. The smaller choke has a higher impedance above 3 MHz due to the

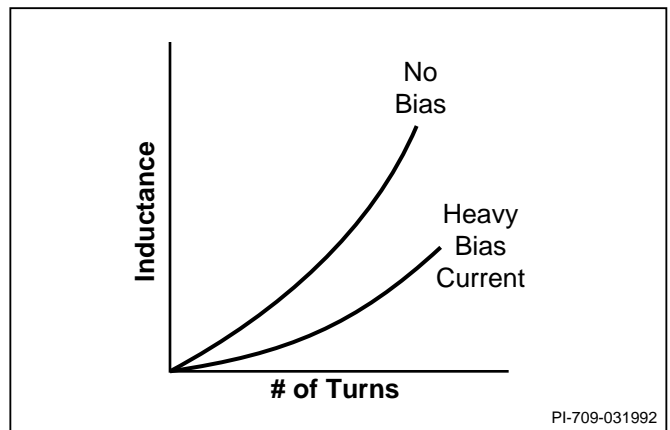


Figure 17. Inductance Under Current Bias.



higher self-resonant frequency. Installing the larger choke to attenuate the fundamental may have the effect of letting through current components above 3 MHz.

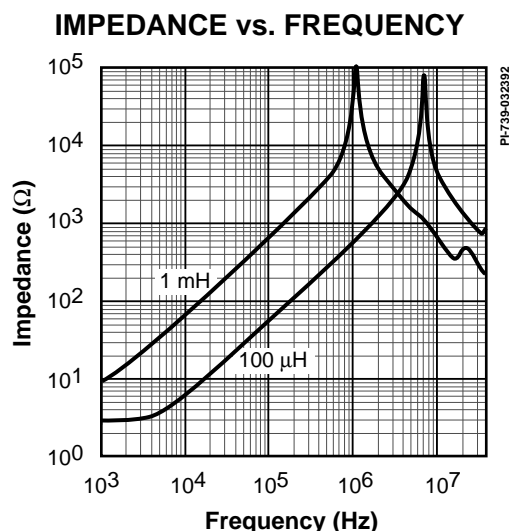


Figure 18. Typical Differential Mode Choke Impedance.

Differential mode chokes are usually used in EMI filters for both differential mode and common mode filtering only for the lowest output power levels (under 5 Watts). At higher power levels, a properly selected common mode choke will also have differential mode inductance for essentially no additional cost.

### Common Mode chokes

Common mode chokes are specialized inductors designed specifically for common mode EMI filters. The common mode choke consists of two identical windings wound such that the magnetic fields caused by differential mode currents cancel. Figure 19 shows a toroidal implementation which is good for illustration purposes but (as will be seen shortly) is not the best choice for low-cost and practical EMI filter implementations. Figure 19 shows three current components  $I_D$ ,  $I_{C1}$ , and  $I_{C2}$ .  $I_D$  is a differential mode current (shown also in Figure 7) which circulates by starting at the AC mains source, flows through one common mode choke winding towards the power supply, flows through one bridge rectifier diode, charges the high voltage energy storage capacitor  $C_{IN}$ , flows back through another bridge rectifier diode, and then flows back towards the source through the other common mode choke winding. The magnetic fields within the core due to the circulating differential current  $I_D$  cancel perfectly because of dot polarity. Note that the “start” of both windings enters the core on the same side and the “finish” of both windings leaves the core on the other side. Common mode chokes behave like short circuits for circulating differential mode currents such as  $I_D$  which flow in through one common mode choke winding and flow out through the other common

mode choke winding.  $I_{C1}$  and  $I_{C2}$  are “common mode” currents which may or may not be related in magnitude and phase. The common mode choke behaves like a large inductor to common mode currents.

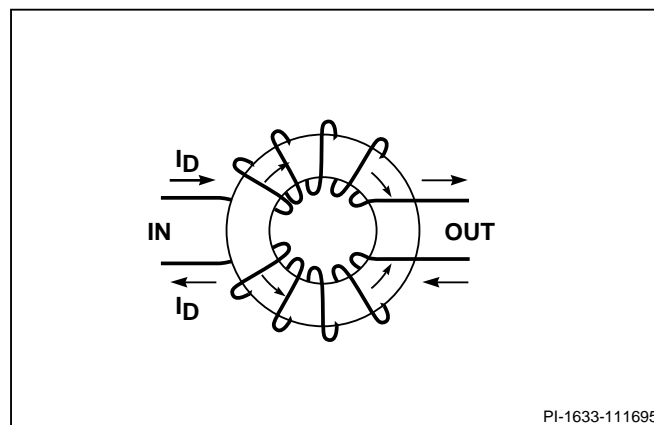


Figure 19. Ideal Common Mode Choke.

Two low-cost bobbin style common mode chokes simplify EMI filter design. Figure 20 shows a typical “U-core” style common mode choke in which the windings are wound on a conventional bobbin. Two U-core halves are inserted into the bobbin and secured with clamps. U-core common mode chokes are widely available from several companies such as Tokin<sup>(16)</sup>, Tamura<sup>(17)</sup>, Panasonic/Matsushita<sup>(18)</sup>, TDK<sup>(19)</sup>, and Murata<sup>(20)</sup>.

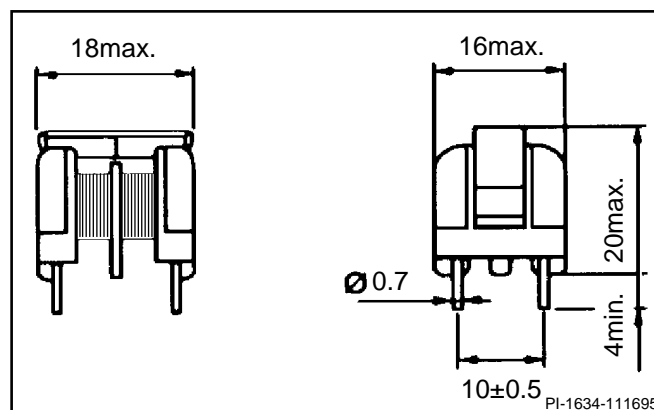


Figure 20. U Core Common Mode Choke (All dimension in mm).

Figure 21 shows a newer common mode choke design with a “spool” style two-piece bobbin. The two-piece bobbin is snapped together around a one-piece ungapped core. A sprocket on the bobbin engages a gear on a winding machine to spool the wire onto the bobbin. Spool style common mode chokes are available from Panasonic/Matsushita<sup>(18)</sup> and Tokin<sup>(16)</sup>.

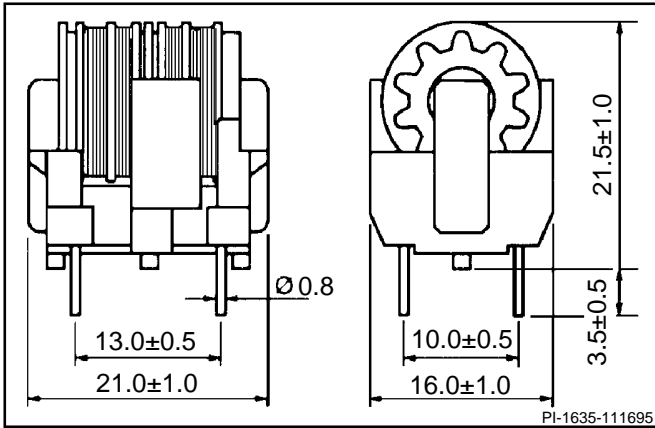


Figure 21. Spool Wound Common Mode Choke (All dimension in mm).

One very important advantage to the bobbin style common mode choke is an “inherent differential mode choke” due to parasitic leakage inductance which usually eliminates any need for additional discrete differential mode chokes. Figure 22 shows the effective common mode choke schematic consisting of a common mode inductance in series with an effective differential mode leakage inductance. Unlike most other magnetic components, leakage inductance in a common mode choke is a desirable parasitic effect which provides balanced differential mode filtering for no additional component cost. The common mode choke is modeled by a common mode inductance in series with a differential mode inductance.

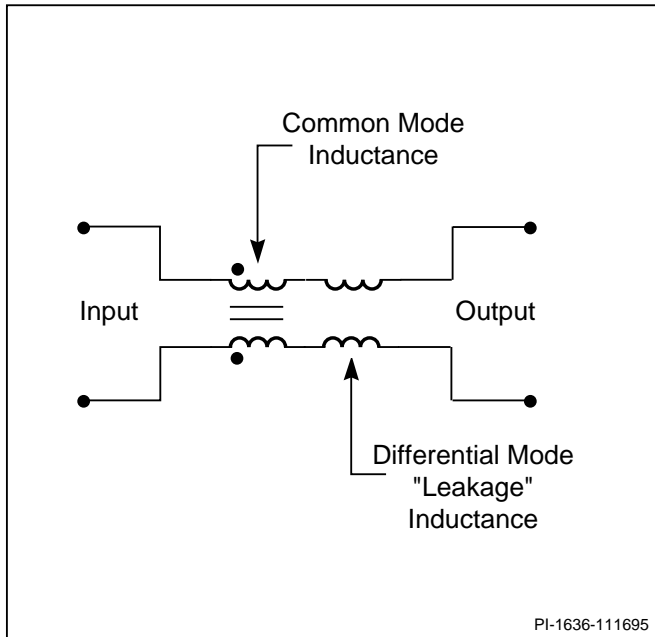


Figure 22. Effective Common Mode Choke Schematic.

Common mode inductance of each winding is the measured inductance of one winding with the other winding open circuited. Differential mode inductance of each winding is equal to half the measured inductance of one winding with the other winding short circuited.

Common mode impedance is shown for the U-core style in Figure 23 and the spool style in Figure 24. Also shown is common mode impedance for a typical toroidal implementation. Note that the toroidal common mode impedance is generally lower than both the U-core and spool style common mode chokes.

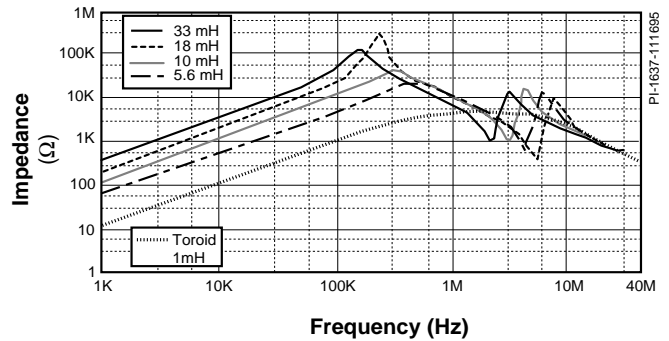


Figure 23. U Core Common Mode Choke (Common Mode Impedance).

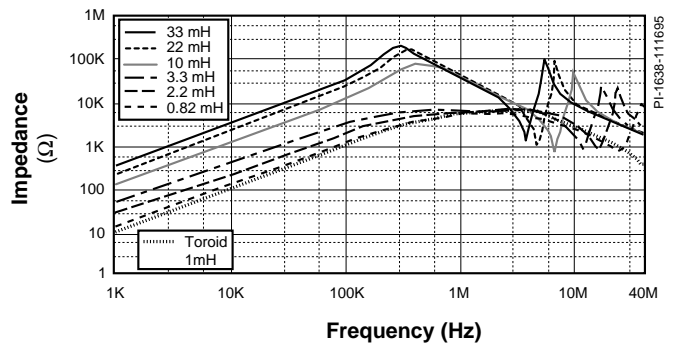


Figure 24. Spool Wound Common Mode Choke (Common Mode Impedance).

Differential mode impedance is shown for the U-core style in Figure 25 and the spool style in Figure 26. Also shown is differential mode impedance for a typical toroidal implementation. Note that the toroidal differential mode impedance is quite a bit lower than both the U-core and spool style common mode choke. With toroidal common mode chokes, additional differential mode chokes are usually required. For these reasons, toroidal common mode chokes are not recommended except for the high frequency, supplemental toroidal common mode choke described below.



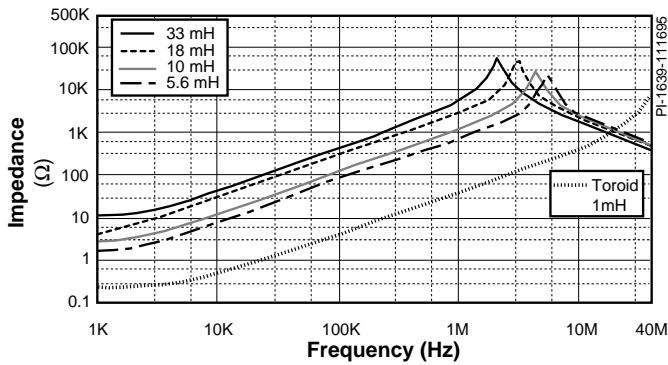


Figure 25. U-Core Wound Common Mode Choke (Differential Mode Impedance).

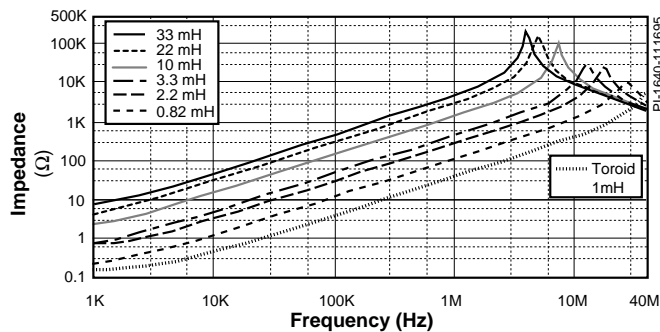


Figure 26. Spool Wound Common Mode Choke (Differential Mode Impedance).

Bobbin style common mode chokes can have either one or two sections in each winding. One section per winding is lowest cost but two sections per winding splits the winding capacitance in half to increase resonant frequency and effective bandwidth. The U-core common mode choke shown in Figure 20 has one section per winding while the spool style common mode choke shown in Figure 21 has two sections per winding. Figure 23 shows that the single section U-core style common mode impedance is lower and resonant frequency is lower with sharper peaking compared with the two section spool style common mode impedance shown in Figure 24. Two sections per winding reduce capacitance to improve common mode impedance at high frequency.

The common mode choke must also survive the surge current occurring when voltage is first applied to the power supply as described earlier, as well as operate at the steady-state RMS input current.

For reducing high frequency common mode conducted emissions in the 10 MHz to 200 MHz range, a simple common mode choke using a small ferrite toroid<sup>(21)</sup> and insulated wire can be wound as shown in Figure 27 and used in addition to one of the bobbin style common mode chokes. Both wires have thick, safety insulated wires with different colors. The wires are held

together in parallel and wound as a pair for typically 3 to 5 turns. The toroidal core should be ferrite and “lossy” at high frequency such as Fair-Rite 75 material. Fair-Rite toroid part number 5975001101 (with 0.5 inch OD x 0.32 inch ID x 0.25 inch thickness) is suitable for most applications<sup>(21)</sup>. This high frequency common mode choke is usually located between power entry and the rest of the power supply EMI filter. This common mode choke technique can also be used on power supply output wires.

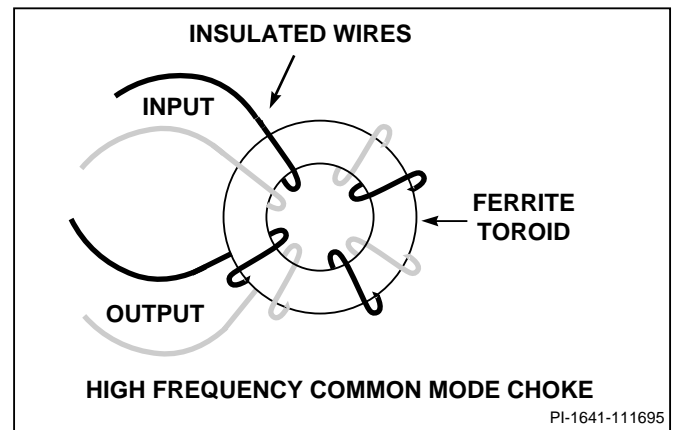


Figure 27. High Frequency Common Mode Choke.

## Flyback Power Supply EMI Signature

Flyback power supplies have a distinctive EMI signature caused by superposition of several waveforms shown in Figure 28. The transformer primary current  $I_{PRI}$ , *TOPSwitch* Drain voltage  $V_{Drain}$ , diode voltage  $V_{Diode}$ , and transformer secondary current  $I_{SEC}$  waveforms each generate emission currents which may exceed the desired EMI specification limits without proper EMI design technique.

### Primary Current Waveform

Primary current  $I_{PRI}$  begins to flow when *TOPSwitch* turns on. Transformer primary current ramps to a peak value determined by input voltage, primary inductance, switching frequency, and duty cycle. This trapezoidal (or triangular) current waveform is characterized in the frequency domain by a spectrum with a fundamental at the switching frequency and harmonics determined by the relative squareness of the waveform and causes primarily differential mode emission currents to circulate between the AC mains and the power supply input. This current waveform can also create common mode emissions due to radiated magnetic fields if the current path defined by the PC board layout encircles a large physical area.

### *TOPSwitch* Drain-Source Voltage Waveform

The Drain-Source voltage waveform  $V_{Drain}$  is characterized by high  $dv/dt$  transitions. Parasitic circuit elements (leakage inductance, *TOPSwitch* output capacitance, and transformer

capacitance) cause additional voltage peaking and ringing at frequencies typically between 3 MHz and 12 MHz. The *TOPSwitch* Drain, transformer primary, and Drain clamping components connected to the Drain node will drive displacement currents to earth ground through transformer capacitance or stray capacitance. This displacement current returns “backwards” through the line and neutral conductors back to the *TOPSwitch* Drain driving node as a common mode emission current. The displacement currents generated by the drain voltage waveform cause spectral energy in the form of a common mode conducted emission currents to be concentrated at the switching frequency and 3 MHz to 12 MHz resonant frequency ( $f_1$ ) of the indicated ringing voltage waveform.

Common mode emission currents will be lower with *TOPSwitch* when compared with discrete MOSFET implementations because *TOPSwitch* has a controlled turn on gate driver to reduce  $dv/dt$ . Common mode emissions currents are also lower because the *TOPSwitch* TO-220 tab is connected to the relatively quiet source pin while a discrete MOSFET has the noisy drain “transmitting” node connected directly to the tab (and heat sink) “broadcasting antenna”.

### Diode Voltage Waveform

The diode voltage waveform  $V_{DIODE}$  is also characterized by fast voltage changes and fast rise and fall times. Parasitic circuit elements (transformer leakage inductance and diode capacitance)

cause additional voltage peaking and ringing at frequencies typically between 20 MHz and 30 MHz. The diode voltage waveform will drive displacement currents to earth ground through transformer capacitance or stray capacitance. The displacement currents generated by the diode voltage waveform cause spectral energy in the form of common mode emission currents to be concentrated at the switching frequency and 20 MHz to 30 MHz resonant frequency ( $f_2$ ) of the indicated ringing voltage waveform.

### Secondary Current Waveform

Secondary current  $I_{SEC}$  begins to flow as soon as *TOPSwitch* turns off. Current starts at a peak value and decreases linearly at a rate determined by secondary inductance and output voltage. This trapezoidal (or triangular) current waveform is characterized in the frequency domain by a spectrum with a fundamental at the switching frequency and harmonics determined by the relative squareness of the waveform. Additional ringing superimposed on the waveform is related to the drain source voltage  $V_{Drain}$  waveform previously discussed. This composite current waveform can cause significant magnetic fields to radiate if the current path defined by the PC board layout encircles a large physical area. Spectral energy in the form of a common mode emission current would be concentrated at the switching frequency and 3 MHz to 12 MHz resonant frequency ( $f_1$ ) of the indicated ringing current waveform.

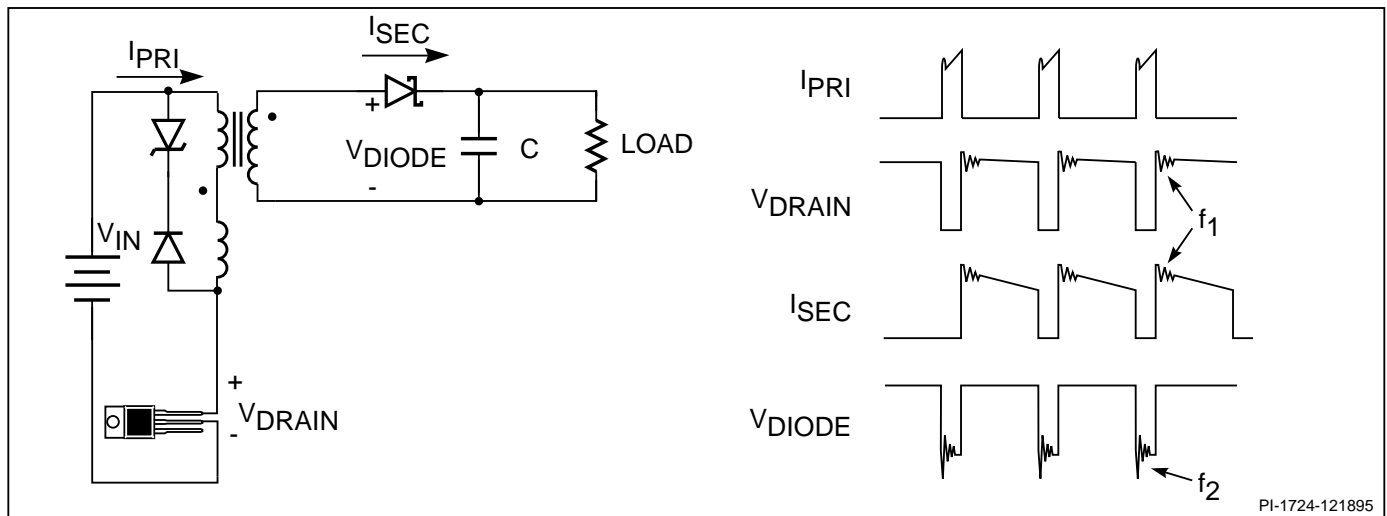


Figure 28. Examples of Typical Flyback Power Supply Waveforms Causing EMI.

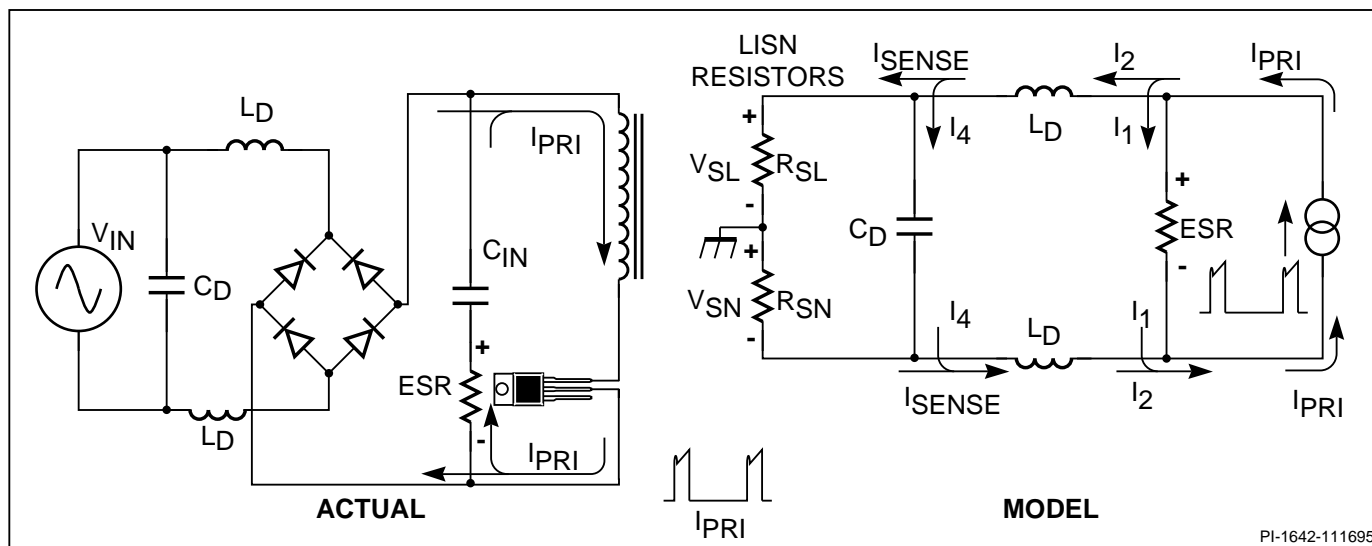


Figure 29. Circuit Origin for Differential Mode Emissions.

## Suppression Techniques

Controlling EMI requires attention to the following areas.

- Differential mode filtering
- Common mode filtering
- Power cord damping
- Transformer construction

### Differential mode Filter Analysis

Differential mode conducted emissions are caused by currents circulating between the power supply and AC mains input which means that a differential current which flows into the power supply through the Line input wire will flow out of the power supply through the Neutral input wire.

Most differential mode conducted emissions are caused by the fundamental and harmonics of the triangular or trapezoidal *TOPSwitch* Drain current waveform. During EMI testing, differential mode currents generate test voltages equal in magnitude and opposite in phase across Line LISN sense resistor  $R_{SL}$  and Neutral LISN sense resistor  $R_{SN}$ .

Differential mode analysis starts by replacing the actual circuitry with an equivalent model as shown in Figure 29. The primary current is modeled by current source  $I_{PRI}$ . The effective impedance of energy storage capacitor C1 over the frequency range of 100 kHz to 1 MHz is modeled by the Equivalent Series Resistance or ESR. The bridge rectifier is assumed to be

conducting current and is replaced with a short circuit. The AC source impedance is modeled by the effective series combination of the 50  $\Omega$  LISN sense resistors  $R_{SL}$  and  $R_{SN}$ . Differential mode filtering is performed by the LC filter consisting of differential mode capacitor  $C_D$  and two identical differential mode chokes  $L_D$ . This model is valid up to roughly 1 MHz.

The primary current switching frequency fundamental and harmonic components  $I_{PRI}(n)$  must be estimated, measured, or derived by simulation. Note that measured harmonic components are given in RMS but calculated or simulated components are given in peak values and must be converted to RMS. A typical harmonics envelope is shown in Figure 30 as a function of frequency.

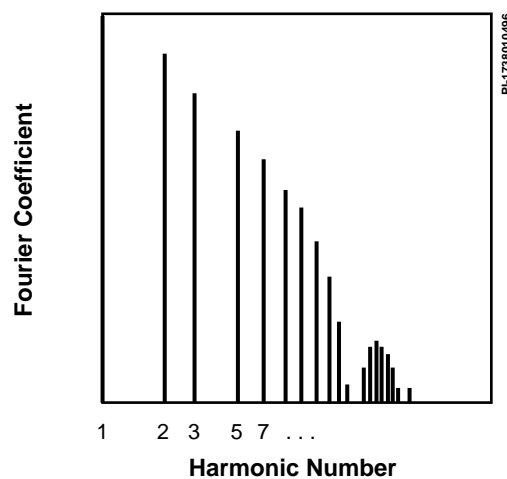


Figure 30. Envelope of Typical Primary Current Fourier Spectrum.

At the fundamental and harmonics of switching frequency  $f_s$ , equivalent series resistance (ESR) of bulk input capacitor  $C_{IN}$  is much lower impedance compared with the  $L_D$  differential mode chokes. Primary current  $I_{PRI}$  flows almost completely through bulk energy storage capacitor  $C_{IN}$  which creates an effective trapezoidal (or triangular) differential mode voltage source proportional to ESR. Differential mode chokes and the differential mode capacitor form a simple low pass filter to attenuate the effective voltage source to a level below the desired specification. Figure 31 shows the final simplified model where the RMS source voltage for each  $n^{th}$  current harmonic (given in peak value) is given by:

$$V_{PRI}(n) = ESR \times I_{PRI}(n) \times \frac{1}{\sqrt{2}}$$

(RMS)                      (Peak)

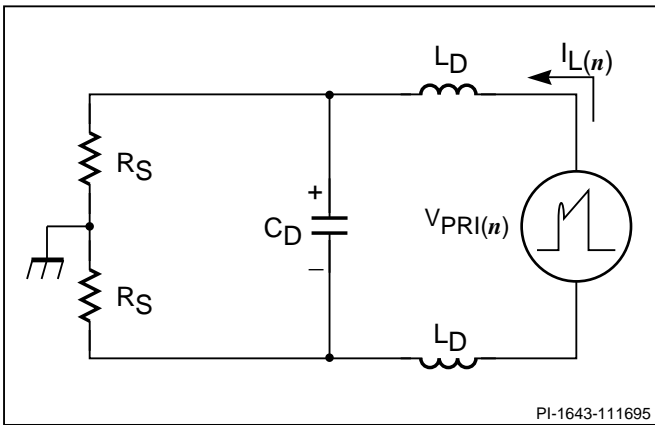


Figure 31. Simplified Differential Mode Model.

Attenuation is determined by the differential between the magnitude of the effective voltage source in  $\text{dB}\mu\text{V}$  and the desired conducted emissions specification. The voltage transfer function  $H(s)$  is given in terms of  $L_D$ ,  $C_D$ , and  $R_S$ .

At high levels of attenuation normally required at the switching frequency, the denominator of  $H(s)$  is dominated by the frequency dependent terms and can be simplified as shown. Simple algebra reveals a very useful frequency domain formula consisting of the product of three separate terms. The first term converts the effective ESR voltage source  $V_{PRI}(s)$  back into differential inductor current  $I_D(s)$ , the second term splits the current between differential mode capacitor  $C_D$  and LISN sense resistors, and the third term senses the LISN current component to create a voltage to be measured with a detector or receiver to compare with limits in  $\text{dB}\mu\text{V}$ . This is a general result with equivalent ESR voltage source  $V_{PRI}(n)$  of each  $n^{th}$  harmonic shown (temporarily) in the frequency domain as  $V_{PRI}(s)$  which is a function of the complex frequency variable  $s$ .

$$H(s) = \frac{V_{SN}(s)}{V_{PRI}(s)} = \frac{1}{2} \times \frac{1}{(2 \times L_D \times C_D \times s)^2 + \left(\frac{L_D}{R_S} \times s\right) + 1}$$

$$\approx \frac{1}{2} \times \frac{1}{(2 \times L_D \times C_D \times s)^2 + \left(\frac{L_D}{R_S} \times s\right)}$$

$$V_{SN}(s) = V_{PRI}(s) \times \frac{1}{2 \times L_D \times s} \times \frac{1}{(2 \times R_S) + \frac{1}{C_D \times s}} \times R_S$$

For EMI filter design, only the magnitude of the most important frequency components are examined which allows simple magnitude expressions in terms of the harmonic integer  $n$  to be used (rather than the complex variable  $s$ ). Filter design begins by identifying a target sense voltage  $V_{SN\text{dB}\mu\text{V}}(n)$  below the specification limits at the appropriate  $n^{th}$  harmonic frequency. For FCC testing, the specification begins at 450 kHz with the fifth harmonic ( $n = 5$ ) while excluding *TOPSwitch* 100 kHz fundamental ( $n = 1$ ) and second through fourth harmonic frequencies ( $n = 2, 3, 4$ ). For European test limits, the 100 kHz fundamental ( $n = 1$ ) and the second harmonic at 200 kHz ( $n = 2$ ) should be examined because the limit changes significantly at 150 kHz. As an example and referring to European EN55022 average limit for class B (Figure 2), the average limit value is 74  $\text{dB}\mu\text{V}$  at 100 kHz ( $n=1$ ) and 53.5  $\text{dB}\mu\text{V}$  at 200 kHz ( $n=2$ ) while the quasi-peak limit values are 10 dB higher. In most low frequency conducted emission measurements, the measured quasi-peak value is slightly less (1 dB to 3 dB) than the peak value. The average value, however, can be 12 dB below the peak value which means that if the filter is designed to meet the average limit, the quasi-peak limit will also be met and with greater margin. In this example and for 12 dB margin overall, the peak value should be designed to touch the average limit and average detection will provide the remaining 12 dB attenuation. The target sense voltages are therefore equal to the average limit or 74  $\text{dB}\mu\text{V}$  at 100 kHz ( $V_{SN\text{dB}\mu\text{V}}(1)$ ) and 53.5  $\text{dB}\mu\text{V}$  at 200 kHz ( $V_{SN\text{dB}\mu\text{V}}(2)$ ).  $V_{SN\text{dB}\mu\text{V}}(n)$  is converted from  $\text{dB}\mu\text{V}$  to an absolute value sense voltage  $V_{SN}(n)$ .

$$V_{SN}(n) = 1.e^{-6} \times 10 \frac{V_{SN\text{dB}\mu\text{V}}(n)}{20}$$

$V_{SN}(1)$  is 5.01  $\text{mV}_{\text{RMS}}$  and  $V_{SN}(2)$  is 473  $\mu\text{V}_{\text{RMS}}$ . Sense voltage  $V_{SN}(n)$  is then converted into an RMS current magnitude  $I_L(n)$  flowing through each differential mode inductor  $L_D$ .





$$I_L(n) = V_{SN}(n) \times \sqrt{\frac{1}{R_s^2} + (4 \times \pi \times n \times f_s \times C_D)^2}$$

RMS differential current  $I_L(1)$  is 638 $\mu$ A and  $I_L(2)$  is 119 $\mu$ A. The target differential inductance  $L_D$  can now be calculated.

$$L_D = \frac{V_{PRI}(n)}{I_L(n) \times 4 \times \pi \times n \times f_s}$$

The ST202A power supply operating from 115 VAC and delivering 15 Watts is found to operate in the discontinuous mode with a triangular drain current waveform. Peak Drain current  $I_p$  is 0.8 A and duty cycle is 0.3.  $C_6$  (0.1  $\mu$ F) is differential capacitor  $C_D$ . ESR of input capacitor  $C_1$  is 0.375  $\Omega$ . From simulation, calculation, or measurement with the power supply connected to the LISN but without an EMI filter, the equivalent source voltage fundamental  $V_{PRI}(1)$  is 59.3 mVRMS and second harmonic  $V_{PRI}(2)$  is 43.0 mVRMS. Differential inductance  $L_D$  is found to be 74 $\mu$ H in each leg for attenuation of the fundamental but the second harmonic requires a higher inductance value of 144 $\mu$ H in each leg to achieve the desired attenuation because the EN55022 specification is more stringent at 200 kHz. The higher inductance value is used in the design. Note also that different combinations of L and C are possible but the LC product will remain the same. Note also that, in common mode chokes, total measured differential inductance is twice the value calculated for each leg (288 $\mu$ H in this example).

Peak load current normally limits the size of discrete chokes to between 100 $\mu$ H and 1 mH (especially in mains applications with peak-charging capacitive input filters). Practical discrete chokes are cost effective only at the lower output power levels (5 Watts and below). Single discrete chokes attenuate the differential mode but have little effect on common mode emission currents. These limitations for discrete, differential mode chokes can be overcome by selecting a common mode choke with parasitic leakage or differential inductance equal to or greater than the differential choke inductance value calculated above. (Note: with a common mode choke, measure inductance of one winding with the other winding shorted for total leakage or differential inductance. The effective differential inductance in each leg is half the measured value.)

Filter effectiveness decreases as parasitic elements of the filter components themselves become significant. The effective circuit model above 1 MHz is shown in Figure 32. Note the additional ESL terms in both energy storage capacitor  $C_{IN}$  and differential capacitor  $C_D$ . Note also the shunt winding capacitance  $C_W$  across each differential mode filter choke  $L_D$ . As the frequency increases, the parasitic components begin to dominate, reducing filter effectiveness. Fortunately, the

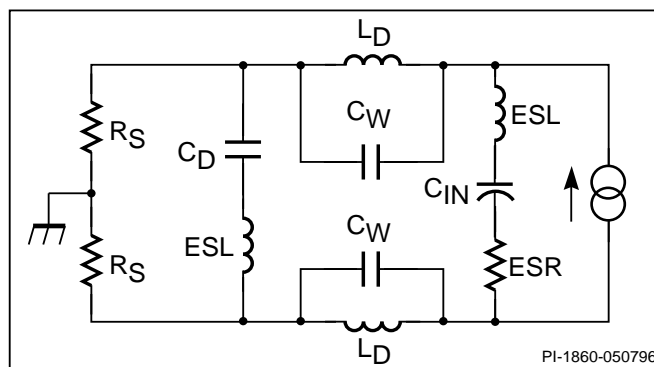


Figure 32. High Frequency Model of the Differential Mode Filter.

harmonics of the trapezoidal (or triangular) *TOPSwitch* Drain current waveform are also decreasing above 1 MHz, which tends to offset the degradation in filter performance. Above 1 MHz, current emissions which exceed the desired specification are usually common mode emissions caused by either ringing waveforms identified earlier or resonances caused by parasitic components themselves.

Physical component layout becomes increasingly critical above 1 MHz. Improper layout can lead to increased capacitor ESL. It is also possible for noise voltages or currents to couple around the EMI filter directly into the mains.

### Common Mode Filter Analysis

Common mode conducted emissions are caused by common mode currents that do not circulate between the AC mains and power supply input. Balanced common mode currents flow simultaneously in power supply line and neutral input wires such that common mode line current is equal in magnitude and in phase with common mode neutral current. Unbalanced common mode currents flow in either power supply line or neutral input wires separately. Common mode conducted emissions are caused by *TOPSwitch* Drain Voltage  $V_{DRAIN}$  and output Diode Voltage  $V_{DIODE}$  as shown in Figure 33.

*TOPSwitch* Drain voltage  $V_{DRAIN}$  drives displacement current through various stray parasitic capacitance terms.  $C_{S1}$  is stray *TOPSwitch* Drain capacitance to earth ground.  $C_{OSS}$  is *TOPSwitch* output capacitance.  $C_{BD1}$  through  $C_{BD4}$  are the effective capacitance terms across each bridge diode.  $C_{AC}$  is the capacitive coupling across the AC mains input (which is very low when testing with LISNs). Note that secondary is shown connected directly to earth ground. Transformer capacitance is distributed but can be modeled with the following six discrete capacitance terms:

$C_{W1}$ : Winding capacitance from “noisy” or switching side of the transformer primary to “noisy” side of the secondary.

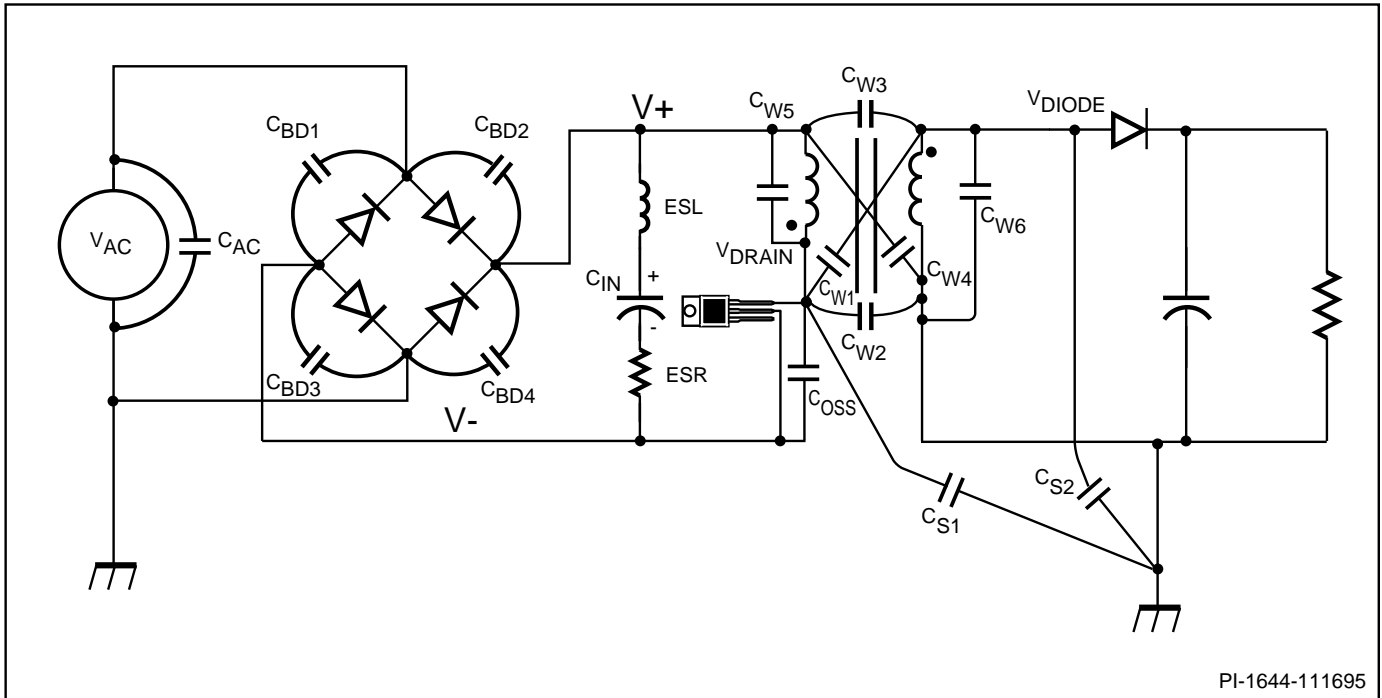


Figure 33. Circuit Origin for Common Mode Emissions.

$C_{W_2}$ : Winding capacitance from “noisy” or switching side of the transformer primary to “quiet” side of the secondary.

$C_{W_3}$ : Winding capacitance from “quiet” side of the transformer primary to the “noisy” or switching side of the secondary.

$C_{W_4}$ : Winding capacitance from “quiet” side of the transformer primary to the “quiet” side of the secondary. (This is actually a “good” stray capacitance term which is usually augmented with an additional, Y-capacitor to return displacement currents back to the driving source).

$C_{W_5}$ : Winding capacitance across the primary.

$C_{W_6}$ : Winding capacitance across the secondary ( $C_{W_5}$  and  $C_{W_6}$  combine to cause a transformer resonant frequency of 400 kHz to 2 MHz above which each winding impedance is capacitive rather than inductive).

The *TOPSwitch* Drain node directly drives displacement current into each of the following stray capacitance terms:  $C_{S1}$ ,  $C_{W1}$ ,  $C_{W2}$ ,  $C_{OSS}$ , and  $C_{W5}$ . Each displacement current ( $I_{CS1}$ ,  $I_{CW1}$ ,  $I_{CW2}$ ,  $I_{COSS}$ , and  $I_{CW5}$ ) must eventually return to the driving node (*TOPSwitch* Drain pin). Each current splits many times but some fraction of each displacement current may flow through the power supply AC input conductors and be measured as

common mode emission currents as follows:

$I_{CS1}$ : Splits into earth ground component and secondary component. Secondary current component returns to the *TOPSwitch* Drain through transformer capacitance. Earth ground component returns up from ground into the neutral wire (and is sensed by the LISN), AC couples into the Line wire (and is sensed by the LISN), flows back through bridge diodes (either superimposing on line frequency current during bridge diode conduction or through effective bridge diode capacitance when diodes are not conducting), to V+ and V- bus wires, and returns to the *TOPSwitch* Drain pin through  $C_{W5}$  and  $C_{OSS}$ .

$I_{CW1}$ : Splits between  $C_{W3}$  returning to primary,  $C_{S2}$  to earth ground, and  $C_{W6}$ . Splits again between  $C_{W2}$  and earth ground. Earth ground components returns through neutral and line wires (as explained above) and is sensed by the LISN.

$I_{CW2}$ : Splits between  $C_{W6}$  and earth ground. Earth ground component returns through neutral and line wires (as explained above) and is sensed by the LISN.

$I_{COSS}$ : Splits between  $C_{IN}$  (note that  $C_{IN}$  Equivalent Series Inductance or ESL will choke off high frequencies) and V-. V- component flows out the bridge rectifier, down through line and neutral wires (and is sensed by the LISN) to earth

ground, up to secondary, and couples back to *TOPSwitch* through transformer winding capacitance.

$I_{CWS}$ : Splits between  $C_{IN}$  (with ESL) and  $V+$ .  $V+$  component flows out the bridge rectifier, down through line and neutral wires (and is sensed by the LISN) to earth ground, up to secondary, and couples back to *TOPSwitch* through transformer winding capacitance.

Superposition of all these different displacement currents will lead to some cancellation but there will always be “leftover” high frequency current components measured as common mode conduction emissions. The asymmetries in various parasitic capacitance terms also explain how common mode emission currents can become “unbalanced” creating net current flowing only in the line or neutral wire. A similar analysis can be performed using the output rectifier anode as the driving point voltage source. For obvious reasons, common mode emission currents are best measured because analysis is quite difficult.

Common mode filters require relatively high values of inductance because safety standards restrict common mode Y-capacitor size to limit leakage current as previously discussed. Common mode chokes between 10 mH and 33 mH are used in most applications because inductance normally required is unaffected by the circulating differential mode current. Discrete chokes can also be used in some low power applications if the peak current is taken into account and a discrete choke is placed in each leg for balanced high frequency impedance.

Physical component layout becomes increasingly critical above 1 MHz. Improper layout can lead to increased capacitor ESL. It is also possible for noise voltages in close proximity to the EMI filter to couple around the filter directly into the mains. Common mode capacitors must have extremely short traces connecting directly to the transformer pins and to each other as well.

## Power Cord Damping

Applications with 3-wire power cords require special attention. A six foot power cord can be modeled as a transmission line with distributed inductance and capacitance, characteristic impedance of approximately  $100 \Omega$ , and little damping which leads to a sharp, well defined resonance, typically between 15 and 25 MHz. This resonance can amplify existing common mode emission currents to levels in excess of the desired limit. A small, lossy ferrite bead or toroid placed over the earth ground lead wire reduces the resonant peak by adding series damping.

The ferrite bead or toroid<sup>(21)</sup> should have an effective impedance of  $100 \Omega$  in the 15 to 25 MHz range. The bead is placed over the safety ground wire between the enclosure power entry connector and the internal safety ground attachment point of the enclosure. The toroid is installed in similar fashion but can accommodate up to 5 or 6 turns of the insulated safety ground wire.

## Transformer Construction

Flyback transformers use gapped ferrite cores which may have fringing fields as shown in Figure 34. Gaps should be confined to the center leg of either one or both core halves so that the fringing field can be effectively shielded by the windings. End gaps “leak” magnetic flux due to the fringing field which can produce common mode emissions.

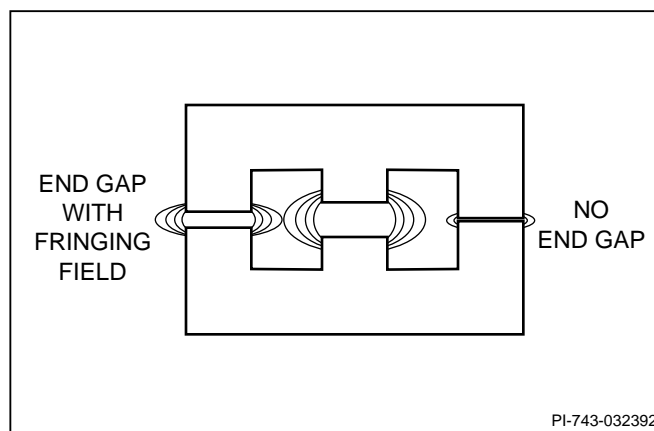


Figure 34. End Gap Magnetic Flux Leakage.

Proper transformer construction techniques are necessary for reducing common mode emissions. Figure 35 shows a typical insulated wire wound transformer cross section. The transformer primary connects between the relatively quiet high voltage DC bus and the noisy *TOPSwitch* Drain pin (which has the high voltage switching waveform). When the primary is wound with two layers, the primary half with the dot mark is connected to *TOPSwitch* which is then buried or shielded under the primary half connected to the high voltage DC bus as shown. One layer of 2 mil tape separates the two primary halves to reduce capacitance and high frequency ringing. Another layer of tape separates the primary winding from the insulated wire wound secondary. The combination of tape and insulation thickness reduces capacitance between primary and secondary which reduces common mode emission currents. One more layer of tape separates the secondary from the primary referenced bias winding.

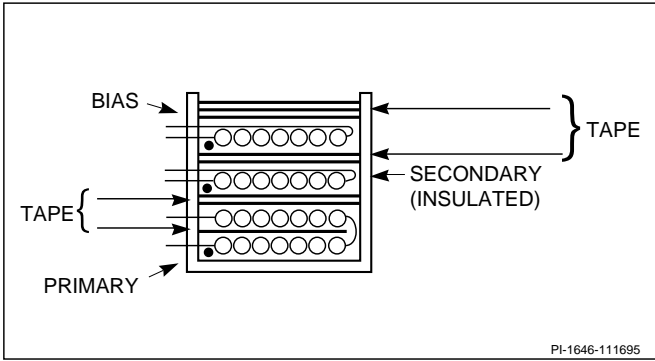


Figure 35. Insulated Wire Wound Transformer Cross Section.

Figure 36 shows a typical margin wound transformer cross section. Each winding is placed between symmetric margins as shown while the safety insulation extends beyond the margins up to the walls of the bobbin flange. The split primary construction shown reduces leakage inductance in higher power applications. The primary half connected to *TOPSwitch* is the first layer followed by one wrap of 2 mil polyester film tape for basic insulation. The bias winding is wound next in a single layer. The bias winding is usually just a few turns but wound using up to three parallel wires to cover more of the bobbin width and effectively shield the noisy *TOPSwitch* Drain connected primary half. Three wraps of 2 mil polyester film tape (3M 1298 or equivalent UL recognized tape) provide the necessary reinforced insulation for safety as well as reducing capacitance between primary and secondary to minimize common mode emission currents. Secondary is now wound between margins followed again by three wraps of 2 mil polyester film tape for reinforced insulation. The second half of the primary is wound followed by three wraps of 2 mil polyester film tape as final insulation.

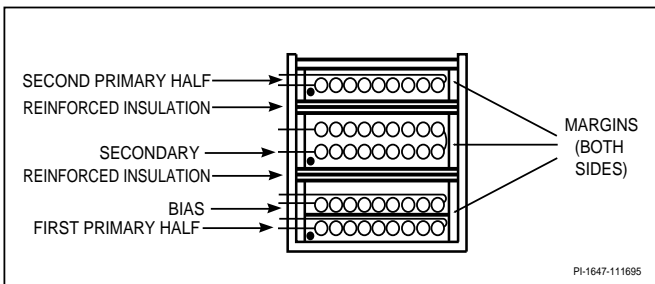


Figure 36. Margin Wound Transformer Cross Section.

In some applications, a copper foil “flux band” over the outside of the completed transformer as shown in Figure 37 may reduce some common mode emissions. The copper foil wraps one complete turn over the exposed (but insulated) windings and each core end gap. Foil width is cut to fit between bobbin flanges while maintaining required creepage distance for reinforced insulation. For best effect, the foil is electrically

connected to each core half (manganese zinc ferrite normally used in 100 kHz flyback transformers is conductive) and the foil ends are electrically connected to create a shorted turn. The flux band can usually be left floating without making additional bobbin changes to meet reinforced insulation requirements for safety. The flux band (and core) may also be connected back to primary or to secondary but bobbin construction must be examined to ensure creepage distance is sufficient to meet reinforced insulation requirements for safety.

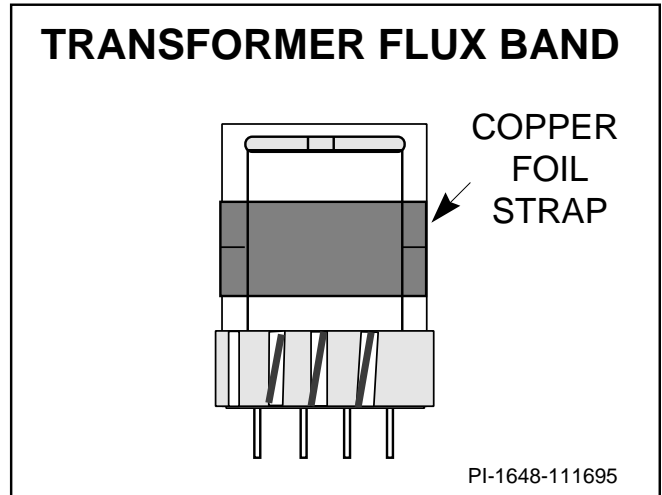


Figure 37. Transformer Flux Band.

In most *TOPSwitch* power supplies, a transformer shield is not necessary because *TOPSwitch* has controlled turn on which limits high voltage  $dv/dt$  and reduces common mode emission currents. For those few applications where further reduction in common mode emission currents is desired, Figure 38 shows proper shield placement within the transformer. The shield intercepts interwinding capacitive displacement currents and returns them to the primary circuitry. Figure 38 shows the shield connected to  $V+$  but the shield can also be connected to  $V-$  if more convenient for construction or layout reasons. Note that safety insulation or creepage distance is required between the primary connected shield and SELV secondary outputs. The foil shield width is selected to fit between primary safety margins. Length is precut for one full turn with slight overlap at the ends. The termination lead wire is soldered to the copper foil shield in the center (equidistant from each end). Tape insulation is usually applied to the copper foil shield before placing on the transformer. The ends must be insulated such that the foil shield does not form a shorted turn inside the transformer (compared with the external “belly band” described earlier which has a shorted turn but is physically located outside the transformer).

Refer to AN-18 for more information on transformer construction.



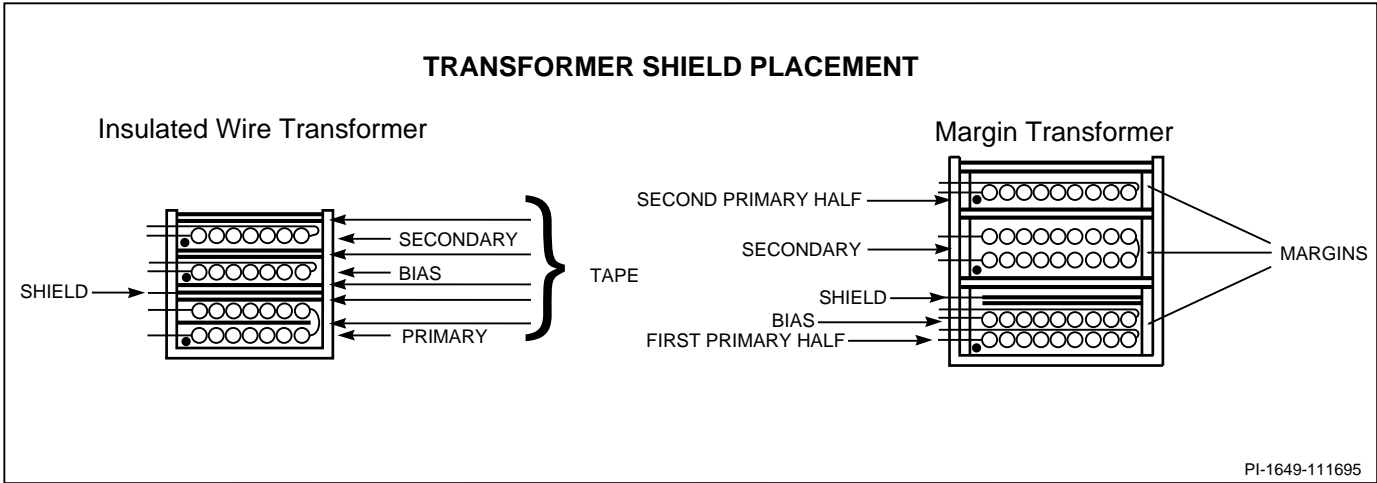


Figure 38. Transformer Shield Placement.

## General Purpose TOPSwitch EMI Filters

### 2-Wire AC Input

A typical TOPSwitch power supply and EMI filter for 2 wire AC input applications is shown in Figure 39. X-capacitor C6 and the differential mode inductance of common mode choke L2

attenuate differential mode emission currents. C7 (Y1-safety capacitor) and the common mode inductance of common mode choke L2 attenuate common mode emission currents. Note that C7 can be replaced by two series connected Y2-safety capacitors, each with twice the value shown.

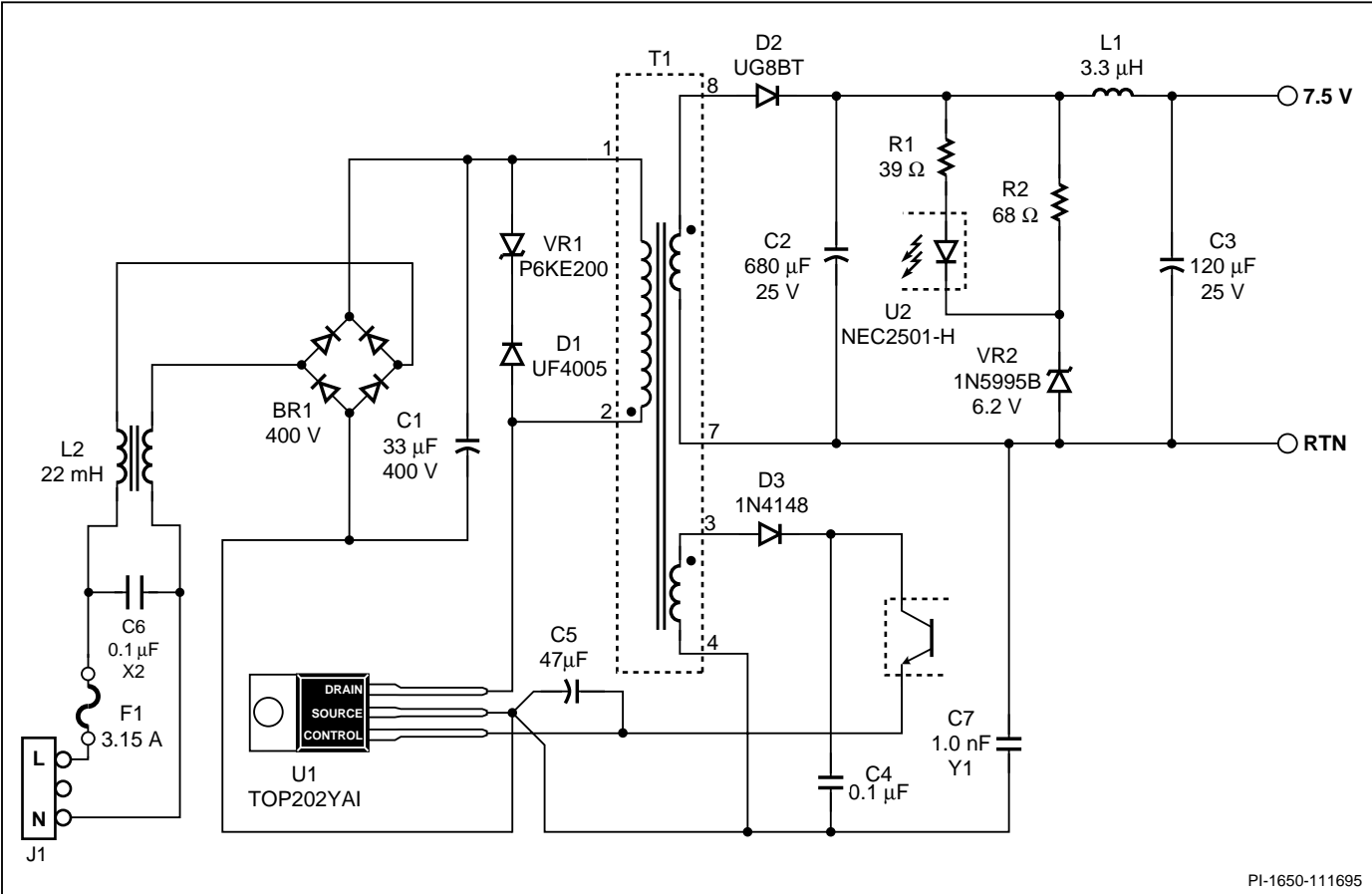


Figure 39. Typical 2-Wire TOPSwitch Power Supply and EMI Filter.



**3-Wire AC Input**

A typical EMI filter for 3 wire AC input applications is shown in Figure 40. X-capacitor C6 and the differential mode inductance of common mode choke L2 attenuate differential

mode emission currents. C7 (Y1-safety capacitor), the common mode inductance of common mode choke L2, and small, lossy ferrite toroid L3 attenuate common mode emission currents<sup>(2)</sup>. L3 also damps power cord resonances as previously described.

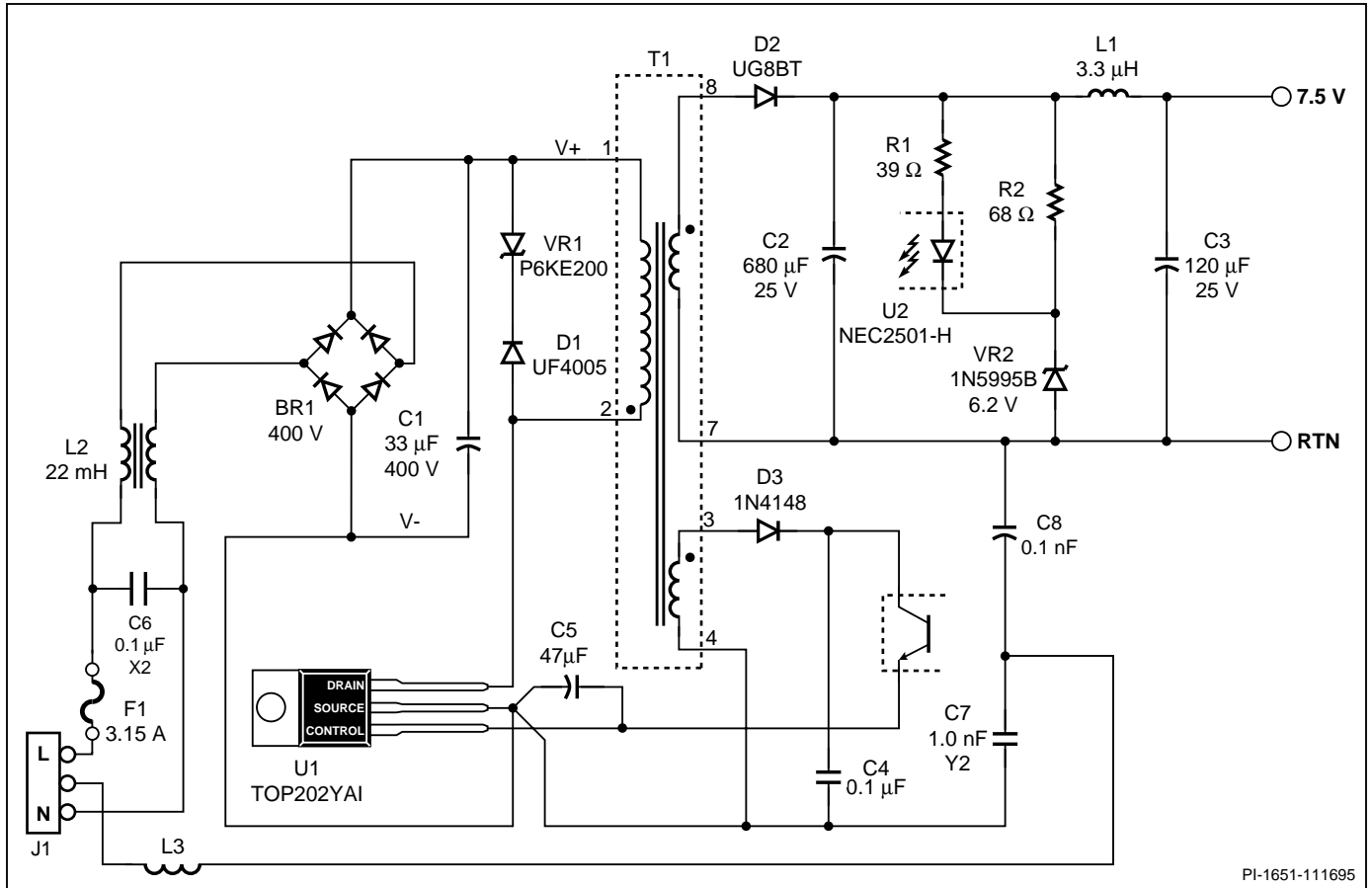
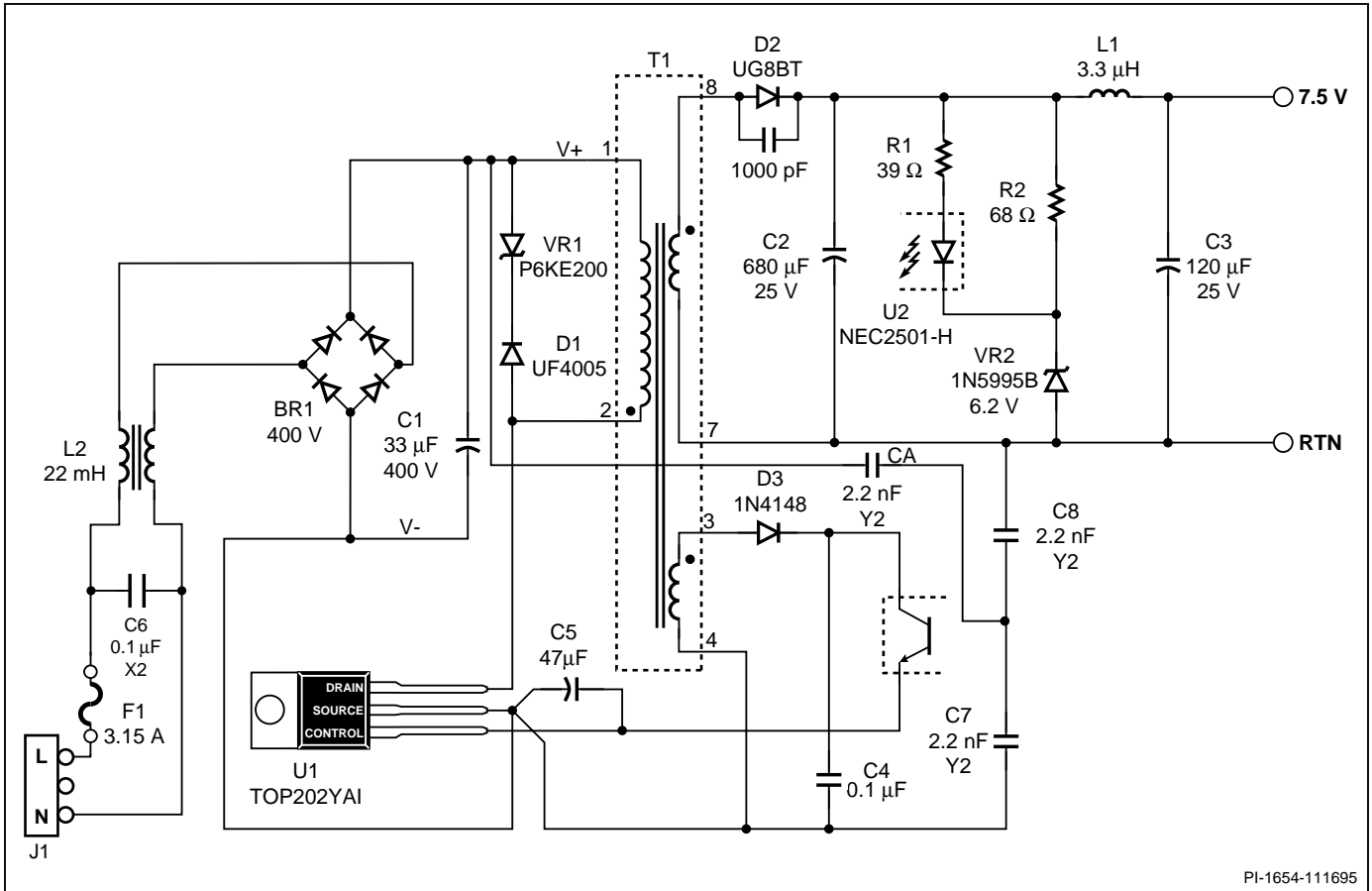


Figure 40. Typical 3-Wire TOPSwitch Power Supply and EMI Filter.







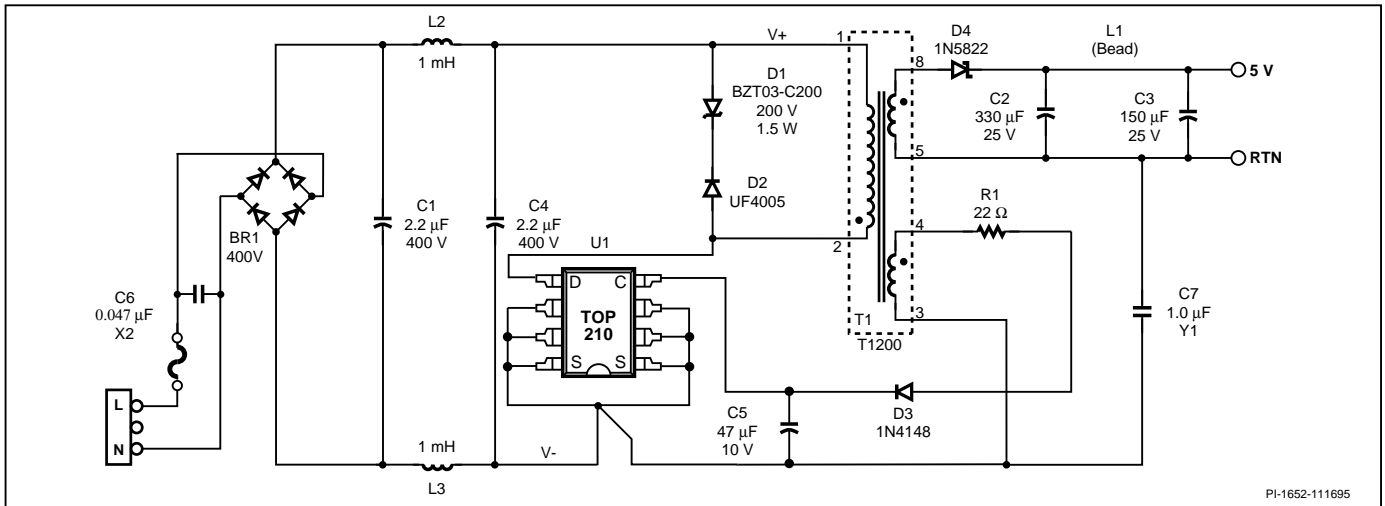
PI-1654-111695

Figure 42. 2-Wire TOPSwitch Power Supply and EMI Filter with Three Y-Capacitors For Video Applications and Reduced Radiated Emissions.

**Alternative Filter Approach Without Common Mode Choke**

An alternative filter for lower power (below 5 watt) applications is shown in Figure 43. This filter splits the high voltage energy storage capacitor to create a filter. Peak current in L2 and L3 is approximately half the peak current in bridge rectifier BR1 due

to capacitor C1. Differential mode attenuation is provided by C1, C4, L2, and L3. The secondary is AC coupled back to the primary by Y1-safety capacitor C7. Note that C7 can be replaced by two series connected Y2-safety capacitors with twice the value as shown.



PI-1652-111695

Figure 43. Low Power (Below 5W) TOPSwitch Supply Using Discrete Normal Mode Chokes.











## EMI Filter Layout Issues

Filter layout is extremely important to obtain the desired attenuation. Poor layout practice can cause conducted emissions to actually couple around the filter components directly into the AC mains conductors or cause radiated emissions.

Keep power stage and output components away from the EMI filter to prevent coupling around the filter. The best approach is to place the EMI filter at one end of a rectangular power supply shape and place the output at the other end as shown in Figure 47. Square power supply shapes should be avoided if possible since the power stage and the output components will be in close proximity to the EMI filter, allowing noise to couple directly into the mains.

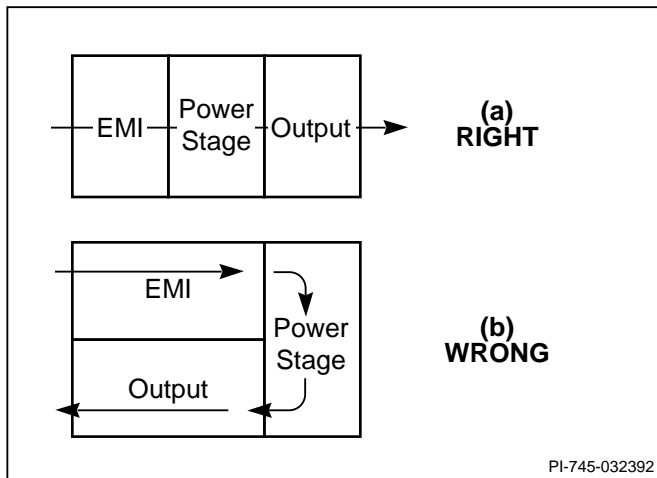


Figure 47. Power Supply Layout to Minimize Noise Coupling.

Capacitor lead length must be minimized as much as possible to reduce ESL. This includes the traces on the PC board leading up to the capacitor pads. Y-capacitor lead lengths and trace lengths are the most critical because the Y-capacitors couple high frequency currents (10 MHz to 200 MHz) back to primary circuitry. Figure 48 shows the right and wrong way to route PC traces to capacitors.

Locate the differential mode filter capacitor across the AC input conductors as close as possible to the power entry point.

### Practical Considerations

Successful EMI filter design begins with knowledge of the switching power supply noise sources generating differential mode and common mode conducted emissions which have

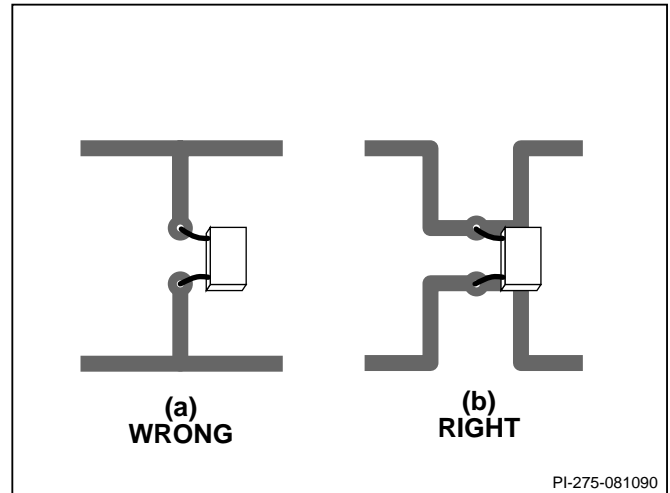


Figure 48. Bending the Bus to Minimize Resistive Effects.

been described in detail. The EMI filter must attenuate the emissions below the specification limit. Implementing a successful EMI filter is an iterative process. The basic steps include:

- Identify and attenuate the differential mode fundamental
- Identify and attenuate the common mode fundamental
- Identify other emissions over the spec limit.
- Determine whether each emission is differential mode or common mode.
- Use average or quasi-peak measurements on peak emissions to verify that the emission actually has insufficient margin compared to the EMI standard.
- Determine whether each emission is coupling around or passing through the EMI filter.
- Change the filter design or control the circuit source to attenuate each emission below the specification limit.
- Go back and check the earlier emission levels to make sure a change did not cause a different problem to occur.

### Differential mode Versus Common mode

The first frequency sweep for EMI conducted emissions on a power supply with no EMI filter will usually produce a spectrum as shown in Figure 49. The fundamental is outside the specification limit as well as some of the harmonics. Each

harmonic is composed of both differential mode and common mode emissions.

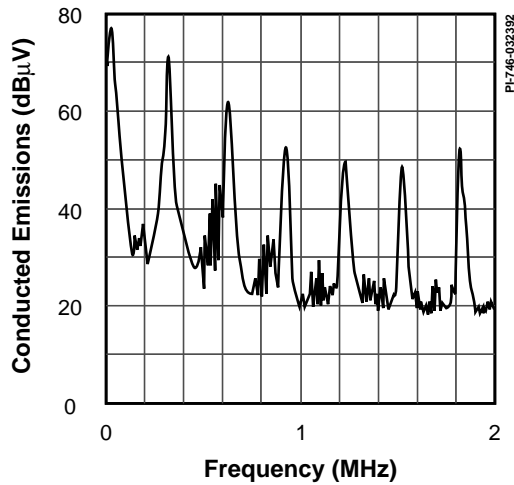


Figure 49. Typical Conducted Emissions Data without EMI Filter.

In Figure 50, a differential mode component with magnitude 70 dBµV is shown relative to a common mode component with magnitude 50 dBµV. 20 dB differential between the two components is actually an order of magnitude between the absolute values of the two components. The signals will add and superimpose if the phasing is correct but the overall effect on the measured signal level is slight (10% increase, or less than 1 dB).

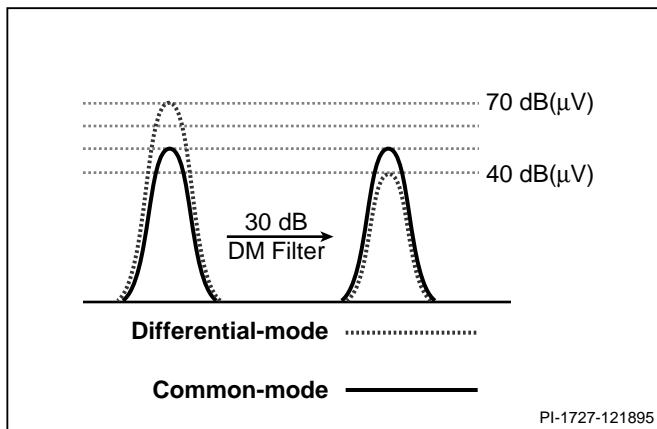


Figure 50. Superimposed Common Mode and Differential Mode Harmonics.

A differential mode filter with 30 dB attenuation at the harmonic frequency of interest will not attenuate the measured peak by 30 dB. The differential mode component will be attenuated from 70 dBµV to 40 dBµV, but the 50 dBµV common mode peak will now dominate the measurement. Further differential mode attenuation will have no effect on the measured harmonic because the signal is common mode. Common mode filtering

will have to be employed to reduce this harmonic any further.

### Using Splitters

Splitters combine the output signals of the two LISNs to determine if a specific emission is common mode or differential mode. Two splitters are required: an in-phase splitter (Mini Circuits Lab ZSC-2-2) where  $V_{OUT}$  is the sum of the two LISN signals and a 180 degree out-of-phase splitter (ZSCJ-2-2) where  $V_{OUT}$  is the differential between the two LISN signals<sup>(22)</sup>. The splitter setup is shown in Figure 51.

Splitter units are also available which allow switching between differential mode and common mode tests.

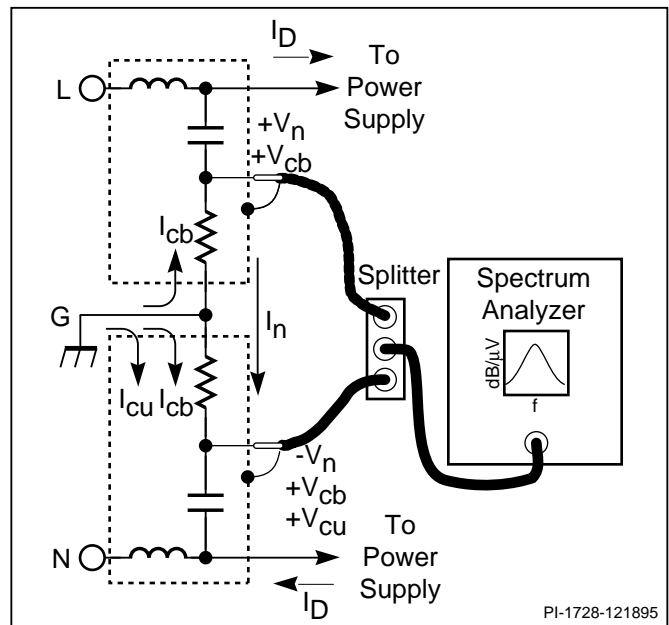


Figure 51. Separating Differential Mode from Common Mode Using a Splitter.

### Differential Mode Splitter Measurement

Differential mode emission current ( $I_D$ ) circulates from the power supply through the first LISN sense resistor (producing an in phase sense voltage), through the second LISN sense resistor (producing an out of phase sense voltage), and back to the power supply. The output voltage of the in-phase splitter will have no differential mode component because the opposite phased sense voltages effectively cancel. The output voltage of the 180 degree out-of-phase splitter will have differential mode components 6 dB higher than those measured directly at the LISN as the sense voltages are now effectively in phase and sum.

### Balanced Common Mode Splitter Measurement

Balanced common mode currents ( $I_{CB}$ ) are defined as currents with similar amplitude and phase that flow from ground through

each LISN sense resistor (producing sense voltages with the same phase) and through the AC input to the power supply. The output voltage of the in-phase splitter will have balanced common mode components 6 dB higher than those measured directly at the LISN because the in phase sense voltages effectively add. The output voltage of the 180 degree out-of-phase splitter will have no balanced common mode components as the sense voltages are now effectively out of phase and cancel.

**Unbalanced Common Mode Splitter Measurement**

Unbalanced common mode currents ( $I_{cu}$ ) flow from ground through either LISN sense resistor. Unbalanced common mode currents are found when the EMI filter does not have balanced impedance in each leg or when noise from the power path returns from the AC mains asymmetrically through one side of the EMI filter (typically caused by asymmetric parasitic capacitance). The output voltage of the in-phase splitter will have unbalanced common mode components equal to those measured directly at the LISN because there is no cancellation. The output voltage of the 180 degree out-of-phase splitter will also have unbalanced common mode components equal to those measured directly at the LISN for the same reason.

The results of using the two splitters on each of the three types of emissions is shown in Table 5.

LISN OUTPUT	IN PHASE	180° OUT OF PHASE
Differential ( $V_D$ )	0	$V_D + 6 \text{ dB}$
Balanced Common-mode ( $V_{cb}$ )	$V_{cb} + 6 \text{ dB}$	0
Unbalanced Common-mode ( $V_{cu}$ )	$V_{cu}$	$V_{cu}$

Table 5. Splitter Signal Levels.

**EMI Filter Component Measurements**

No EMI filter component is perfect. At some frequency all components “give up” their basic characteristic to the effects of parasitics.

Measure all capacitors. Identify specifically the self-resonant frequency (above which the capacitor looks like an inductor) and effective Q.

Measure all chokes. Identify specifically for the self resonant

frequency (above which the choke behaves like a capacitor) and effective Q. Identify multi-resonant behavior due to multiple layer winding.

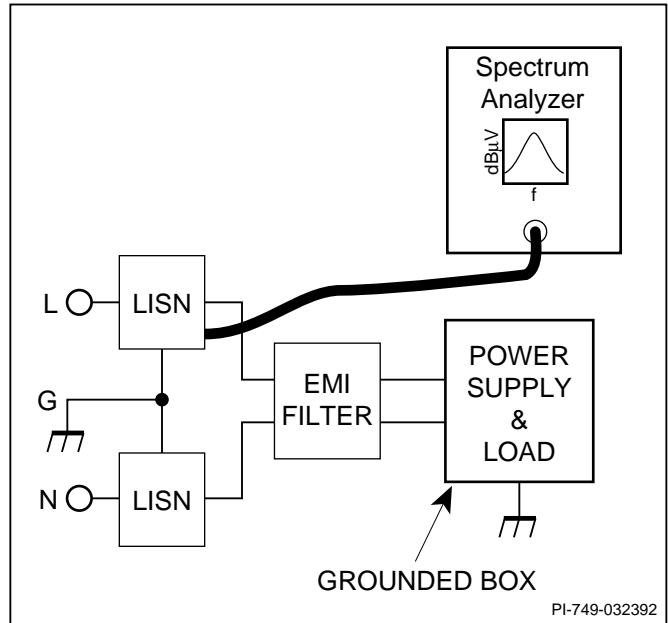


Figure 52. Typical Conducted Emissions System Test Set-Up with Grounded Box.

**Spatial Coupling**

As power supplies get smaller the EMI filter gets physically closer to circuitry acting as noise generators. High dv/dt voltage waveforms and high di/dt current loops generate fields which may spatially couple around the EMI filter and induce emission currents directly in the mains. Noise currents which couple around the filter must be distinguished from the noise currents which are passing through the filter.

One way to separate conducted emission currents is to place the power supply power circuitry and load within a grounded box as shown in Figure 52. The EMI filter is connected between the enclosed power supply and AC mains. The box will contain the fields, allowing conducted emission currents to be directly measured. This is especially effective when analyzing the fundamental. (Note that this technique is for investigative purposes only and must not be used for final test data).

The spatial coupling emissions can be reduced by containing the fields at their sources with local shields. Local shields over primary power circuitry such as the flyback transformer, primary damper, clamp diode, and TOPSwitch can be used to contain fields. Local shields can also be used over secondary circuitry such as output rectifiers. Shields can also be applied around the EMI filter although the preferred approach is to contain the field



at the source. Heat sinks can also be used as shields.

Another type of shield is a conductive plate approximately the same size as the printed circuit board. This plate can be connected to earth ground, primary reference, or secondary reference (depending on the safety insulation system chosen). This is attractive for applications connecting to the third wire ground but without a conductive enclosure.

### Lossy Beads

Small beads can be used in circuit leads to damp or eliminate high frequency ringing. Ferrite beads from Fair-Rite<sup>(21)</sup> are available in a variety of shapes. These beads feature low impedance at low frequency for minimal effect on the current waveform but have high impedance at high frequency with significant parallel resistance to damp and reduce ringing voltage waveforms.

### Grounding

Some applications connect the output voltage to earth ground. Others have no connection at all to earth ground. It is important to identify the earth ground connection expected for each application and to test in that configuration. EMI testing should be performed in both grounded and ungrounded configurations.

### Power Cord

The power cord resonance previously described can interfere with conducted emissions testing. Switch between two power cords of different lengths to separate power cord resonances from other conducted emissions.

### Miscellaneous Test Tricks

Terminate opposite LISN with 50 ohm terminator. The LISN sense impedance is actually determined by the termination and will change if not properly terminated.

Warm up equipment including Device Under Test (DUT) for at least 1 hour before testing so results will be repeatable.

Make sure analyzer sweep speed is low enough to capture the peaks of each harmonic. The bridge rectifier conducts current (both power and emission currents) for a short time compared with the full line cycle which effectively “pulse width modulates” the emission currents. Slower sweep speeds will collect enough data to accurately measure the peak of each current emission. A peak hold test can also be used to fill in the peaks in a few sweeps.

Peak measurements take the least amount of time but specifications are given in quasi-peak or average limits. Both quasi-peak and average measurement techniques give lower readings when compared to the peak value. If the peak value meets the average or quasi peak limit specification, there is no need to take further data with the average or quasi peak methods. To save test time, use the quasi-peak and average test

methods only when the measured peak value is close to or exceeds the target specification.

### Recommended Step-by-Step Procedure

- 1) Determine differential mode fundamental (and low frequency harmonics).
- 2) Calculate and select filter X-capacitance and target differential inductance. Select bobbin style common mode choke with sufficient differential mode inductance and AC current rating (use discrete chokes only for low power, under 5 Watts).
- 3) Measure impedance versus frequency for each component. Select components with resonant frequencies that do not coincide with waveform ringing frequencies in the power supply.
- 4) Use a nominal value AC source through the LISNs to provide power to the power supply. Use the 180 degree splitter to extract the differential mode fundamental current component. Measure the fundamental with slow sweep speed and measure peak value using maximum hold. Compare the measured differential mode fundamental with the calculated value. Use average or quasi-peak detection as required to properly compare measured reading with limits of the chosen standard. Increase X2-capacitance or select common mode choke with higher differential inductance if necessary.
- 5) Examine entire frequency range for differential mode components close to or in excess of the specification limit. Make measurements on both Line and Neutral LISN. Pay special attention to frequency ranges around measured component resonances and identified circuit ringing frequencies. Use average or quasi-peak detection as required to properly compare measured reading with limits of the chosen standard. Modify differential mode filter design if necessary.
- 6) Use in-phase splitter to extract the balanced common mode fundamental and low frequency harmonic current components. For 2-wire applications, place the largest value Y1-safety capacitor (subject to leakage current limitations but typically 1 nF) to output return. Two series connected Y2-safety capacitors (typically each with 2.2 nF value) can also be used. For 3-wire applications, place the largest value Y2-safety capacitor (subject to leakage current limitations but typically 1 nF up to 33 nF) from the power supply high voltage return to earth ground and then connect a 0.1 uF low voltage ceramic capacitor from earth ground to secondary return. **Y-capacitor leads must be very short to attenuate high frequency current emissions.** Measure balanced common mode fundamental and low frequency harmonics. Increase size of common mode choke if necessary. Select the smallest, widest bandwidth common mode choke (with sufficient RMS current ratings) that attenuates the balanced common mode fundamental to the desired level. Measure impedance versus frequency for each component. Select components with resonant frequencies that do not coincide with

waveform ringing frequencies in the power supply.

7) Examine entire frequency range for balanced or unbalanced common mode components close to or in excess of the specification limit. Make measurements on both Line and Neutral LISN. Pay special attention to frequency ranges around measured component resonances and identified circuit ringing frequencies. Use average or quasi-peak detection as required to properly compare measured reading with limits of the chosen standard. Modify common mode filter design if necessary.

8) Remove splitters. Measure signals from both Line and Neutral over the entire frequency range. Emissions above 1 Mhz are usually common mode. Solve spatial coupling effects using additional Y-capacitors (see Figures 41 and 42), improved transformer construction (see Figures 35 and 36), higher bandwidth two section common mode choke (see Figure 21), additional high frequency common mode choke (see Figure 27) or shielding techniques (Figures 44-46).

9) Perform final test with secondary connected to ground and also with secondary isolated from LISN ground.

## BIBLIOGRAPHY

- (1) FCC Harmonizes Its ITE Rules with C.I.S.P.R. Requirements, Compliance Engineering, pp 7, January/February, 1994.
- (2) European Standard EN55022, "Limits and Methods of Measurement of Radio Interference Characteristics of Information Technology Equipment", Cenelec, 1994.
- (3) "Limits and Methods of Measurement of Radio Interference Characteristics of Information Technology Equipment", C.I.S.P.R Publication 22, 1993.
- (4) Dash, D. and Straus, I.; "EMC Regulations in Germany"; Compliance Engineering 1994 Reference Guide, pp. 85-96, Compliance Engineering, Boxborough, MA.
- (5) Regulation Vfg 243/1991, "Radio Interference Suppression of Radio-Frequency Equipment for Industrial, Scientific, Medical (ISM) and Similar Purposes and Equipment used in Information Processing Systems; General License", German Federal Minister for Post and Telecommunications.
- (6) Amending Regulation Vfg 46/1992, German Federal Minister for Post and Telecommunications.
- (7) "Meeting the EC Emissions Requirement", Handbook of EC EMC Compliance, Compliance Design Incorporated, 1993, pp 7-15.
- (8) M. Engelson, "EMI Applications using the Tektronix 2712 Spectrum Analyzer", Application Note, Tektronix, 1993.
- (9) M. Engelson, "Measures of EMC, A Review of Basic EMC Measurement Techniques and Standard Practices", EMC Test and Design, November/December 1990.
- (10) Elliott Laboratories, 684 W. Maude Avenue, Sunnyvale, CA, 94086 (408-245-7800).
- (11) IEC1950, "Safety of Information Technology Equipment Including Electrical Business Equipment", Second Edition, International Electrotechnical Commission, 1991.
- (12) Evox Rifa, "Capacitors 1995", Catalogue, Evox- Rifa Inc., 100 Tri-State International, Suite 290, Lincolnshire, Illinois 60069, (1-708-948-9511).
- (13) European Standard EN 132400, "Sectional Specification: Fixed Capacitors for Electromagnetic Interference Suppression and Connection to the Supply Mains", (Assessment Level D), Cenelec, 1994.
- (14) Murata Electronics, North America, Inc., "Disk Ceramic Capacitor Catalog no. C-OS-C 1991", Murata Erie North America, 2200 Lake Park Drive, Smyrna, Georgia 30080, (1-800-831-9172).
- (15) Roederstein, "EMC Radio Interference Suppression Components", 1991, (704-872-8101).
- (16) Tokin, "EMC Line Filters Vol. 2 CD-07JE", Tokin America Inc., 155 Nicholson Lane, San Jose, CA 95134 (1-408-432-8020).





- 
- (17) Tamura , "Tamura Common Mode Coils for AC Line & EMI Filtering, 0.1 to 50 Mhz", Electronic Engineers Master Catalogue (EEM), 1995, Tamura Corporation of America, P.O. Box 892230, Temecuca, CA 92589, (1-909-699-1270).
- (18) Panasonic Industrial Company, "Panasonic Line Filters", Digikey Catalog 961, January-February 1996 (1-800-344-4539). Also "Panasonic Inductors/Transformers 94/95" #9404155S1, Panasonic Industrial Co., 2 Panasonic Way (7H-3), Secaucus, New Jersey 07094 (1-201-348-4630).
- (19) TDK Corporation of America, "TDK EMI Prevention Components", EVE-005B, 1993, (708-803-6100).
- (20) Murata Erie, "EMI/RFI Filter Catalog", E-06-A, 1993, (1-800-831-9172).
- (21) Fair-Rite, "Fair-Rite Soft Ferrites" Catalogue, 12th edition, Fair-Rite Products Corporation, P.O. Box J, 1 Commercial Row, Wallkill, NY 12589, (914-895-2055).
- (22) Mini-Circuits Laboratories, 13 Neptune Ave., P.O. Box 350166, Brooklyn, NY 11235, (718-934-4520).

# NOTES



---

# NOTES

Power Integrations reserves the right to make changes to its products at any time to improve reliability or manufacturability. Power Integrations does not assume any liability arising from the use of any device or circuit described herein, nor does it convey any license under its patent rights or the rights of others.

PI Logo and **TOPSwitch** are registered trademarks of Power Integrations, Inc.  
©Copyright 1994, Power Integrations, Inc. 477 N. Mathilda Avenue, Sunnyvale, CA 94086

---

**WORLD HEADQUARTERS**

Power Integrations, Inc.  
477 N. Mathilda Avenue  
Sunnyvale, CA 94086  
USA  
Main: 408•523•9200  
Customer Service:  
Phone: 408•523•9265  
Fax: 408•523•9365

**AMERICAS**

For Your Nearest Sales/Rep Office  
Please Contact Customer Service  
Phone: 408•523•9265  
Fax: 408•523•9365

**EUROPE & AFRICA**

Power Integrations (Europe) Ltd.  
Mountbatten House  
Fairacres  
Windsor SL4 4LE  
United Kingdom  
Phone: 44•(0)•1753•622•208  
Fax: 44•(0)•1753•622•209

**JAPAN**

Power Integrations, K.K.  
Keihin-Tatemono 1st Bldg.  
12-20 Shin-Yokohama 2-Chome, Kohoku-ku  
Yokohama-shi, Kanagawa 222 Japan  
Phone: 81•(0)•45•471•1021  
Fax: 81•(0)•45•471•3717

**ASIA & OCEANIA**

For Your Nearest Sales/Rep Office  
Please Contact Customer Service  
Phone: 408•523•9265  
Fax: 408•523•9365

**APPLICATIONS HOTLINE**

World Wide 408•523•9260

**APPLICATIONS FAX**

Americas 408•523•9361  
Europe/Africa  
44•(0)•1753•622•209  
Japan 81•(0)•45•471•3717  
Asia/Oceania 408•523•9364

