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# FILTER DESIGN PROCEDURE OF CONDUCTED EMI BASED ON NOISE IMPEDANCES

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***Abstract:*** *This paper presents a filter analysis of conducted Electro-Magnetic Interference (EMI) for switching power converters (SPC) based on noise impedances. The EMI characteristics of the SPC can be analytically deduced from a circuit theoretical viewpoint. The analytical noise model is first investigated to get a full understanding of the EMI mechanism. It is shown that with the suitable and justified model, filters pertinent to EMI noise are investigated. The EMI noise is identified by time domain measurements associated with an isolated half-bridge ac–dc converter. The practical filters like LC filter, π filter and Complete EMI filters are investigated. The analysis and results proposed can provide a guideline for future effectiveness of filtering schemes in SPC. Experimental results are also included to verify the validity of the proposed method. The results obtained satisfy the Federal Communications Commission (FCC) class A and class B regulations.*

***Keywords:*** *Common-mode (CM), Differential-mode (DM), Eelectromagnetic compatibility (EMC), Conducted EMI, Switching power converters (SPC).*

## 1. Introduction

EMI emission is always a great concern for the modern trends in power electronics by fast switching large amount of current at high voltages and high frequency in switching devices like Metal Oxide Semiconductor Field Effect Transistor (MOSFET), Insulated Gate Bipolar Transistor (IGBT) and Gate Turn Off thyristor (GTO). The frequency range of the conducted emission limit is from 450 kHz to 30 MHz for the FCC class A and class B regulations and 150 kHz to 30 MHz for the German Verband Deutscher Elektroniker (VDE) regulations. Radiated emissions are generally measured at much higher frequencies, namely beyond 30 MHz up to several GHz.

Conducted EMI in power converters has become increasingly important due to its wide applications. Recently, EMI problems have become more serious for power electronic systems. The process of EMI mitigation normally involves filter designing. Traditional cut-and-try approach has been abandoned by engineers for its time-consuming and inefficient defects. Therefore, the effective solutions for conducted EMI noise analysis and suppression have become great concerns.

Conventionally, the conducted EMI noise is caused by two mechanisms: Differential Mode noise and Common Mode noise. Furthermore, the EMI filter should be designed for CM noise and DM noise, respectively. So, the first step in designing the EMI filter is to separate the CM and DM noise.

A two way 0˚ combiner (ZFSC-2-6-75) and a two way 180˚ power combiner (ZFSCJ-2-1) are used to measure the total, DM and CM noise via election of a three way built-in switch. Moreover, in order to suppress the conducted EMI noise, the proper design of EMI filter is necessary. In general, the DM noise is related to switching current and the CM noise is related to capacitive coupling of switching voltage with Line Impedance Stabilizing Network (LISN), which is used in standard conducted EMI measurement.

Accurate modeling of EMI noise generation and propagation in power converters are the first step in predicting and managing the EMI noise in a system. The effective EMI prediction often relies on the engineer's experience or extensive numerical simulation models [1]–[3]. Owing to the effectiveness of filtering the conducted EMI noise separately by common-mode and differential-mode, each mode of the noise is dealt with the respective section of an EMI filter [4]. New techniques for designing EMI filters have been developed recently by the use of a noise separator, which can be used to separate DM and CM noise [5].

## 2. Model Description

Figure. 1 shows the configuration of the conducted EMI measurement for an isolated half-bridge ac–dc converter. The power source is provided through a LISN, which is required by the conducted EMI measurement.

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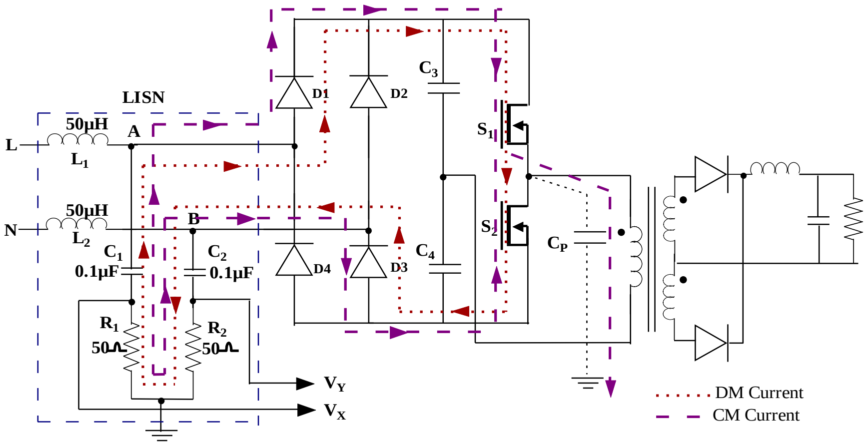


**Figure.1.** Conducted EMI coupling path of Half-bridge isolated ac–dc converter

## 2.1. DM/CM coupling

LISN contains two 50 µH inductors - L1 and L2, two 0.1 µF capacitors - C1 and C2 and two 50 Ω resistors - R1 and R2. For noise frequency, the two inductors L1 and L2 present high impedance and two 0.25 µF capacitors C1 and C2 present low impedance. The noise voltage coupled through the two resistors R1 and R2 is counted as the conducted EMI. For power-line frequency, the inductors are essentially short, the capacitors are essentially open, and the power is passed through to supply the test system.

According to conventional theory, DM noise is defined as the voltage difference between the two LISN resistors, i.e., $V_X - V_Y$ and CM noise is defined as the average voltage of the two LISN resistors, i.e., $(V_X+V_Y)/2$. For conducted EMI noise frequency, two 50µH inductors present high impedances and two 0.1µF capacitors present small impedances. The voltages measured across the two 50Ω impedances are defined by the conducted EMI, $V_X$ is the line-side EMI, and $V_Y$ is the neutral-side EMI. DM noise voltage and CM noise voltage are expressed as

$$V_{CM} = (V_X + V_Y)/2 \tag{1}$$

$$V_{DM} = (V_X - V_Y)/2 \tag{2}$$

During normal operation, dc link is clamped at a fixed voltage by the capacitances $C_3$ and $C_4$. When ac side line voltage is larger than the capacitance voltage, the diode bridge is ON, and when line voltage is smaller than the capacitance voltage, the diode bridge is OFF. DM noise current coupling path as indicated in Figure 1 passes through the input rectifier diodes (D1 and D3).

The measured DM noise fluctuates with time because of the rectifier diodes ON and OFF during half a supply cycle. But CM noise is independent of the conduction state of the rectifier. The noise magnitude should be larger when diodes are “ON” because the noise current can couple R1 and R2 with less impedance in the path. The DM noise is higher when the rectifier is OFF than that when rectifier diodes are ON.

## 2.2. EMI Noise Separator

The noise separator is built using the principle of noise rejection accomplished by using power combiners. Figure.2 (a) shows the diagram depicting the basic concept of a noise separator in which the two signals derived from the LISN, consist of both CM and DM noises. One of the signals is the vector sum of the two modes of noise (CM+DM) and the other signal is the vector difference of the two modes of noise (CM-DM). The CM current is evenly divided between the two input terminals, which is often true except in extreme cases.

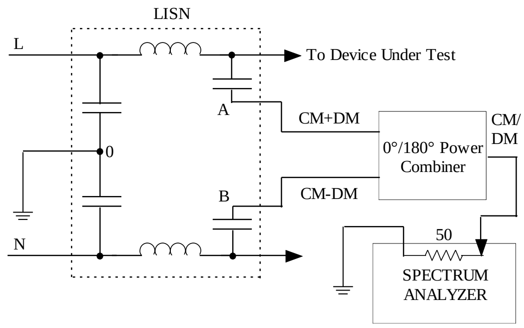


**Figure.2 (a)** 0°/180° Power Combiner

There are four parameters defined to evaluate a noise separating network, that is, the common-mode insertion loss (CMIL), differential-mode insertion loss (DMIL), common-mode rejection ratio (CMRR) and differential-mode rejection ratio (DMRR). The transmission coefficient of noise separating network are defined as

$$S_{21} = 20 \log (V2/V1)\ (dB) \tag{3}$$

where V1 and V2 are the input and output of the network. As the transmission coefficient $S_{21}$ is CMIL/DMIL, the V1 and V2 represent the same mode signals, and as $S_{21}$ is CMRR/ DMRR, V1and V2 represent different mode signals [6]. It should be mentioned that the insertion loss should be not more than 5dB and the rejection ratio should be not less than 40 dB. A spectrum analyzer and 0°/180° combiners are used to measure the characterization parameters of noise separating network [7]. The CMIL result of the high-performance noise separating network is shown in Figure.2. (b). As the frequency goes up, the CMIL declines slightly but remains above -2dB. The DMRR result of the high-performance noise separating network is shown with a good performance in Figure.2. (c). The DMRR rises with increasing frequency, and at 30MHz, the maximum frequency for conducted EMI noise measurement, the DMRR successful remains below -40dB. The measurement results prove that the noise separating network can separate the conducted EMI noise efficiently.

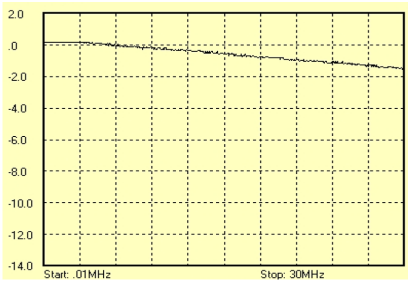


**Figure.2 (b)** CMIL of the noise separating network

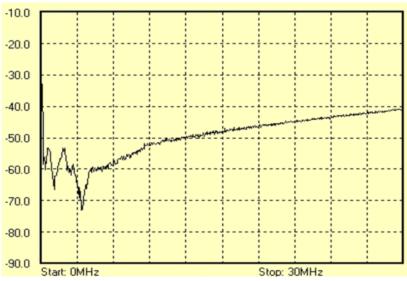


**Figure.2 (c)** DMRR of the noise separating network

## 2.3. Model Implementation

Figure.3 shows an equivalent circuit model for description of the essential coupling paths between the converter and the LISN three physical terminals L, N, and G. Assume a linear model for the converter physical circuit by a simple combination of three impedance elements together with three voltage sources. The overall noise is equal to the sum of all components from individual analysis for all harmonics for a linear model. Dealing with the EMI noise in this way, all details of the internal responses with the converter is lost, and only the external responses at the LISN input terminals are maintained [8].

The characteristics of the LISN are well defined; each phase of the LISN can be represented by 50Ω impedances. EMI coupled from three voltage sources through three impedances, i.e., $V_{SDM}$ and $Z_{SDM}$ are DM noise voltage source and impedance, $V_{SCM1}$, $V_{SCM2}$, $Z_{SCM1}$, and $Z_{SCM2}$ indicate CM noise voltage sources and impedances respectively. The LISN is represented by two resistors $R_N$.

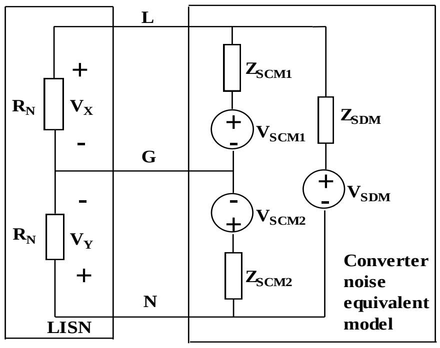


**Figure.3** Noise model for essential coupling path

These voltage sources depend on the high dv/dt and high di/dt slew rates and circuit parasitic parameters, device package, and layout. Normally $Z_{SCM1}$ and $Z_{SCM2}$ are high source impedances because they are associated with parasitic capacitance to the earth. Assume that $Z_{SCM1}$ >>50 and, $Z_{SCM2}$ >>50 then obtain Figure 4.

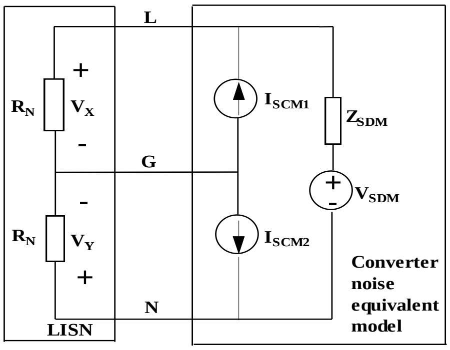


**Figure.4** Simplified noise model

The CM currents are expressed as

$$I_{SCM1} = \frac{V_{SCM1}}{Z_{SCM1}} \quad (4)$$

$$I_{SCM2} = \frac{V_{SCM2}}{Z_{SCM2}} \quad (5)$$

The effectiveness of an EMI filter depends not only on the filter itself but also on the noise source impedance. For CM noise, the source is modeled by a current source in parallel with high source impedance $Z_s$. For DM noise, the source is modeled by a voltage source in series with low impedance or a current source in parallel with high impedance dependent on the state of the input rectifier diodes. When two of the rectifier diodes are conducting, the noise source is modeled by a voltage source in series with a low impedance source $Z_s$.

When all of the four diodes are cut off, the noise is modeled by a current source in parallel with a high source impedance $Z_p$. The DM noise equivalent circuit therefore fluctuates between these two models at two times the line frequency. $Z_s$ is associated with wire inductance and resistance, and $Z_p$ is associated with diode parasitic capacitance. These source impedances depend on parasitic parameters and are therefore package dependent. Although the source impedances can be measured, this is practically difficult. Two simplified equivalent circuits, one for CM emissions and one for DM emissions are obtained.

## 2.4. EMI Model for Symmetrical Circuit

The model in Figure. 3 contain all essential coupling paths and circuit parameters that can provide a full picture of the EMI conduction and coupling mechanism. The DM voltage (with $R_N$= 50Ω) can be written as

$$V_{DM} = \frac{100V_{SDM}}{Z_{SDM}+100} + \frac{50Z_{SDM}}{Z_{SDM}+100}(I_{SCM1} - I_{SCM2}) \quad (6)$$

From the equation (6), total DM noise has two parts: 1) the first part is determined by DM noise source $V_{SDM}$ and impedance $Z_{SDM}$ called, Intrinsic Differential Mode (IDM) noise and 2) the second part is the difference of two CM current sources through the DM impedance and LISN, called Non-Intrinsic Differential Mode (NIDM) or Mixed Mode (MM) noise. It is clear from that, the MM noise is caused by unbalanced CM current which flows through the two LISN branches. By following the above deducing, the DM and CM noise source model can then be reduced to a simple two-port lumped circuit model, as shown in Figure. 5. $Z_N$ is $2R_N$ in the case of DM and $0.5R_N$ in the case of CM, $Z_S$ is $Z_{SDM}$ in the case of DM and $0.5Z_{SCM1}$ in the case of CM, $V_S$ is $V_{SDM}$ in the case of DM and $V_{SCM1}$ in the case of CM.

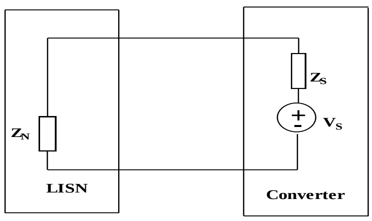


**Figure.5** Simplified EMI noise lumped circuit model

For reducing the complexity, the EMI issues using very simple models always consider the combination of impedance $Z_S$ and source $V_S$. These sources and impedances depend on parasitic parameters and, therefore, are determined by converter package and layout. Although the sources and impedances can be measured by a particular method [9], [10], this is not in the scope of this paper. The major components of the DM noise source impedance are the turned-on resistance of rectifying diodes and the equivalent series resistance (ESR) and equivalent series inductance (ESL) of the bulk capacitor. The major components of CM noise source impedance are the unintentional capacitance between the switching device and its heat-sink, parasitic capacitance between the heat-sink and the grounded chassis, and parasitic capacitances between other devices or wires, which carry pulsating voltage waveform, and the grounded chassis. Therefore, $Z_{SDM}$ is no more than several ten ohms, but $Z_{SCM1}$ and $Z_{SCM2}$ are always bigger than several hundreds ohms in a frequency range of 150 kHz–30 MHz. The DM noise source is always considered the pulse current of the power switch device. The CM noise source is associated with the high rate of the changing voltage put on the parasitic capacitances between the circuit and ground. However, owing to the complexity of the EMI noise coupling, theoretical modeling of parasitic elements has been a very challenging task because the parasitic elements are difficult to identify and the parasitic elements may be physically inaccessible inside the module package.

## 3. Filters Analysis for Noise Suppression

After the filter circuit has been selected, the component values in that circuit must be determined. The Y-capacitors can easily be determined from the leakage current limit, which is given in the safety standards applicable to the particular equipment. With known Y-capacitors, the CM inductance can be calculated from:

$$L = \frac{1}{4\pi^2 f_c^2 C} \quad (7)$$

where *fc* should be the corner frequency, which can be defined as the intersection between the 0 dB line and a line with slope 40 dB/decade that is tangent to the required CM Insertion loss. Similarly, the DM corner frequency is determined from the required DM Insertion loss. When the CM inductance is being calculated, the capacitance C in (7) should be twice the $C_Y$ value. The CM inductance consists of the CM choke's inductance and half of the inductance of the DM inductors, if these are decoupled inductors. If the DM inductors are coupled, they have negligible CM inductance and then the CM inductor should have an inductance, equal to the required CM inductance, obtained by (7). The selection of the filter components involves not just the calculation of suitable capacitance and inductance values, but also other considerations, such as inductor core material, the dielectric of the capacitor, voltage, and current ratings, the voltage drop at line frequency, size, weight, etc. The $L_{DM}$ is a DM choke $C_X$ is a DM capacitor or (called "X" capacitor).

When filtering DM noise, an X capacitor is used to shunt noise from one line to the other through the low impedance of the capacitor, thus returning the noise to the EMI source. For DM noise, DM choke should have suppressing effect when the two paths are unbalanced. Even though current flow is unbalanced, the magnitude of each current is attenuated already by the DM choke and therefore the DM noise is suppressed. To suppress the DM noise effectively, the capacitor is always used together with the DM chokes. The measurement of the conducted EMI noises is obtained using PSPICE software simulation. The waveforms of DM and CM voltage spectrum can be sampled in time domain during a full 50 Hz line cycle. $V_{CM}$ and $V_{DM}$ can be obtained by performing a Fast Fourier Transformer (FFT) on the time domain waveforms. After considering the measure bandwidth of 10 kHz and normalized in dBμV [11], the spectra of total EMI is given in Figure. 6.

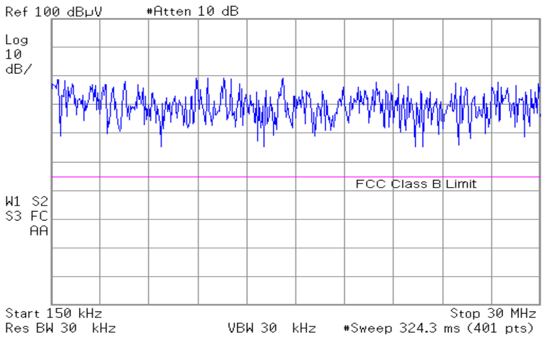


**Figure.6** Total noise spectrum of the ac–dc SPC without filter

The fluctuation of EMI at a particular frequency can be displayed by using the zero span mode of a spectrum analyzer. Using this mode, a magnitude of the selected noise frequency is displayed with respect to time. Another important issue is that the insertion of the filter should not affect the stability of the SPC. The EMI filter will not affect the power converter's stability and dynamics only if the output impedance of the DM-filter part is much less than the input impedance of the converter from dc to the gain crossover frequency of converter's control loop [12]–[15].

This requires that the DM filter should be well damped to avoid impedance peaking and consequent possible violation of this criterion. To effectively suppress the DM noise, the balancing capacitor should provide balanced impedance such that the currents flowing through the two LISN resistors R1 and R2 are the same, irrespective of $C_p$'s charged condition. Under this condition, the choice of $C_x$ value depends on the frequency and the filter topology. The value of $C_x$ can be calculated by using the following formula.

$$X_c = 1/(2\pi f C_x); \quad X_c \ll 50\,\Omega \qquad (8)$$

f = starting frequency of EMI noise;
(150 kHz for VDE and 450 kHz for FCC)
The filter attenuation is defined as

$$A_{filter} = \frac{V_{noise}(without filter)}{V_{noise}(with filter)} \qquad (9)$$

$V_{noise}$ without filter is measured at the LISN output, when no EMI filter element is added. $V_{noise}$ with filter is measured at the LISN output when a filter is added. The topology should be selected according to the source and load impedances. The selection of the filter components involves not just the calculation of suitable capacitance and inductance values, but also other considerations, such as inductor core material, the dielectric of the capacitor, voltage, and current ratings, the voltage drop at line frequency, size, weight, etc. This approach considered filter topology only but not mentioned the relationship of noise source impedance (Zs) and noise load impedance (LISN), which determine the EMI filter performance significantly [16].

## 4. EMI Filter Design Procedure

In this section, a method to design an EMI filter using the maximum and minimum CM and DM noise impedances is introduced. The ac–dc half-bridge converter is used as an example. The objective was to design an EMI filter, so that the SPC would pass the FCC 15 Class B regulations for conducted EMI. The CM and DM filters were designed separately. The maximum or minimum worst case noise impedance was considered for each filter. After completion of the CM and DM filters, they were put together to make the complete filter [17].

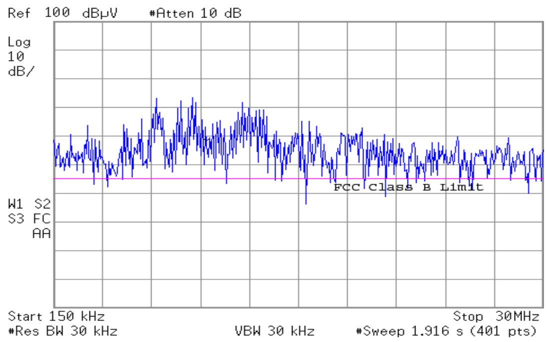


**Figure.7** CM noise spectrum without a DM filter

The procedure to design the EMI filter is summarized as follows.

1) Separate the CM and DM noise spectrum of the SPC.
2) Measure the noise voltage, $V_{noise}$, with and without a simple filter (eg., a capacitor), and then use the attenuation to calculate the maximum CM and DM noise impedances for the frequency range under consideration (0.15–30 MHz for the FCC 15).
3) Design the EMI filter using the maximum magnitude, or minimum magnitude of the noise impedance, which ever yields the least attenuation.
4) Test the completed EMI filter.

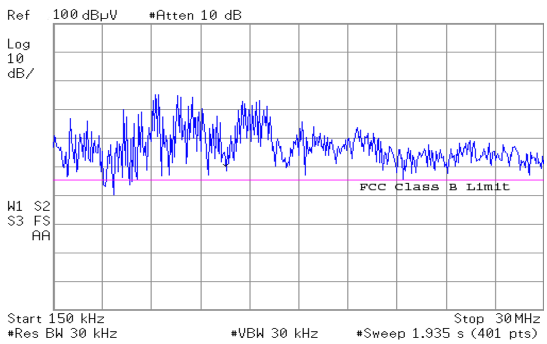


**Figure.8** DM noise spectrum without a DM filter

The CM and DM noise spectra of the unfiltered ac-dc SPC, operating at full load, are illustrated in Figure.7 and Figure.8, respectively. The impedance of the CM inductor is $Z_{fCM}$. $R_{loadCM}$ is the equivalent resistance of the LISN. $I_{sCM}$ represents the SPC noise source and $Z_{sCM}$ represents the noise source impedance. From the test results, it is noted that the CM noise was the dominant factor. Calculations were made at different frequency points using measurements of the noise voltage $V_{noiseCM}$ along with (9) and (10)–(13).

1) When $A_{TCM} \geq 10$

$$\left|Z_{sCM}\right|_{Max} \approx \left| R_{loadCM} + \frac{\left|Z_{fCM}\right|}{\left|A_{TCM}\right|} \right| \tag{10}$$

$$\left|Z_{sCM}\right|_{Min} \approx \left| R_{loadCM} - \frac{\left|Z_{fCM}\right|}{\left|A_{TCM}\right|} \right| \tag{11}$$

2) When $A_{TCM} < 10$

$$\left|Z_{sCM}\right|_{Max} \approx \left| R_{loadCM} + \frac{\left|Z_{fCM}\right|}{\left|A_{TCM} - 1\right|} \right| \tag{12}$$

$$\left|Z_{sCM}\right|_{Min} \approx \left| R_{loadCM} - \frac{\left|Z_{fCM}\right|}{\left|A_{TCM} + 1\right|} \right| \tag{13}$$

The maximum possible value and minimum possible value of the DM noise impedance from 0.15 to 2.5 MHz were calculated at different frequency points using measurements of the noise voltage $V_{noiseDM}$ along with (9), (14), and (15). $Z_{sDM}$ is the DM noise impedance of the SPC and $Z_{fDM}$ is X capacitor's impedance.

$$Z_{sDM}\big|_{Min} = Z_{fDM}\left|A_{TDM} - 1\right. \tag{14}$$

$$Z_{sDM}\big|_{Max} = Z_{fDM}\left|A_{TDM} + 1\right. \tag{15}$$

## 4.1. CM Filter Design

In order to design the CM filter, need to know the following: 1) the attenuation required to make the CM noise spectrum pass the EMI standards at the frequencies of interest and 2) the worst case, maximum, or minimum noise impedance for the frequency range of interest. For the CM part of the EMI filter, two basic topologies can be selected. In the first topology, the CM inductor faces the input of the SPC, as illustrated in Figure.9. In the second topology, the Y capacitor faces the input of the SPC, as illustrated in Figure.10. Using the equivalent circuits of Figure.9 and Figure.10 the CM noise voltage across $R_{loadCM}$ was calculated when the filter was added. Two topologies were tested. For the test, the CM inductance used was 100 μH and the Y capacitor used was 330 nF. From the results, it was concluded that the topology of Figure. 9 should be used because it provided more attenuation at high frequencies than the topology shown in Figure. 10. This result also emphasizes the fact that noise impedance has a significant effect on the performance of the EMI filter. In order to determine the lowest corner frequency required for the CM filter, (16) was used.

$$f_C = \frac{F_0}{\sqrt{A_{TCMreq}}} \tag{16}$$

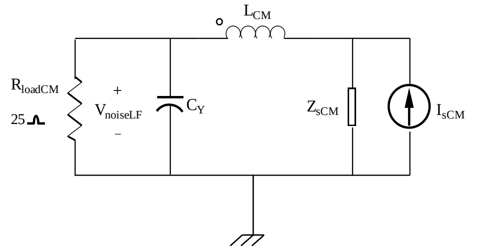


**Figure.9** CM noise equivalent circuit with the CM inductor at the input of the SPC

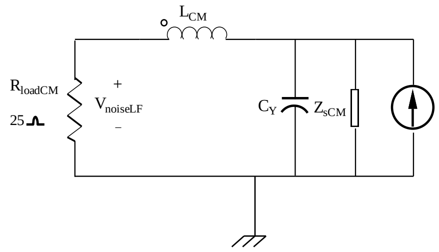


**Figure.10** CM noise equivalent circuit with the Y capacitor at the input of the SPC

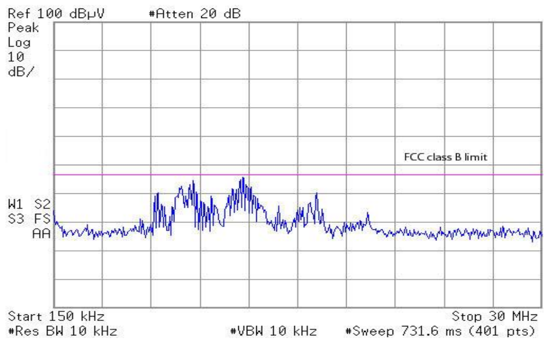


**Figure. 11** Test results of the CM noise spectrum after the CM filter designed based on maximum noise impedance was added

$A_{TCMreq}$ is the required attenuation at frequency $F_0$, and $f_c$ is the corner frequency of the CM filter. The lowest calculated filter corner frequency was $f_c$ = 94 kHz, using $F_0$ = 583 kHz at $A_{TCMreq}$ = 38. To select the value of the CM inductor, the maximum noise impedance was used because it provides the least attenuation for the filter in Figure. 9, where the inductor faces the input of the SPC. If the phase angle of the noise impedance is neglected and to guarantee that CM noise impedance has little effect on the performance of the EMI filter, the impedance of the CM inductor at the frequency of interest must be much larger than the maximum magnitude of the CM noise impedance. In fact, to make sure the desired attenuation is achieved, the CM inductor's impedance must be at least double the maximum magnitude of the CM noise impedance.

Therefore, the CM inductor's impedance at $F_0$ must be larger than 2 x 205Ω = 410Ω, which corresponds to an inductance greater than 2 x 55.9μH = 111.8μH. The inductance chosen can be any reasonable value above 111.8 μH, so in order to leave some margin; a 200 μH inductance was selected. The impedance of this CM inductor at 583 kHz is 732Ω which is approximately three times the maximum magnitude of the CM noise impedance. Ignoring the phase angle of the CM noise impedance (i.e., considering the CM noise impedance as purely resistive), the attenuation of the topology in Figure.9, can be expressed by (17) where $V_1$ and $V_2$ are given by (18) and (19), respectively. Parameters $Z_1$ and $Z_2$ are given by (20) and (21), respectively

$$|A_{TCM}| = \left|\frac{V_2}{V_1}\right| \tag{17}$$

$$V_1 = \frac{Z_1 Z_2}{(Z_{LCM} + Z_1)} I_{sCM} \tag{18}$$

$$V_2 = \frac{R_{load} Z_{sCM}}{R_{load} + Z_{sCM}} I_{sCM} \tag{19}$$

$$Z_1 = \frac{R_{load} Z_{CY}}{R_{load} + Z_{CY}} \tag{20}$$

$$Z_2 = (Z_{LCM} + Z_1)\left(\frac{Z_{sCM}}{Z_{sCM} + Z_1 + Z_{LCM}}\right) \tag{21}$$

$Z_{LCM}$ is the impedance of the CM inductor. $Z_{CY}$ is the impedance of the Y capacitor. $V_2$ is the voltage across the LISN equivalent resistor shown in Figure. 9 and is the noise voltage across the LISN resistor without the CM filter. The final step in the filter design is to select the capacitance. A capacitance of 136 nF was selected using two $C_Y$ capacitors of 68 nF each in parallel. Using this capacitance and (9), the calculated attenuation of the filter at 583 kHz was 41, which was greater than the required attenuation of 38.

Figure. 11 shows the experimental results of the CM spectrum after the CM filter designed using the maximum noise impedance. It is noted that the SPC with the CM filter passed the FCC 15 Class B requirements over the required frequency range and the attenuation achieved was greater than the 3-dB margin over most of the frequency range. If the capacitance was selected based on an ideal LC filter design, the calculated capacitance required would be 14.3 nF using (22). The closest available capacitance is 16.4 nF (2 x 8.2 nF in parallel).

$$f_C = \frac{F_0}{2\pi\sqrt{L_{CM} C_Y}} \tag{22}$$

Then, using (17), with the parameters as follows $C_Y$ = 16.4nF, $R_{loadCM}$ = 25Ω, $L_{CM}$ = 200μH, and $Z_{SCM}$ = 205Ω, the calculated attenuation is 5.9 at 583 kHz. This is less then the required attenuation, $A_{TreqCM}$ = 38. It is obvious that this filter does not meet the FCC Class B requirements at approximately 560 kHz. From the above analysis, it is clear that neglecting the noise impedance leads a designer to design a filter that will not meet the requirements. Neglecting the noise impedance, the selected Y capacitance is 16.4 nF, but when the noise impedance is taken into account, the selected capacitance of 136 nF is nearly ten times larger. The experiment results also verified this conclusion. From above analysis and experimental results it is again demonstrated that the noise impedance has a significant effect on the performance of the EMI filter.

## 4.2. DM Filter Design

The DM filter design method is similar to that of the CM filter. In order to design the DM filter need to know: 1) the attenuation required making the DM noise spectrum pass the EMI standards at the frequencies of interest and 2) the maximum possible value and minimum possible value of the noise impedance for the frequency range of interest. A л topology was selected to suppress the DM noise, because this topology provides better performance than other simple topologies such as the LC topology. The topology, along with the noise source, its impedance and $R_{loadDM}$, is displayed in Figure.12.

Since an X capacitor was used at the input of the SPC, the minimum DM noise impedance was used for the filter design. The attenuation of this topology can be expressed using (23), where $Z_{Cl}$ and $Z_{Cs}$ are given by (24) and (25), respectively. In (24) and (25), $Z_{CX}$ is the impedance of the X capacitor and in (23); $Z_{LDM}$ is the impedance of the differential mode inductance.

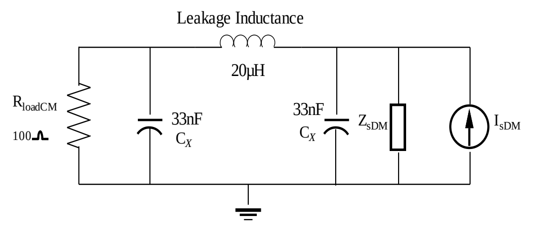


**Figure.12** DM noise equivalent circuit including the л DM filter topology

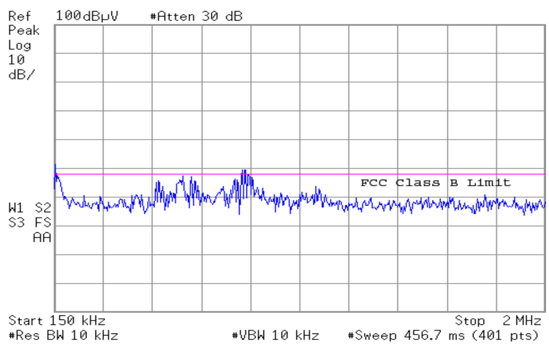


**Figure.13** Test results for DM noise spectrum after the PI filter was added

In (23) and (24), $R_{loadDM}$ is the LISN equivalent load resistance of 100Ω

$$|A_{TDM}| = \left|\frac{R_{loadDM} Z_{sDM}(Z_{C1} + Z_{LDM} + Z_{Cs})}{(R_{loadCM} + Z_{sDM}) Z_{Cs} Z_{C1}}\right| \tag{23}$$

$$Z_{C1} = \frac{R_{loadCM} Z_{CX}}{R_{loadCM} + Z_{CX}} \tag{24}$$

$$Z_{Cs} = \frac{Z_{CX} Z_{sDM}}{Z_{CX} + Z_{sDM}} \tag{25}$$

Using the experimental results the lowest corner frequency required was at $F_0$ = 583 kHz. At this frequency, the attenuation needed was 4.5. It should be noted that the leakage inductance of the CM inductor was used as the DM inductance, so only one inductor was required to meet the attenuation requirements for the CM noise and the DM noise. Using (23), along with $L_{DM}$ = 20µH, $Z_{sDMmin}$ = 0.06Ω, and $C_X$ = 33nF, the calculated attenuation $A_{TDM}$ was 7.87 at $F_0$ = 583 kHz, which was greater than the required attenuation of 4.5. The test results of the DM noise spectrum after the DM filter was added are shown in Figure. 13. It is clear that the DM filter exceeded the FCC 15 Class B specifications over the entire frequency range by more than 3 dB.

## 4.3. Complete EMI Filter Design

Finally, the CM and DM filters were assembled together. The final EMI filter is illustrated in Figure. 14. A 200-µF electrolytic capacitor was added to the output of the filter to ensure that the EMI filter would not destabilize the feedback loop of the SPC [18]. Since the resonant frequency of this electrolytic capacitor is relatively low, it does not affect the filter design.

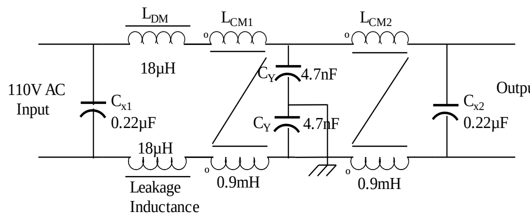


**Figure.14** Complete EMI filter containing the DM and CM filters for the ac–dc SPC

The test results for the filter are shown in Figure.15. From the test results, it can be noted that the filter allows the SPC to pass the FCC 15 Class B requirements for conducted EMI. The proposed method was used to design an EMI filter for an ac–dc power converter.

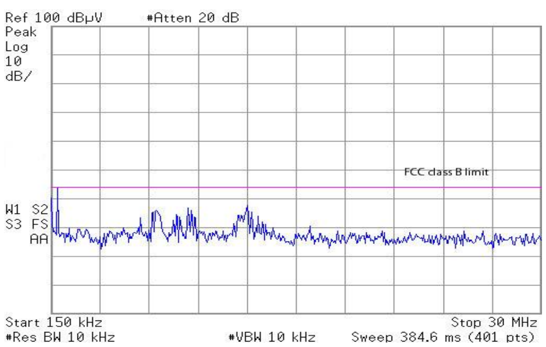


**Figure.15** Test results of the total noise spectrum after the EMI filter was added to the input of the ac–dc SPC

Experimental results are presented for a commercially available, ac–dc power module with a nominal input voltage of 120 $V_{RMS}$, and an output voltage of 5.1 $V_{DC}$ at a full load current of 22A. After measuring the DM and CM noise spectrum and calculating the noise impedance, the filter topology was selected [19]. A T topology was selected for CM filter because the worst case CM noise impedance was calculated as 1.9 kΩ. The attenuation required was 57 at 0.6 MHz.

A л topology was selected to suppress the DM noise, since this topology provides higher attenuation than other simple topologies such as the LC topology. The worst case DM noise impedance was 3Ω. The required attenuation was 20 at 0.15 MHz. It is clear that the filter exceeded the FCC 15 Class B specifications across the entire frequency range by more than 3 dB [20].

**Table.1** Comparison of hardware waveform results

| Frequency in MHz | Without Filter (dBµV) | | | With Filter (dBµV) | | |
|---|---|---|---|---|---|---|
| | Total | CM | DM | CM | DM | EMI |
| 10 | 79 | 65 | 67 | 40 | 42 | 29 |
| 20 | 78 | 62 | 62 | 34 | 38 | 27 |
| 30 | 76 | 58 | 55 | 28 | 35 | 26 |

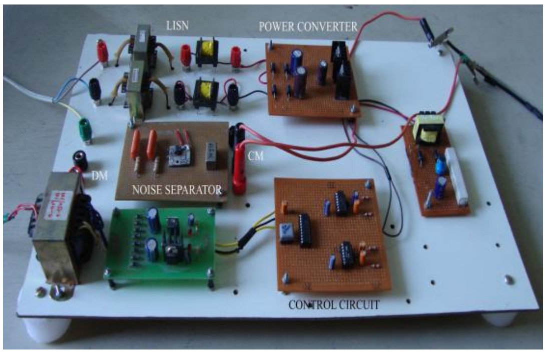


**Figure.16** Photograph of the measurement arrangement

The method presented is not only limited to single power converter, but it can be used to design EMI filters for paralleled switching power converter modules. The paralleled power modules can be simplified to one equivalent power module, so the method presented can be used to design the EMI filter for a power converter consisting of multiple power converters. Table.1 shows the comparison of hardware results for various frequency ranges using filters. Figure.16 shows the Photograph of the measurement arrangement. An example of an EMI filter for a Half-bridge isolated ac–dc converter was designed to illustrate the proposed method. As expected, the Half-bridge isolated ac–dc converter with the designed filter meets the requirements set in the standard.

## 5. Conclusions

The proposed method is based on measuring the noise spectrum and using the data to calculate the maximum possible value and minimum possible value of the noise impedance, and using it in the EMI filter design. This approach considers filter topology only but does not mention the relationship of noise source impedance and noise load impedance, which determines the EMI filter performance significantly. The various filters like LC filter, π filter and Complete EMI filters are analyzed. From the Table, it is well understood that the waveforms with filters have below the FCC class B level. The analysis and results proposed here can provide a guideline for future effectiveness of filtering schemes in switching power converters. The estimation of parameters for the model can be done by measurement or analytical calculation based on the geometry of the circuit. Based on the model and analysis proposed in this paper, future works will focus on parameter extraction and accurately EMI prediction.

Experimental Results have shown that the proposed method is very effective and accurate in identifying and capturing EMI features in switching power converters. The method presented is not only limited to half-bridge converters, but it can be applied to many different converter topologies, such as buck, flyback, boost, with single-phase diode bridge front-end. This model contains the salient features of conducted EMI, is convenient to use, and gives adequate prediction of EMI behavior in switching converters.

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