RIGA TECHNICAL UNIVERSITY

Deniss STEPINS

STUDY AND QUALITY IMPROVEMENT OF FREQUENCY MODULATED SWITCH-MODE POWER CONVERTERS

Summary of doctoral thesis

Riga 2011

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Faculty of Electronics and Telecommunications
Institute of Radioelectronics

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A DOCTORAL THESIS SUBMITTED TO RIGA TECHNICAL UNIVERSITY IN FULFILLMENT OF THE REQUIREMENTS FOR THE DEGREE OF DOCTOR OF SCIENCE IN ENGINEERING

The doctoral thesis for the degree of doctor of science in engineering will be presented in the auditorium 210 at Riga Technical University, Faculty of Electronics and Telecommunications, 12 Azenes street, on the 9^{th} of June 2011, at 16:30 p.m.

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CONFIRMATION

I confirm that the doctoral thesis, submitted for the degree of doctor of science in engineering at the Riga Technical University, is my own original work. The doctoral thesis has not been submitted for a degree at this or any other university.

Deniss Stepins …………………………….(Signature)

Date: ………………………

The doctoral thesis is written in Latvian, contains introduction, three chapters, conclusions, references, 9 appendices, 108 figures, 144 pages in total. A list of references consists of 122 sources.

Contents

TOPICALITY OF THE THESIS

Nowadays switching power converters (SPC) are essential to electric power conversion with high efficiency (linear regulators used widely in the past had very low efficiency and therefore they are used rather rarely nowadays). SPC are necessary for power supplies of electronic devices and systems, lighting equipment electronic ballasts, etc. [1,2,3]. A switching power converter is essential part of power electronics that nowadays is very important field of science and technology. Electrical power conversion principles, power electronic components as well as converter topologies are continuously improving [1]. Modern SPC development trends are mainly increasing specific power and efficiency, as well as reducing electromagnetic interference (EMI), output voltage ripples, improving thermal conditions, etc.

Advantages of switching power converters in comparison with other power converters (e.g. linear regulators) are mainly high efficiency (it can be higher than 90% for modern SPC) and high specific power (recently it has exceeded 3kW/dm^3) [1].

Despite their advantages, SPC have also their drawbacks. The main SPC disadvantages are high EMI both conducted and radiated as well as high output voltage ripples. EMI in SPC is caused mainly by rapidly switching power semiconductor devices with high current (*di/dt*) and voltage (*dv/dt*) rate of change [4]. With the introduction of the international electromagnetic compatibility directives (e.g. CISPR 22), there is an increasing awareness of the EMI problems of SPC [5]. The classical ways of mitigation of these problems usually include the application of appropriate design of SPC (e.g. the use of input and output filters, correct design of printed circuit boards, grounding, shielding, etc) and the use of soft switching techniques [6-10]. However the techniques can substantially increase SPC size

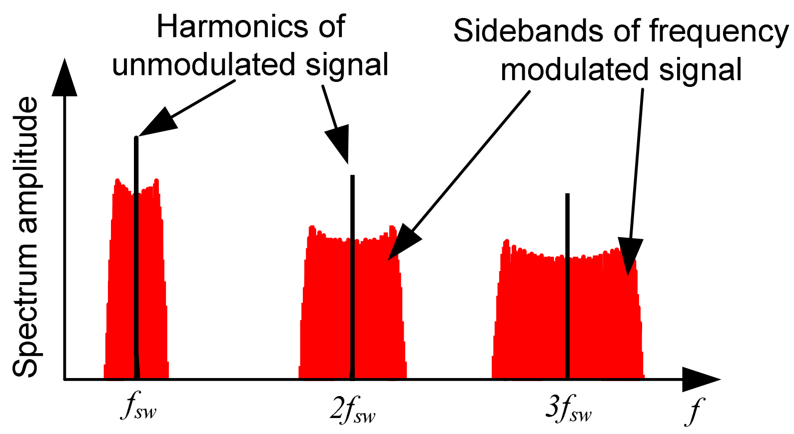


Fig. 1.1 Spectra of unmodulated and FM switching waveform.

and cost.

Another successful approach for EMI reduction known as spread spectrum has been developed over the last two decades. It is based on modulating the switching frequency in random, chaotic or periodic manner [5,11,12]. As a result, energy of discrete harmonics of unmodulated switching frequency is spread over a wider frequency range, thus peak EMI levels are significantly reduced [6] (as it can be seen in Fig. 1.1). The first publications on the use of the spread spectrum to reduce EMI in electronics (e.g. [13-15]) appeared in mid 1990s [6]. Over the last 15 years spread spectrum has been extensively used not only for EMI reduction in microprocessor systems and traditional SPC [5,6,11,13,14,16,19,20,27] but also in power factor correctors [21,22], lighting equipment electronic ballasts and inverters [23,24], and even in D class audio amplifiers [25]. It should be noted that despite the fact that random or chaotic frequency modulation techniques are sometimes applied to SPC because of better conducted EMI attenuation, periodic FM has several advantages: firstly, spectrum of signal, which frequency is modulated by periodic signal (sinusoidal, triangular, etc), is discrete, while that for random FM is continuous [6]; secondly, periodic FM can be implemented easily and, thirdly, periodic FM introduces appreciably lower low frequency ripples in output voltage of SPC [11]. In the research SPC with periodic FM will be investigated, however some results can also be used for the other FM techniques.

Despite the fact that FM can reduce conducted EMI, it could be sufficiently ineffective for EMI reduction with specific modulating signal $m(t)$ parameters. It is shown in [23] that amplitude modulation (AM) of electronic ballast output current can result in distortion and asymmetry of the sideband of switching frequency (f_{sw}) fundamental harmonic and consequently worsen effectiveness of the use of FM for EMI reduction. In our research the similar effect has also been observed for power factor correctors [21] and boost SPC [20]. To solve the problem in electronic ballasts, double-slope sawtooth modulating waveform was proposed in [23]. Using the waveform the sideband asymmetry can be effectively neutralized and conducted EMI attenuation can be improved at least by several dB [23]. However it is proved in our research that the technique is only effective for one combination of modulating signal and power circuit components parameters. This means that this technique is not optimal for other values of the parameters. That is why we, firstly, examine if this method can be applied for input EMI reduction in other SPC topologies, and secondly, we improve effectiveness of the use of the technique by optimizing it.

Despite its advantages FM similarly to other EMI suppression techniques has also its drawbacks:

- increase in peak-to-peak output voltage ripples [6,11,12,16,27-30];

- in addition to “natural” high-frequency (HF) output voltage ripples, low-frequency (LF) ripples (and LF harmonics) arise [6,11,27-30];
- deterioration of input power quality that is mainly important for AC mains-operated SPC [6,18,22,24,31];
- possible decrease in efficiency [6,18,22,24,32].

Since the disadvantages are very important in SPC, then negative effect of FM on the SPC performance should be examined and reduced. After analyzing scientific publications related to the similar problems it is revealed that the publications have many drawbacks such as:

- research results are mainly based on experiments (without appropriate theoretical analysis);
- the main causes of the problems are not revealed;
- for the most part, the publications consider only the use of FM in unregulated SPC;
- recommendations to solve the problems effectively are not proposed in the publications.

That is why we will try to examine the problems deeply in the research.

OBJECTIVES OF THE THESIS

The main aim of the doctoral thesis is to improve effectiveness of the use of FM in SPC, to examine and reduce the negative effects of the technique on input and output power quality as well as efficiency of SPC.

In order to achieve the aim, main tasks of the research are:

- to examine the influence of modulating signal and power stage circuit parameters on EMI attenuation using periodic FM;
- to improve input EMI attenuation in FM SPC using optimized double-slope modulating waveform;
- to investigate effect of FM on SPC output voltage ripples and to reduce them;
- to investigate negative effect of FM on the efficiency of SPC and to propose some recommendations to increase it;
- to examine negative effect of FM on SPC input power quality and to improve it.

SCIENTIFIC NOVELTY AND RESEARCH RESULTS

Novelty of the doctoral thesis and the main research results are related to improvement of usefulness of FM technique for EMI suppression in SPC and to reduction of negative effects of FM on important FM SPC characteristics. The results obtained and novelties mainly are as follows:

- effectiveness of the use of FM for maximum EMI suppression in FM SPC is appreciably improved by using optimized double-slope modulating waveform;
- a procedure for the calculation of double-slope modulating waveform optimum parameters values is developed to get maximum conducted EMI attenuation in FM SPC;
- formerly unknown causes of the increase in output voltage ripples and deterioration of input power quality of regulated FM SPC in both continuous conduction mode (CCM) and discontinuous conduction mode (DCM) are revealed;
- original recommendations for reducing output voltage ripples and improving input power quality of FM SPC are proposed;
- original recommendations for improving the efficiency of FM SPC are proposed and verified;
- a procedure for the calculation of the output voltage ripples and the efficiency in FM SPC is developed. The procedures could be useful for FM SPC design and optimization;
- original research on the negative effect of FM on SPC output voltage ripples and input power quality in DCM is conducted;
- original solution to get small difference between the switching delays and therefore very small LF ripples in CCM is proposed and verified.

DEFENDABLE THESES

For a defense of the doctoral thesis the following theses are proposed:

1) conducted EMI attenuation in FM SPC can be increased by using optimized dual-slope modulating waveform or selecting the switching frequency according to recommendations proposed. In order to get maximum EMI attenuation, optimum values of double-slope modulating waveform parameters should be calculated according to the procedure proposed;

2) output voltage ripples of FM SPC can effectively reduced using recommendations for selecting modulating signal, power stage and control unit parameters;
3) recommendations proposed can help to mitigate negative effect of FM on the efficiency of FM SPC;
4) input power quality of FM SPC can be significantly improved using the recommendations proposed.

RESEARCH METHODOLOGY

In the doctoral thesis, FM SPC have been studied theoretically, experimentally and using computer simulations. Initially the problems were analyzed theoretically and analytical expressions were also derived. Theoretical results have been verified using the computer simulation. Further the results obtained have been verified experimentally.

PRACTICAL SIGNIFICANCE

The results obtained and the recomendations proposed could be applied for designing high-quality FM SPC as well as for their development.

APPROBATION AND PUBLICITY

The research results have been presented in 9 international scientific conferences:

- IEEE International Symposium on Industrial Electronics (IEEE ISIE10), Bari, Italy, July 4-7, 2010.
- International Biennial Baltic Electronics Conference (BEC2008), TUT, Tallinn, Estonia, Oct. 6-8, 2008.
- International Biennial Baltic Electronics Conference (BEC2010), TUT, Tallinn, Estonia, Oct. 4-6, 2010.
- The 14th International Conference „Electronics '10", Kaunas, Lithuania, May 18-20, 2010.
- The 12th International Conference „Electronics '08", Kaunas, Lithuania, May 20-22, 2008.
- The 13th International Conference „Electronics '09", Kaunas, Lithuania, May 12-14, 2009.

- The 10th International Conference „Electronics '06", Kaunas, Lithuania, May 23-25, 2006.
- The 9th International Symposium on Electronics and Telecommunications (ISETC2010), Timisoara, Romania, Nov. 11-12, 2010.
- The 50th RTU International Conference, Rīga, Latvija, 2009. gada 14.-16. Oktobrī.

The research results have been published in 12 internationally recognized sources:

1. Stepins D. Analysis of Output Voltage of Switching Frequency Modulated DC-DC Converter Operating in Discontinuous Conduction Mode // *Proceedings of IEEE International Symposium on Industrial Electronics (IEEE ISIE10)*, Bari, Italy, July 4-7, 2010. – pp. 670-675.
2. Stepins D. Examination of influence of periodic switching frequency modulation in dc/dc converters on power quality on a load // *Proceedings of the 11th Biennial Baltic Electronics Conference*, Tallinn, Estonia, Oct. 6-8, 2008. – pp. 285-288.
3. Stepins D. Examination of Power Converters with Modulated Switching Frequency // Electronics and Electrical Engineering. - 9(105). – 2010. - pp. 33.-38.
4. Jankovskis J., Stepins D., Pikulins D. Improving effectiveness of the use of frequency modulation in power converters // *Proceedings of the 12th Biennial Baltic Electronics Conference*, Tallinn, Estonia, Oct. 4-6, 2010. – pp. 327-330.
5. Stepins D. Improving EMI attenuation in a frequency modulated boost converter // RTU zinātniskie raksti. 7. sēr., Telekomunikācijas un elektronika. - 9. sēj. (2009), 17.-23. lpp.
6. Jankovskis J., Stepins D., Tjukovs S., Pikulins D. Examination of Different Spread Spectrum Techniques for EMI Suppression in dc/dc Converters // ELECTRONICS AND ELECTRICAL ENGINEERING. - 6 (86). – 2008. – pp. 60.-64.
7. Jankovskis J., Stepins D., Pikulins D. Lowering of EMI Noise in Boost Type PFC by the use of Spread Spectrum// ELECTRONICS AND ELECTRICAL ENGINEERING. - 6 (94). – 2009. – pp. 15.-18.
8. Jankovskis J., Stepins D., Pikuļins D. Efficiency of PFC Operating in Spread Spectrum Mode for EMI Reduction // Electronics and Electrical Engineering. - 7. – 2010. – pp. 13.-16.
9. Jankovskis J., Stepins D., Pikuļins D. Effects of Increasing Switching Frequency in Frequency Modulated Power Converters // *Proceedings of the "2010 9th International Symposium on ELECTRONICS AND TELECOMMUNICATIONS"*,

Timisoara, Romania, November, 11-12, 2010. – pp. 115-118.

10. Jankovskis J., Stepins D. An Examination of Parasitic Processes in DC/DC Power Converters // ELECTRONICS AND ELECTRICAL ENGINEERING. - 4(68). - 2006. – pp. 15.-18.
11. Stepins D. On the Figures of Merit for Planar Spiral Inductors // RTU zinātniskie raksti. 7. sēr., Telekomunikācijas un elektronika. - 7. sēj. (2007), 16.-18. lpp.
12. Stepins D. Performance-Analysis-Based Examination of Planar Spiral Inductors // Materials of the 4th International Young Scientist Conference "Modern Issues in Radio Engineering and Telecommunications", Sevastopol, Ukraine, April 21-25, 2008. – pp. 312.

Four publications (1,2,4,9) are also available in IEEEXplore Digital Library.

STRUCTURE OF THE THESIS

The doctoral thesis consists of introduction, three chapters, conclusions, references and appendices. Topicality, scientific novelty and objectives of the thesis as well as literature review, approbation and practical importance are provided in the introduction. The second chapter gives a brief review of EMI modes, international EMI specifications, traditional conducted EMI suppression techniques for SPC and their advantages and disadvantages, as well as spread spectrum technique. The third chapter presents a brief review of frequency modulation, considers spectrum of a periodic rectangular pulse train as well as the attenuation. Theoretical study of EMI in FM SPC is performed in the chapter 3.2. The use of double-slope sawtooth modulating waveform for increasing EMI suppression is analyzed in the chapter 3.3. Chapter 3.4 provides results of the experimental verification, but chapter 3.5 draws some conclusions. The fourth chapter is dedicated to the negative effect of FM on crucial FM SPC quality characteristics: peak-to-peak output voltage ripples (chapter 4.1), the efficiency (chapter 4.2) and input power quality (chapter 4.3). Appendices show SIMULINK models for both unregulated and regulated FM SPC with buck and boost topologies, present regulated FM buck converter schematic diagram used in the experiments, and present the calculation procedures for peak-to-peak output voltage ripples and efficiency of FM SPC.

2. ELECTROMAGNETIC INTERFERENCE OF SWITC-MODE POWER CONVERTERS AND ITS SUPPRESSION TECHNIQUES

In the second chapter EMI modes, its causes in SPC and international EMI specifications are discussed briefly. A brief review of traditional conducted EMI suppression techniques for SPC and their advantages and disadvantages, as well as novel technique, known as spread spectrum, is presented in this chapter.

SPC are usually one of the main contributors to EMI. Since EMI generated by SPC can disturb normal operation of other electronic equipment, EMI suppression is of importance when designing SPC. EMI in SPC is caused by rapidly switching power semiconductor devices with high *di/dt*, *dV/dt* [8]. As a result switching waveform spectrum consists of multiple switching frequency (f_{sw}) harmonics that occupy wide bandwidth. Since f_{sw} of SPC are rather high (they can range from several tens of kHz up to several MHz), the problem is very important [7,10,15]. Conducted EMI is more pronounced at lower frequencies, but radiated EMI: at higher frequencies. That is why conducted EMI is usually measured in the frequency range from 150 kHz to 30 MHz, but radiated EMI – from 30 MHz to 1 GHz [4]. There are various standards that specify EMI measurement limits. One of the world-wide accepted standards is CISPR Publication 22 [4].

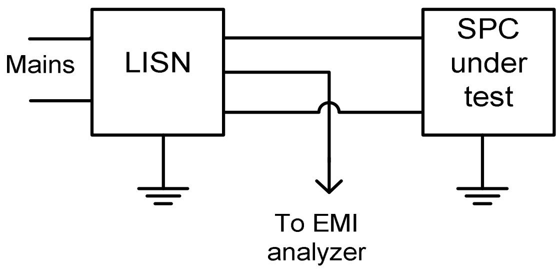


Fig. 2.1 Typical EMI measurements block diagram.

Conducted EMI as such can travel through input wires to a power line and consequently disturb normal operation of other electronic equipment connected to the same line. For conducted EMI measurements a line impedance stabilization network (LISN) and EMI or spectrum analyzer are used. LISN is usually connected between a power line and SPC under test (see Fig. 2.1). The standards also specify EMI analyzer input filter bandwidth, also known as resolution bandwidth (RBW), for different EMI measurement ranges. For example, according to CISPR 22 conducted EMI measurements should be performed in the frequency range from 150 kHz to 30 MHz with RBW=9kHz, but as per CISPR 16 specifications the measurements should be performed in the frequency range from 9 kHz to 150 kHz with RBW=200 Hz [10].

Conducted EMI can be categorized into two separate modes of current flow: differential mode and common mode noise [4]. Differential mode EMI is due to SPC input current, but common mode EMI is mainly caused by parasitic capacitances to ground and rapidly switching power transistor with high *dv/dt*.

Classical ways for suppressing conducted and radiated EMI are usually include the use of input EMI filters, that can effectively reduce both differential and common mode EMI; special snubber RC or RCD circuits, that can be connected across power transistors or diodes; ferrite beads, that attach easily to SPC input/output wires and power components terminals; shielding of SPC for suppressing radiated emissions; correct design of printed circuit board according to electromagnetic compatibility requirements; soft switching techniques [8,9].

Each EMI suppression technique has its own drawbacks. For instance, despite the fact that passive input filters can effectively reduce conducted EMI, they are highly responsible for increasing size, weight, cost of SPC and decreasing their efficiency.

Considerable recent attention has been focused on the use of another successful approach for EMI reduction, known as spread spectrum. The technique, in fact, has been adopted from communication theory and, as it was mentioned before, it has been used for EMI suppression in microprocessor systems, as well as various SPC and recently even in D class amplifiers [6,8,25]. As a result, energy of discrete harmonics of unmodulated switching frequency is spread over a wider frequency range, thus peak EMI levels are significantly reduced [6], as it can be seen in Fig. 1.1. Spread spectrum can help to reduce not only conducted EMI but also radiated one. Spreading the spectrum can be achieved through different ways: one of the most effective techniques is frequency modulation (FM) [5,11]. In this work, frequency modulated SPC are to be examined, because FM can give better EMI attenuation than other spread spectrum techniques. In general the switching frequency can be modulated periodically, chaotically or in random manner. It is mentioned in several publications (e.g. [11]), that random and chaotic FM can give better EMI attenuation, than periodic FM (under the assumption that frequency deviation is the same). However periodic FM has several advantages. For example, it can give lower output voltage ripples (mainly LF ripples) and can be implemented easier, than chaotic or random FM. This PhD thesis is dedicated to SPC with periodic FM, nevertheless several results can be applied for other FM techniques.

3. IMPROVING EFECTIVENESS OF THE USE OF FM IN SPC

The third chapter presents a brief review of frequency modulated signals and their main parameters, considers spectrum of a periodic rectangular pulse train as well as the attenuation that can be achieved using FM in SPC. Original theoretical research on EMI in FM boost SPC has been done taking into account LISN and power stage parasitic parameters. In the third chapter it also proved that optimized double-slope modulating signal can substantially improve usefulness of FM and increase input EMI attenuation in traditional SPC. The procedure for the calculation of double-slope modulating waveform optimum parameters values to get maximum conducted EMI attenuation in FM SPC is also developed in the chapter.

FM rectangular pulse train representing approximately, for example, MOSFET control signal or several other power stage voltages responsible for EMI can be expressed as:

$$s(t) = sign[\cos(2\pi f_{sw} t + \theta(t))], \tag{3.1}$$

where f_{sw} is central switching frequency; $\theta(t)$ – time-dependent phase angle.

Spectrum of rectangular pulse train which frequency is modulated by a sine wave can be calculated using the Bessel functions. For other periodic modulating waveforms *m(t)*, Fourier series in complex exponent form should be used. It should be noted that if f_{sw}/f_m (f_m is modulating frequency) is integer number, then FM signal period equals *m(t)* period T_m. FM rectangular pulse train spectrum is made up of sidebands comprising side frequencies $nf_{sw} \pm kf_m$ (*n*, *k* are integers) [33], as it is shown in Fig. 3.1. Attenuation due to FM is different for each individual unmodulated f_{sw} harmonic (as it can be seen in Fig. 3.1.). Since unmodulated fundamental harmonic of f_{sw} usually is dominant in conducted EMI spectrum, then it makes sense to estimate the attenuation just for this harmonic. The attenuation can be

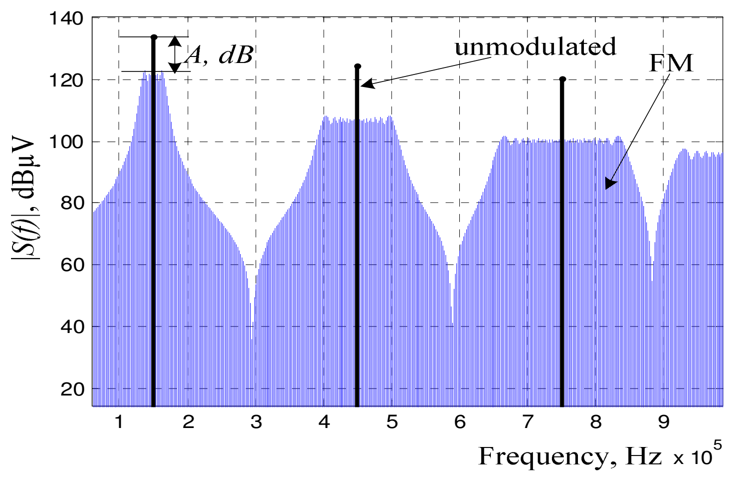


Fig. 3.1 Spectrum of unmodulated and FM rectangular pulse train (f_{sw}=150kHz; f_m=2kHz; frequency deviation Δf_{sw}=20kHz; *m(t)* is sawtooth waveform).

defined as the difference expressed in dB between amplitude of unmodulated fundamental harmonic of f_{sw} $|C_1|=2|d_1|$ and maximum amplitude of the first sideband harmonics of FM signal [24]:

$$A = 20\log_{10}\left(\frac{|C_1|}{\max(|S_{1\,\mathrm{mod}}(f)|)}\right), \tag{3.2}$$

where $|S_{1mod}(f)|$ is the amplitude spectrum of the first sideband.

The higher modulation index $\beta=\Delta f_{sw}/f_m$ is, the better f_{sw} harmonics amplitude attenuation is [6,11]. Notice that the attenuation is independent on f_{sw}. The attenuation that can achieved by frequently used modulating waveforms (i.e. sine, sawtooth, triangle) is almost the same up to $\beta=10$. For higher β values, sawtooth $m(t)$ gives the best attenuation, but sine wave – the worst [6]. Increase in the attenuation for sufficiently high β values is slow. FM signal spectrum of the first sideband is symmetrical to f_{sw}.

Since rectangular pulse trains are only analyzed in many publications (e.g. [6,11,15]) about the use of FM for EMI suppression, it hinder us to reveal the causes worsening the effectiveness of FM in SPC. That is why in the research EMI is analyzed taking in account LISN and FM SPC power stage parasitic parameters. For this purpose total EMI model [35] shown in Fig. 3.3 is used. The model can give us a possibility to derive transfer function between equivalent EMI source V_s and LISN radio frequency (RF) output voltage V_{LISN}:

$$\underline{K}_{EMI}(f) = \frac{50}{50+\underline{Z}_{C5}}\left(\frac{(25+\underline{Z}_{C5}/2)(5+\underline{Z}_{L3})/(55+\underline{Z}_{C5}+\underline{Z}_{L3})}{\underline{Z}_{Ch}+(25+\underline{Z}_{C5}/2)(5+\underline{Z}_{L3})/(55+\underline{Z}_{C5}+\underline{Z}_{L3})} - \frac{\underline{Z}_{Cin}}{2(\underline{Z}_{Cin}+\underline{Z}_{L})}\right), \tag{3.3}$$

where $\underline{Z}_{C5}$, $\underline{Z}_{Ch}$, $\underline{Z}_{L3}$ are complex impedances.

In making analysis of V_{LISN} spectrum for different f_{sw} values it is concluded that the first harmonic amplitude attenuation A_{EMT} depends not only on β (as it is for FM rectangular

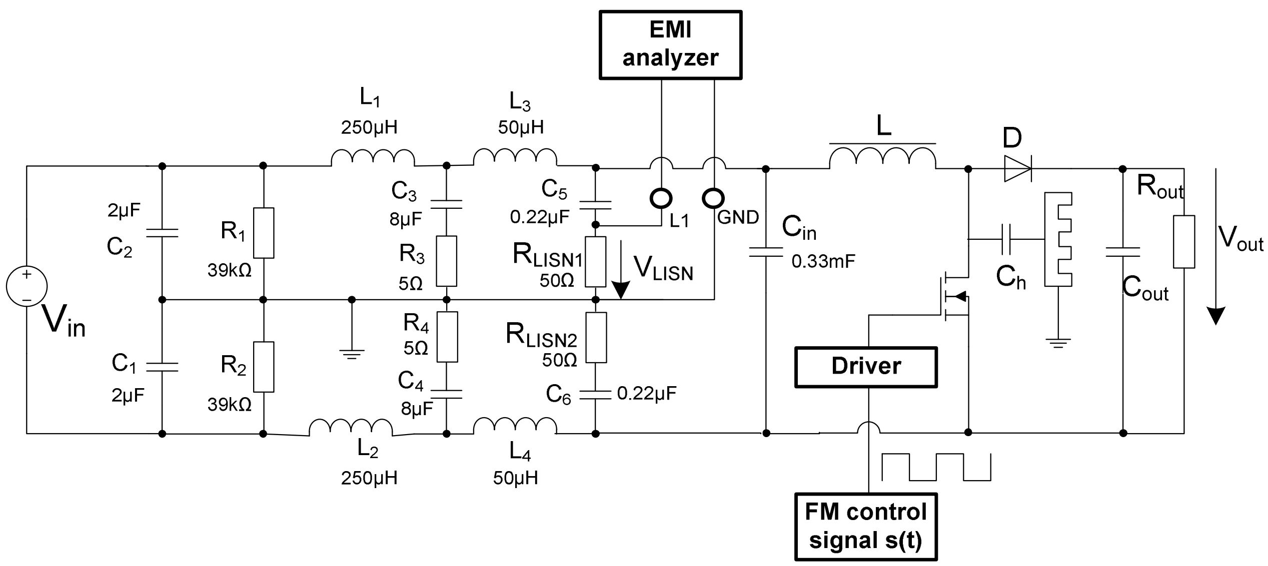


Fig. 3.2. Boost SPC schematic diagram with LISN.

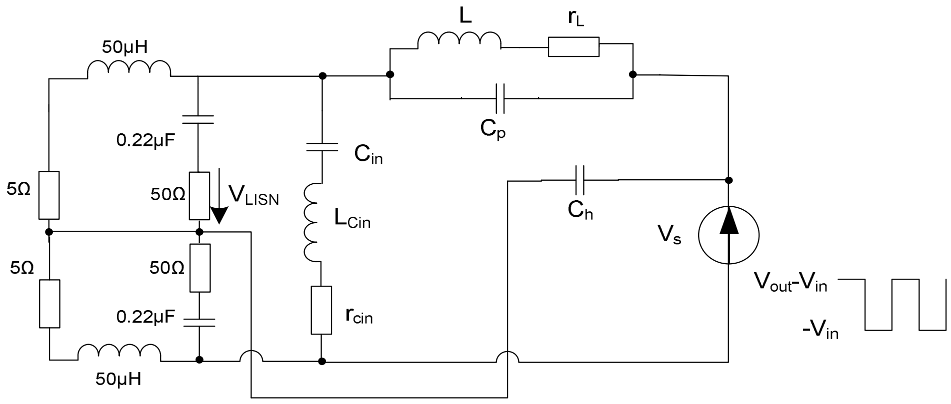


Fig. 3.3. Total boost converter EMI model [35].

pulse train) but also on f_{sw}. The frequency determines the sideband location in the frequency domain as it can be seen in Figs. 3.4 and 3.5. If the sideband is in the range where $|\underline{K}_{EMI}(f)|$changes steeply (this mainly corresponds to the LF range where differential-mode EMI is dominant), then difference between A and A_{EMT} (ΔA) is high and in turn the effectiveness of the use of FM is low. If the sideband is in the range where $|\underline{K}_{EMI}(f)|$ changes slowly (this mainly corresponds to higher frequencies), then ΔA is small. The best results

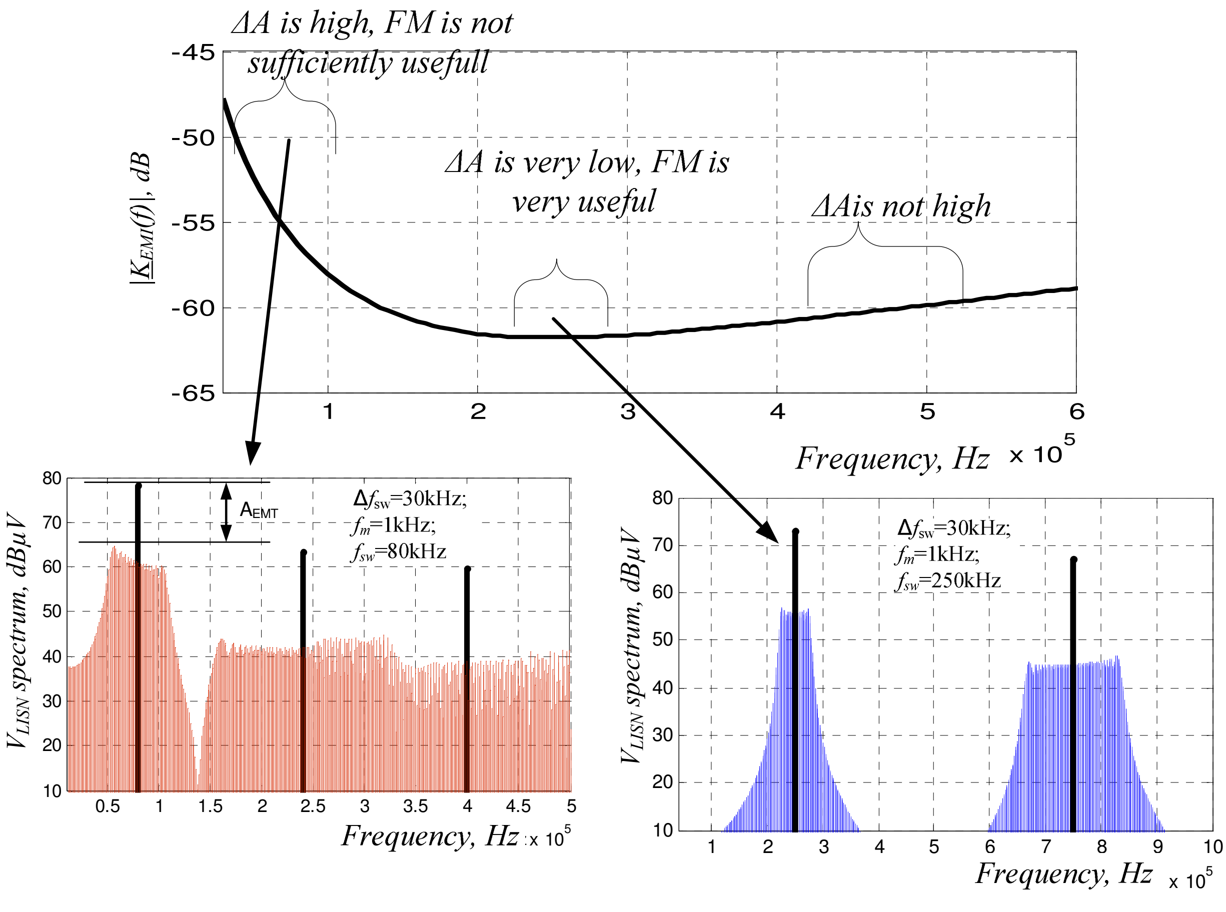


Fig. 3.4 Transfer function $|\underline{K}_{EMI}(f)|$. Effectiveness of the use of FM depends on the location of the fundamental harmonic sideband.

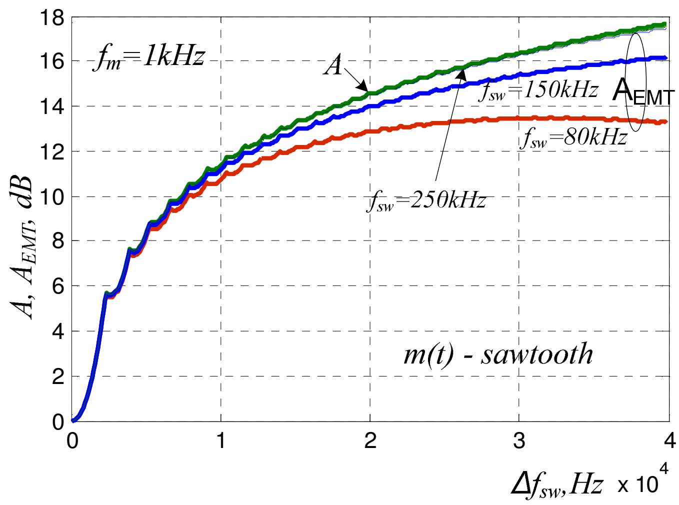


Fig. 3.5 EMI attenuation of fundamental f_{sw} harmonic amplitude both for FM rectangular pulse train (A) and FM SPC V_{LISN} (A_{EMT}).

can be achieved when the sideband is located in the range where common-mode and differential-mode EMI are comparable. In the range $|\underline{K}_{EMI}(f)|$is almost independent on the frequency. When increasing Δf_{sw}, the difference ΔA between A_{EMT} and A also increases. However the difference is very small for low values of Δf_{sw}. For a given modulation index sinusoidal FM gives the highest ΔA but triangular FM the lowest ΔA. The difference ΔA decreases slightly for higher f_m.

It should be noted that sometimes selection of f_{sw} depends on other SPC parameters, mainly the efficiency. That is why it is of importance to propose an alternative solution to get maximum A_{EMT} (or minimum ΔA) for given $m(t)$ parameters without changing f_{sw}. A technique to reduce AM and fundamental f_{sw} harmonic sideband asymmetry in FM electronic ballast output current was proposed in [23]. The technique is based on the use of double-slope sawtooth modulating waveform shown in Fig. 3.6. Since the technique was not applied to other SPC topologies for conducted input EMI suppression, then one of our tasks is to verify and prove that the method is also effective in FM SPC for input conducted EMI suppression and improving effectiveness of the use of FM. The asymmetry of the sidebands can be appreciably reduced using different slopes for the sawtooth modulating waveform [23] (see Fig. 3.6). The slopes can be adjusted by changing t_0 values. It is noted in [23] that output current fundamental f_{sw} harmonic sideband asymmetry and AM can

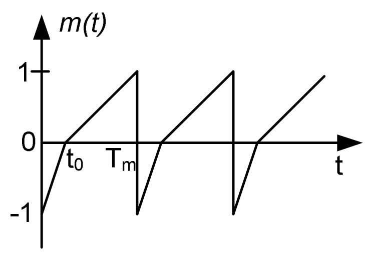


Fig. 3.6. Double-slope sawtooth modulating waveform.

be reduced if t_0=0.35T_m. However as it is proved in the doctoral thesis, the sideband asymmetry and in turn ΔA are dependent both on parameters of the power stage components including parasitics and modulating signal and switching signal parameters. Thus t_0=0.35T_m can give minimum ΔA or maximum A_{EMT} only for a single combination of the parameters. For other values of the parameters it can even worsen A_{EMT} in comparison with traditional sawtooth waveform with t_0=0.5T_m. That is why one more task of the research is to develop a procedure to calculate optimum t_0 to get minimum ΔA or maximum A_{EMT} for a given $m(t)$ and SPC parameters values, and also examine if the technique does not worsen EMI attenuation at higher frequencies. It should be noted that not only modified sawtooth but also other $m(t)$ can be used provided that the similar modification is done. Changing t_0 value with small step and calculating V_{LISN} spectrum using the procedure and the expressions for the spectrum calculation derived in the thesis, optimum t_0 values can be found. Thus we can conclude that the use of the optimized double-slope modulating waveform can appreciably improve A_{EMT} and in turn reduce ΔA, and improve FM usefulness in FM SPC, as it can be seen in Fig. 3.7 and 3.8 (note that A_{EMT1} denotes EMI attenuation using modified $m(t)$). It is worth noting that the technique does not worsen EMI attenuation at higher frequencies.

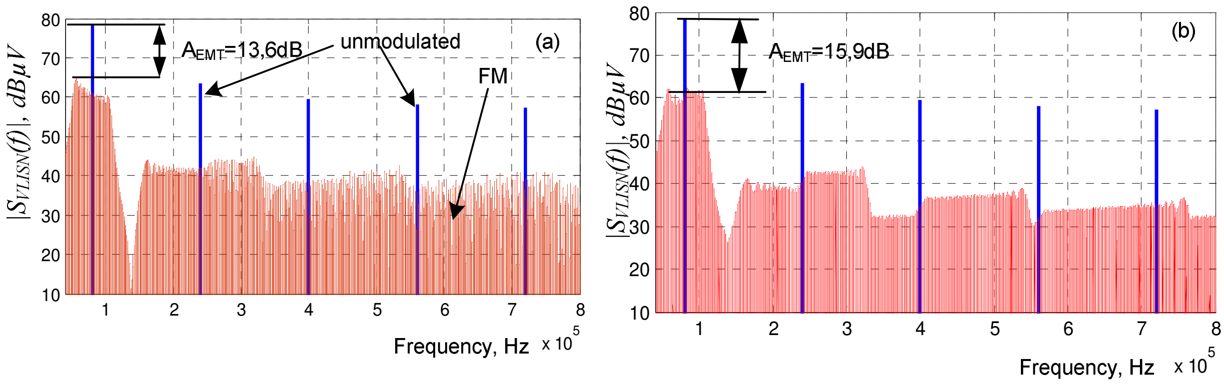


Fig. 3.7. Calculated V_{LISN} spectrum for unmodulated and FM SPC. Modulation parameters: Δf_{sw}=30kHz, f_m=1kHz, f_{sw}=80kHz, (a) t_0=0.5T_m; (b) t_0=0.29T_m.

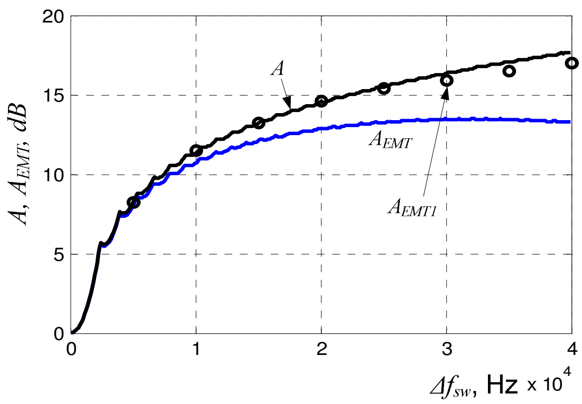


Fig. 3.8. Calculated A_{EMT} (for t_0=0.5T_m), A and A_{EMT1} (using optimum t_0) as a function of Δf_{sw} for f_m=1 kHz and f_{sw}=80kHz.

4. STUDY OF IMPACT OF FM ON SPC PARAMETERS; IMPROVING QUALITY OF SPC

4.1. Influence of FM on output voltage ripples

In the chapter 4.1 the output voltage ripples both for unregulated and regulated SPC in CCM and DCM are investigated. Expressions to calculate the peak-to-peak output voltage ripples in both CCM and DCM are derived. The results obtained are confirmed experimentally. Original recommendations to reduce the output voltage ripples effectively are proposed and verified. A circuit to mitigate the LF ripples in CCM is also developed.

It is measured in several publications (e.g. [6,11,16]) that FM SPC output voltage ripples increase due to FM. Thorough examination of FM effect on the output voltage ripples of unregulated (operating in open-loop) FM SPC in CCM was performed only in [27] and independently in our publications ([17,30] in CCM and [29] in DCM). Although the main causes of the increase in the output voltage ripples are revealed in [27] for unregulated buck FM SPC, effect of FM on the ripples of regulated (operating in closed-loop) SPC and FM SPC operating in DCM are not investigated in the publication. Moreover, expressions or procedure to calculate the output voltage ripples are not derived and recommendations to mitigate the ripples effectively are not proposed in [27].

In making thorough examination we have concluded that FM increases the output voltage ripples both in CCM and DCM. In general the ripples consist of HF switching ripples and

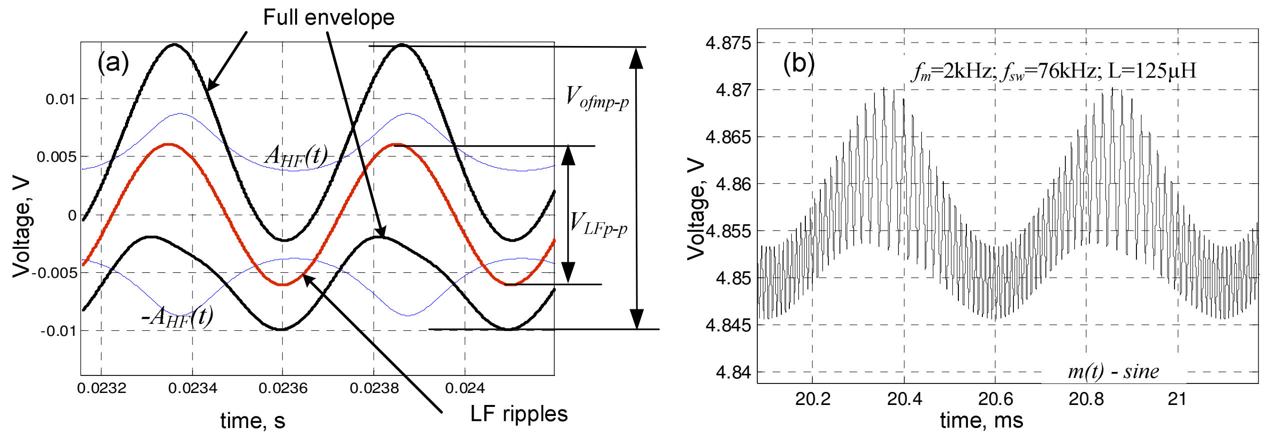


Fig. 4.1. Calculated envelope of the output voltage ripples (a) un simulated output voltage ripples (b) for buck SPC in CCM taking in account $|t_d|$=110ns. Parameters: C_{out}=330uF; L=125uH; V_{in}=10V; *m(t)*-sine; f_m=2kHz; Δf_{sw}=30kHz; f_{sw}=76kHz; r_{cout}=0.04Ω; r_L=0.12Ω; D=0.5; output load R_{out}=10Ω.

LF ripples with f_m, as it is shown in Fig. 4.1. Generally an envelope of the total ripples in FM SPC is as follows:

$$V_{ofmp-p} = \max\{\tilde{v}_{LF}(t) + A_{HF}(t)\} - \min\{\tilde{v}_{LF}(t) - A_{HF}(t)\}, \qquad (4.1)$$

where $\tilde{v}_{LF}(t)$ is LF ripples; $A_{HF}(t)$ is envelope of the HF ripples. An expression for the envelope of the HF ripples can be derived using an expression for the unmodulated SPC ripples. For this purpose constant f_{sw} should be substituted with instantaneous switching frequency $f_{sw}(t)$. So the HF ripples, for instance, for FM SPC with typical output electrolytic capacitor in CCM is:

$$A_{HF}(t) = \frac{r_{cout} V_{out} (1 - D)}{2L f_{sw}(t)} \quad , \tag{4.2}$$

where r_{cout} is equivalent series resistance of the output capacitor (ESR); D is average duty ratio; V_{out} is DC output voltage.
The instantaneous switching frequency is:

$$f_{sw}(t) = f_{sw} + \Delta f_{sw} m(t) . \tag{4.3}$$

The cause of increasing the HF ripples is natural: minimum switching frequency $f_{swmin}=f_{sw}-\Delta f_{sw}$ gives maximum peak-to-peak HF ripples [27,30]. Relative increase in peak-to-peak HF ripples in comparison to ripples of the unmodulated SPC is lower for higher f_{sw} but the same Δf_{sw}. Increase in the HF ripples of FM SPC with typical electrolytic capacitors is much less pronounced for boost SPC or related topologies (such as flyback or buck-boost) where output capacitor current also depends on power inductor or transformer secondary DC current I_{Lavg} than in buck SPC or related topologies where the HF ripples depend only on the power inductor AC current. This means that from the HF ripples point of view, FM is much less harmful in boost SPC or related topologies than in buck SPC.

The genuine cause of the LF ripples for FM SPC in CCM is nonzero difference $|t_d|$ between the switching delays [17,27,30]. SPC switching delays are mainly related to power transistor switches, their drivers and logic circuits (e.g. flip-flops) of pulse width modulated (PWM) integrated circuits. Sometimes the switching delays are intentionally applied for SPC control signals, for instance, in synchronous buck SPC where the delay is necessary for both power transistors not to be in on state simultaneously because it can cause high shoot-through current. If $|t_d|\neq 0$, then FM causes parasitic PWM in CCM and consequently instantaneous duty ratio d changes in time with f_m. This effect causes high LF ripples in the output voltage. In the doctoral thesis a general expression for the calculation of the LF ripples of regulated FM SPC in CCM is derived in operator form considering nonzero $|t_d|$ as follows:

$$\tilde{v}_{LF}(s) = \tilde{d}(s) H_{co}(s) / (1 + T(s)) = | t_d | \Delta f_{sw} m(s) H_{co}(s) / (1 + T(s)) , \tag{4.4}$$

where $H_{co}(s)$ is control-to-output transfer function; $T(s)$ is open loop gain. Using the Laplace transform of the equation (4.4), a steady-state LF ripples $\tilde{v}_{LF}(t)$ can be calculated in time domain. Note that only the steady-state solution is of interest and that $m(t)$ is a periodic function of time. The LF ripples in the time domain for sinusoidal FM can be easily obtained from (4.4) as follows:

$$\tilde{v}_{LF}(t) = |t_d|\Delta f_{sw} \cos\left[2\pi f_m t + \arg\left(\frac{H_{co}(j2\pi f_m)}{1+T(j2\pi f_m)}\right)\right]\left|\frac{H_{co}(j2\pi f_m)}{1+T(j2\pi f_m)}\right|. \qquad (4.5)$$

It follows that the LF peak-to-peak ripple for sinusoidal FM in CCM is:

$$V_{LFp-p} = 2|t_d|\Delta f_{sw}\left|\frac{H_{co}(j2\pi f_m)}{1+T(j2\pi f_m)}\right| \quad . \qquad (4.6)$$

From this results the conclusion that the LF peak-to-peak ripple $V_{LFp\text{-}p}$ is directly proportional to $|t_d|$ and Δf_{sw}, and it also depends on f_m and $m(t)$. If $|t_d|=0$, then output LF ripples are nullified even in regulated SPC. It follows that $V_{ofmp\text{-}p}$ depends only on the HF ripples.

Recommendations for decreasing the output voltage ripples in CCM.

1) To neutralize the LF ripples, the power MOSFET control circuit should be designed so that $|t_d|$ is low. This can be achieved, for instance, using two MOSFET drivers with delay RCD circuit connected between them, as it is shown in Fig. 4.2. It should be noted that the technique does not require to increase cost and printed circuit board size, because modern MOSFET drivers integrated circuits include two drivers with individual outputs.

2) If it is possible to choose FM SPC topology, then it is better to use boost or related topologies where output capacitor current depends on I_{Lavg} substantially.

3) In regulated FM SPC, f_m should be far away from f_{max} at which $|H_{co}(s)/(1+T(s))|$ is maximal. Note that f_{max} in buck SPC is equal to output filter resonance frequency, as it is shown in Fig. 4.3.

4) As the modulating signal triangular $m(t)$ should be used. It is recommended to use sawtooth $m(t)$ when $f_m > f_{max}$. The sawtooth $m(t)$ should not be used, if its first harmonic is in

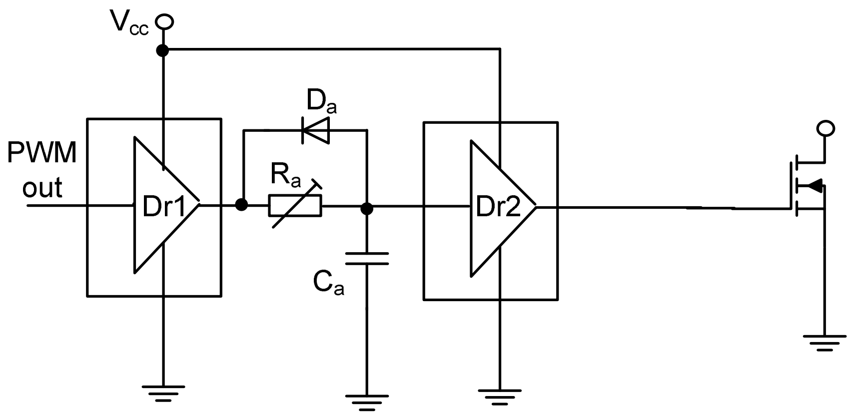


Fig. 4.2. An artificial delay circuit to get small $|t_d|$ and the LF ripples.

the range where |$H_{co}(s)/(1+T(s))$| rises.

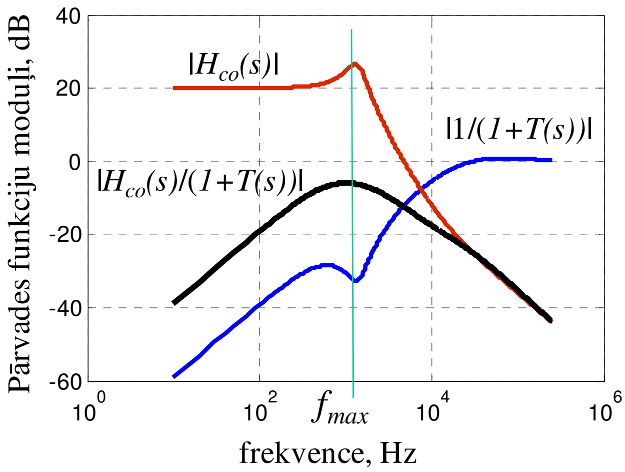


Fig. 4.3. Regulated buck FM SPC transfer functions in CCM. Parameters: C_{out}=330uF; L=125uH; Δf_{sw}=30kHz; r_c=0.04Ω; r_L=0.12Ω; R_{out}=10Ω; f_{sw}=80kHz; f_{cut}=5kHz.

FM causes the peak-to-peak output voltage ripple V_{ofmp-p} to increase in FM SPC in DCM. As it is for CCM, in DCM the ripples also consist of HF switching ripples and LF ripples with f_m. Moreover the output voltage spectrum consists of LF f_m harmonics and side frequencies. Similarly to CCM, in DCM f_{swmin} also determines increase in peak-to-peak HF ripples.

The cause of the LF ripples in DCM is not the same as in CCM: the LF ripples are due to the fact that output voltage averaged to T_{sw} depends on f_{sw}. This means that the LF ripples are natural and independent on parasitic parameters as it is in CCM. The peak-to-peak LF ripple for sinusoidal FM in DCM is:

$$V_{LFp-p} \approx \frac{\Delta f_{sw}}{f_{sw}^2} \left| \frac{H_{out}(j2\pi f_m)}{1+T(j2\pi f_m)} \right| \frac{D^2 V_{in}(V_{in}-V_{out})}{LV_{out}} \text{ ,} \tag{4.7}$$

where $H_{out}(j2\pi f)$ is output voltage to inductor current transfer ratio in DCM. It can be concluded that V_{LFp-p} is approximately proportional to Δf_{sw} and approximately inversely proportional to f_{sw} raised to the power of 2. This means that for constant Δf_{sw}, the higher f_{sw} is, the lower LF ripples are. In general, to derive the expressions of the output voltage LF ripples for different SPC topologies and $m(t)$, the expressions should be firstly derived for LF ripples $\tilde{v}_{LFunreg}(t)$ in FM SPC using averaged models in DCM. Then using $T(s)$ in DCM, the LF ripples can be calculated for the regulated FM SPC. Note that V_{LFp-p} also depends on f_m and $m(t)$. In the regulated FM SPC, the lower f_m is, the lower V_{LFp-p} is.

Recommendations for decreasing the output voltage ripples in DCM.

1) Keeping $|t_d|$ as low as possible is not necessary in DCM, because it barely affects the LF ripples in DCM. This can be deduced from the expression of power inductor current averaged to T_{sw} for FM buck SPC operating in DCM:

$$\langle i_L \rangle = \frac{V_{in}(V_{in} - V_{out})}{2LV_{out}} \cdot \left(\frac{D^2}{f_{sw}(t)} + 2Dt_d + t_d^2 f_{sw}(t) \right). \tag{4.8}$$

It follows that the first term in the second bracket is the same as for FM SPC with t_d=0. The second term is negligible DC term. But the third term is LF AC component due to nonzero t_d. Considering typical $|t_d|$ and FM parameters, it can be concluded that the third AC term is much lower than the first AC term in the second brackets of (4.8).

2) To reduce the LF ripples in the regulated FM SPC in DCM, f_m should be less than open-loop gain cut-off frequency f_{cut} at least several times, as it can be deduced from Fig. 4.4.

3) As the modulating signal triangular *m(t)* should be used. It is recommended to use sawtooth *m(t)* when $f_m > f_{cut}$. The sawtooth *m(t)* should not be used, if its first harmonics is in the range where $H_{full}(s)$ rises (when $f_m < f_{cut}$).

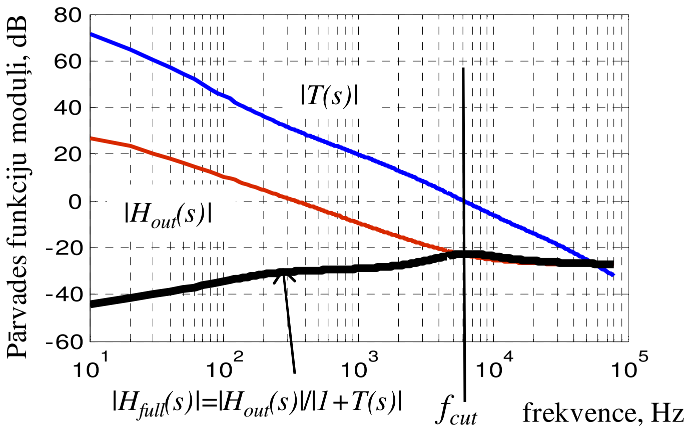


Fig. 4.4. Regulated buck FM SPC transfer functions in DCM. Parameters: C_{out}=470uF; L=42.5uH; Δf_{sw}=30kHz; r_{cout}=0.043Ω; r_L=0.07Ω; R_{out}=30Ω; f_{sw}=80kHz; f_{cut}=6kHz.

4) If it is possible to choose f_{sw}, then the higher f_{sw} is the better, because it allows us to get lower $V_{LFp\text{-}p}$. This is because, firstly, f_{cut} can be increased; secondly, $V_{LFp\text{-}p}$ is inversely proportional to f_{sw} raised to the power of two.

4.2 Influence of FM on efficiency of SPC

In the chapter 4.2 original study of influence of FM on SPC losses and the efficiency η is presented. A procedure to calculate η in FM SPC is developed. Expressions to calculate the losses and η for FM boost SPC in CCM are derived. The theoretical results are confirmed experimentally. Recommendations to eliminate the negative effect of FM on η of FM SPC are proposed and verified experimentally.

In making the analysis of the publications [6,18,22] about FM negative effect on SPC losses and η, it has been concluded that there are severe ambiguities in understanding the problem, mainly because the analysis is experimentally based without relevant theoretical explanation. For instance, it is measured in [22] that η of FM 600W boost power factor corrector (PFC) decreases by several percent due to the use of FM. The authors in [22] explained this by increased losses in power inductor core. However the experimental investigation of the same PFC in [6] showed that FM does not affect η. It is also experimentally revealed in [18] that FM has no influence on η of FM 600W SPC. The theoretical explanation of the problem is only presented in our paper [32]. The explanation is also confirmed experimentally using FM PFC.

In making thorough examination of the problem it is concluded that losses of FM SPC $P_{lossmod}$ can be calculated using the expression for unmodulated SPC losses P_{loss} as follows:

$$P_{loss\,\mathrm{mod}} = \frac{1}{T_m}\int_0^{T_m} p_{loss}(t)dt = \frac{1}{T_m}\sum_{k=1}^{T_m/T_{sw}} P_{loss}(T_{sw,k})T_{sw,k} \tag{4.9}$$

where $p_{loss}(t)$ are instantaneous losses of FM SPC; $P_{loss}(T_{sw,k})$ are losses at k-th switching period $T_{sw,k}$; T_{sw} is unmodulated SPC switching period; T_m is modulating signal period. Note that the expression is only useful when T_m/T_{sw} is an integer number. In making detailed examination, it is concluded that SPC total losses can be divided into three separate components: losses independent on f_{sw} (P_{const}), linearly dependent losses ($P_{lin}(f_{sw})$), and losses nonlinearly dependent on f_{sw} ($P_{nonlin}(f_{sw})$). It is follows that total losses in unmodulated SPC are:

$$P_{loss} = P_{const} + P_{lin}(f_{sw}) + P_{nonlin}(f_{sw}) = A_1 + A_2 f_{sw} + P_{nonlin}(f_{sw})\,, \tag{4.10}$$

where A_1 and A_2 are the coefficients dependent on specific SPC topology. The coefficients derived in the doctoral thesis, for instance, for boost SPC are as follows:

$$P_{const} = I^2_{Lavg}[R_{LDC} + DR_{DS(on)} + (1-D)R_d + V_d(1-D)/I_{Lavg} + D(1-D)r_{Cout}] = A_1 \tag{4.11}$$

$$P_{lin}(f_{sw}) = \frac{1}{2}V_{out}I_{Lavg}(t_r + t_f)f_{sw} = A_2 f_{sw} \tag{4.12}$$

where V_d is Schottky diode threshold voltage, R_{Ldc} is the power inductor DC resistance; R_d is forward resistance of a diode; $R_{DS(on)}$ is a power MOSFET on-resistance; t_r and t_f are the power MOSFET rise and fall times respectively. P_{const} mainly relates to DC conduction losses, $P_{lin}(f_{sw})$ mainly relates to switching losses, but $P_{nonlin}(f_{sw})$ mainly relates to power inductor magnetic core losses and AC conduction losses. It is derived in the doctoral thesis that the losses in the unmodulated boost SPC are as follows:

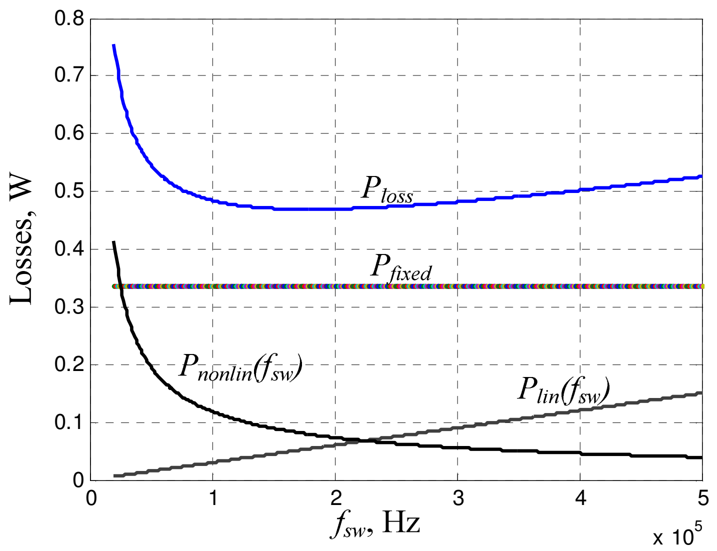


Fig. 4.5 Calculated P_{loss} and its components dependence on f_{sw} for unmodulated boost SPC in CCM with the powdered-iron-core power inductor ($L=46\mu H$; $C_{out}=330\mu F$; $V_{in}=5V$; $R_{out}=12.7\Omega$).

$$\begin{aligned} P_{nonlin}(f_{sw}) &= \frac{V_{in}^2 D^2 [R_{LDC} + DR_{DS(on)} + (1-D)R_d + (1-D)r_{Cout} + r_{Cin}]}{12L^2 f_{sw}^2} + \\ &+ C_m V_c \left(\frac{V_{in}D}{2NS}\right)^{\beta} \frac{1}{f_{sw}^{\beta-\alpha}} = \frac{A_3}{f_{sw}^2} + \frac{A_4}{f_{sw}^{\beta-\alpha}} \end{aligned} \tag{4.13}$$

where r_{cin} is input capacitor ESR; N is number of turns of power inductor; S is the magnetic core cross-section area; α, β are Steinmetz equation empirical coefficients that are usually listed in magnetic core material datasheets, e.g. [34]; V_c is the magnetic core volume. Nonlinearly dependent losses are more pronounced at lower f_{sw}, but linearly dependent ones at higher f_{sw}, that can also be seen in Fig. 4.5.

Equation (4.9) can be rewritten as:

$$P_{loss\,mod} = A_1 + A_2 f_{sw} + \frac{1}{T_m}\sum_{k=1}^{T_m/T_{sw}} P_{nonlin}(T_{sw,k})T_{sw,k} \tag{4.14}$$

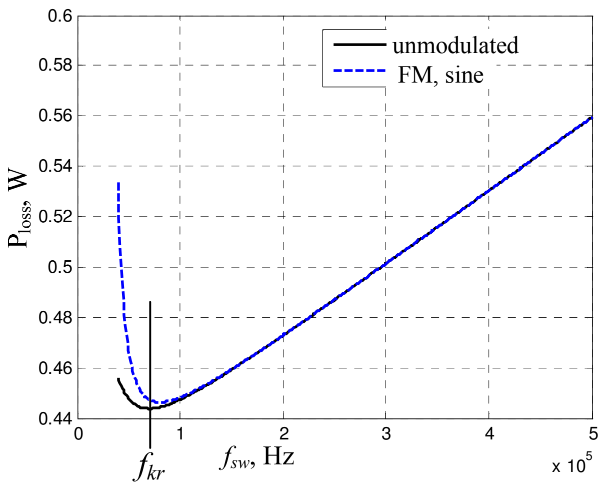
As it can be deduced the first and the second terms in the right of (4.14) are the same as in (4.10) for unmodulated SPC. This means that FM can only affect the losses that are nonlinearly dependent on f_{sw}. Assuming that T_m is higher than T_{sw} at least by an order of magnitude and that f_{sw} is periodically modulated by *m(t)* being an odd function (e.g. sine or triangular), (4.14) can be shown to be:

$$P_{loss\,\mathrm{mod}} = P_{const} + A_2 f_{sw} + \frac{1}{T_m}\int_0^{T_m} P_{nonlin}(f_{sw}(t))dt \tag{4.15}$$

It follows that for triangular or sawtooth modulating signal, nonlinear dependence of the losses on f_{sw} in FM SPC is:

$$P_{nonlin\,\mathrm{mod}}(f_{sw}) = \frac{A_3}{f_{swc}^2 - \Delta f_{sw}^2} + A_4 \frac{(f_{swc} + \Delta f_{sw})^{\alpha-\beta+1} - (f_{swc} - \Delta f_{sw})^{\alpha-\beta+1}}{2\Delta f_{sw}(\alpha-\beta+1)} \quad . \tag{4.16}$$

It is concluded in the doctoral thesis, that the nonlinearity is almost negligible when f_{sw} is higher than the critical frequency f_{kr} at which P_{loss} has its minimum. If FM SPC minimum switching frequency $f_{swmin}=f_{sw}-\Delta f_{sw}$ is higher than f_{kr} then modulating f_{sw} even with very high switching frequency deviation Δf_{sw} have almost insensible impact on the total losses and η. However, if $f_{swmin}<f_{kr}$ where the nonlinearity is dominant, then the total losses increase as Δf_{sw} increases. This can also be deduced from the results obtained both theoretically and experimentally as shown in Figs. 4.6 and 4.7.



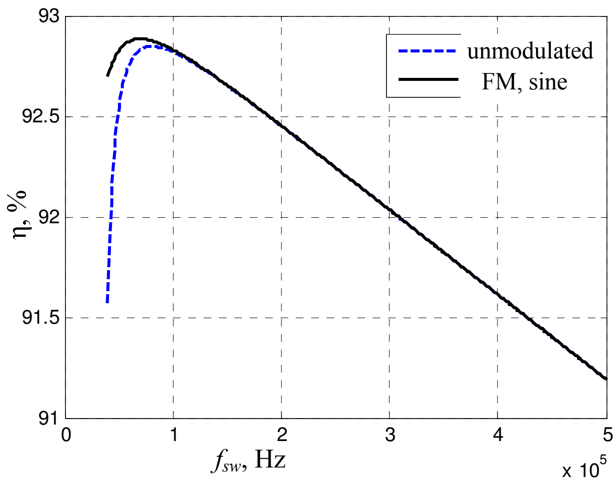


Fig. 4.6 Calculated $P_{lossmod}$ and η versus f_{sw} for unmodulated and FM boost SPC. FM parameters: f_{swc}=70...500kHz; Δf_{sw}=60kHz; f_m=1kHz; *m(t)* is sine wave.

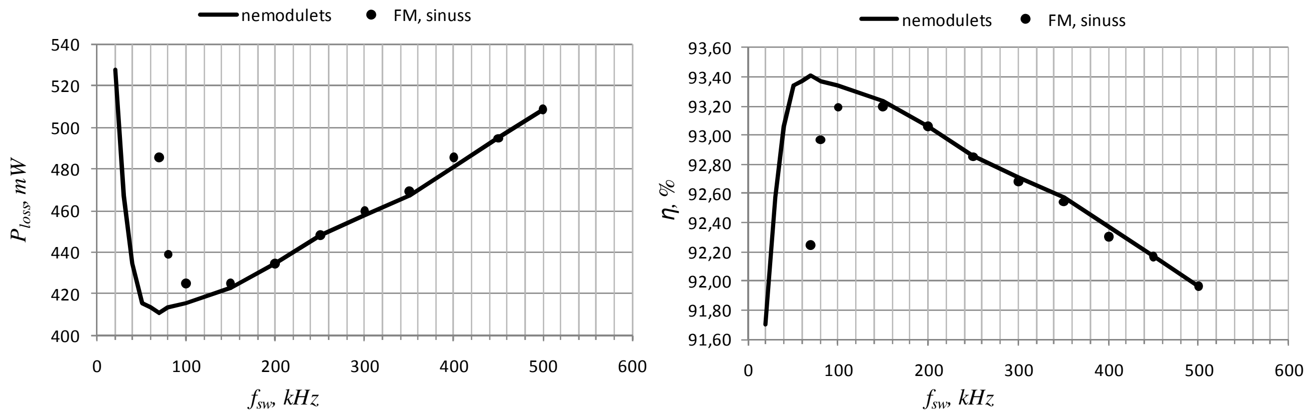


Fig. 4.7 Experimental η and the total losses versus f_{sw} for unmodulated and FM boost SPC. Modulation and power stage parameters and the power inductor are the same as in Fig. 4.6.

Recommendations to increase the efficiency in FM SPC

- When designing FM SPC, minimum switching frequency $f_{swmin}=f_{sw}-\Delta f_{sw}$ should be higher than critical frequency f_{kr}, because it provides SPC operation in the pronouncedly linear frequency region and in turn the negative effect of FM on the losses and η is negligible. If it is impossible to satisfy the condition, then Δf_{sw} should be small as possible considering naturally the trade-off between A_{EMT} and η.
- As modulating signal, *m(t)* with lower averaged rectified value (e.g. triangular) should be used.
- Since FM can cause rather high power inductor current LF ripples, they needs to be lowered.

4.3 Influence of FM on SPC input power quality

In the chapter 4.3 the negative effect of FM on the input current total harmonic distortion (THD), power factor (PF) and the input current spectrum in LF region has been examined thoroughly in both CCM and DCM. The causes of the problems have been found out and recommendations to solve the problems effectively have been proposed and verified experimentally using FM PFC.

In making the analysis of the publications [6,22,24,31] it is revealed that the main causes of worsening the input power quality of FM SPC have not been revealed yet, and the recommendations to mitigate the negative effect of FM on the input power quality have not been proposed yet in the publications.

It is concluded in the doctoral thesis that FM can increase the distortion of the input current and reduce PF in SPC. This is because FM causes LF ripples with f_m in SPC input current. In a general case the input current i_{in} of AC-mains-connected FM SPC consists of three components: LF current pulses with AC mains frequency f_{mains}, HF switching ripples

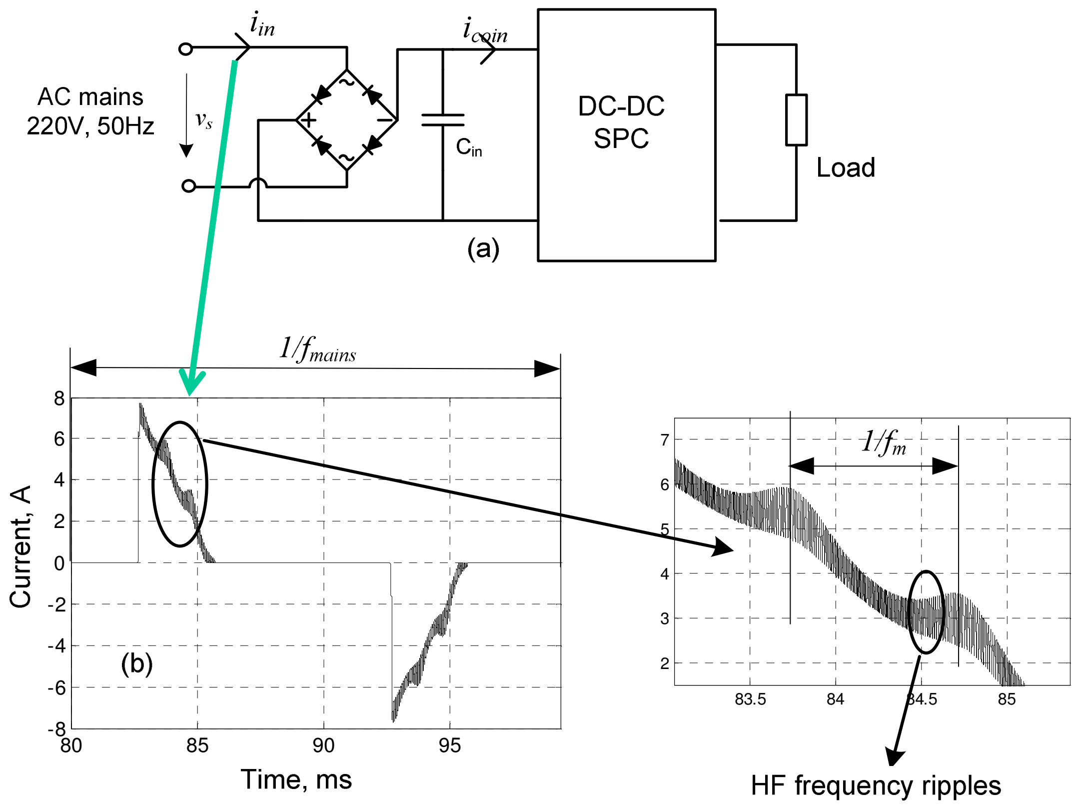


Fig. 4.8 Block diagram of a typical AC-mains-connected SPC (a) and simulated unregulated boost FM SPC input current waveform in DCM (b). Note that the boost FM SPC was simulated with a rectifier, filtering capacitor C_{in}=100uF and an AC sinusoidal source with an amplitude of 300V and frequency of 50Hz connected to its input. Other parameters: C_{out}=330uF; L=250uH; $m(t)$ is sine; Δf_{sw}=30kHz; f_{sw}=100kHz; R_{out}=1kΩ; f_m=1kHz.

and LF ripples with f_m caused by FM (see Fig. 4.8). The LF components can reduce PF and increase the distortion. The input current i_{in} LF ripples are originally generated in DC-DC SPC (which is usually the main part of AC-mains-connected FM SPC) input current i_{coin}. So to examine the effect of FM on the input power quality and calculate THD and PF for FM SPC, it is necessary, firstly, to calculate i_{coin} spectrum in the LF frequency region, secondly, to use standard expressions for THD and PF calculation, thirdly, to estimate impedances of the input capacitor, diode bridge and AC source, and fourthly, to assume that i_{in} flows only short interval of one AC mains cycle.

In making the examination it is concluded that in general DC-DC FM SPC input current spectrum consists of f_m harmonics that are responsible for the input current LF ripples and side frequencies that account for the HF switching ripples. FM causes interharmonics with frequencies $nf_m \pm mf_{mains}$ (where m≥1 and n≥1) in the LF region of the input current of AC-mains-connected SPC. The causes of the LF components in FM DC-DC SPC input current are the same as for the output voltage LF ripples. In other words nonzero difference between the switching delays $|t_d|$ is the main cause of the problems in CCM. But the fact that the input current averaged to T_{sw} depends on f_{sw} which is modulated is the main cause of the problems in DCM. In the doctoral thesis analytical expressions for calculation of the LF components of the input current i_{coin} spectrum are derived. The expressions give us a possibility to estimate THD and PF. It is important to note that amplitudes of the LF components due to FM are directly proportional to Δf_{sw} and $|t_d|$ in CCM, but they are approximately proportional to Δf_{sw} and approximately inversely proportional to f_{sw} raised to the power of two in DCM. This means that increasing Δf_{sw} and $|t_d|$ causes THD to increase but PF to decrease in AC-mains-connected FM SPC in CCM. Meanwhile THD can be reduced but PF can be increased by reducing the ratio $\Delta f_{sw}/f_{sw}$ in DCM.

THD and the LF components due to FM depend also on the modulation frequency. If f_m is slightly higher than maximum power analyzer frequency F_{max}, then from the normative point of view the input power quality will not worsen. However in this case the LF components can be regarded as EMI in the LF region (this is because several standards, e.g. CISPR16 require EMI measurements from 9kHz). If f_m is far beyond maximum of regulated FM SPC inductor current to output voltage transfer function, then attenuation of the LF components is high. This is provided by the feedback loop at lower frequencies and by the output capacitor and power inductor impedances at higher frequencies. The situation is worse in DCM because the LF components can be reduced only if f_m is lower than f_{cut} at least several times or reducing the ratio $\Delta f_{sw}/f_{sw}$.

The use of FM causes appreciably lower problems in traditional AC-mains-connected SPC than in PFC. This is because FM-caused-input-current-distortion is reduced by a high-capacity filtering capacitor following a rectifier and due to the fact that the input current flows only short time interval.

Recommendations to improve the input power quality of FM SPC

- From the normative point of view the negative effect of FM on the SPC input power quality can be completely neutralized when f_m is slightly higher than F_{max};
- as the main mode of the operation of FM SPC, CCM should be chosen. In the mode it should be provided that $|t_d|$ is as low as possible. This can be achieved using, for example, the techniques that were used to get small output voltage LF ripples. f_m should be chosen far beyond the maximum of regulated FM SPC inductor current to output voltage transfer function considering of course that f_m affects A_{EMT}.
- if DCM is chosen to be the main mode of the operation of FM SPC, then f_m should be lower than f_{cut} at least several times (note that for PFC f_{cut} is input current open-loop gain cut-off frequency). One more way to reduce the problem in DCM is to reduce the ratio $\Delta f_{sw}/f_{sw}$. This can be achieved, for instance, by increasing f_{sw} without changing Δf_{sw}.

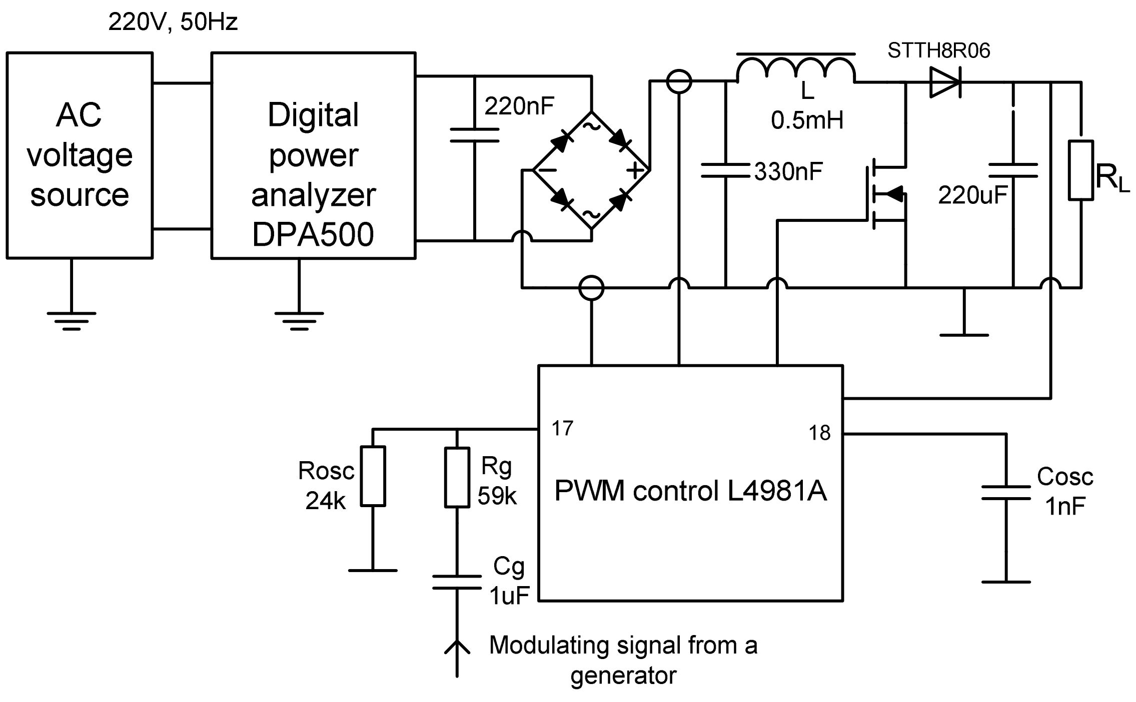


Fig. 4.9 Simplified schematic of the experimental FM PFC.

The recommendations proposed for improving the input power quality in CCM also have been verified experimentally using 360W FM PFC shown in Fig. 4.9. PF and the input current THD were measured using the power analyzer DPA500. Table 4.1 presents measured THD and PF for different Δf_{sw}, f_m and output power P_{out}=360W (nominal) and P_{out}=102W.

Table 4.1

Measured PF and input current THD for unmodulated and FM PFC.

Δf_{sw}, kHz	**f_m, kHz**	P_{out}=102W		P_{out}=360W	
		JK	**THD, %**	**JK**	**THD, %**
unmodulated		0.983	9.8	0.993	6.45
10	1	0.983	9.8	0.993	6.45
20	1	0.981	9.9	0.992	6.5
30	1	0.982	9.85	0.992	6.5
40	1	0.981	9.95	0.991	6.6
10	5	0.983	9.8	0.993	6.45
20	5	0.983	9.8	0.993	6.45
30	5	0.983	9.8	0.993	6.45
40	5	0.982	9.8	0.993	6.45

The results obtained show that the input power quality does not worsen for f_m=5kHz and different Δf_{sw} using FM even with high Δf_{sw}. However when f_m=1kHz then THD increases slightly. This can be explained by the fact that when f_m=5kHz, the frequencies of the LF components caused by FM are beyond the power analyzer measurement range (note that DPA500 maximum analysis frequency is F_{max}=3kHz). That is why the negative effect of FM on PFC input power quality from the normative point of view can be completely neutralized. Meanwhile when f_m=1kHz, increase in THD is negligible, because the recommendations proposed are considered as follows: firstly, f_m at least several times lower than input current open loop gain cut-off frequency (for this PFC f_{cut}=10kHz), and secondly, the artificial delay circuit proposed in the doctoral thesis was used to get small t_d in CCM. It can be concluded that the experimental results confirm that the recommendations proposed can be very effective to eliminate the negative effect of FM on PFC input power quality.

CONCLUSIONS

It can be concluded from the results obtained that FM as one of the spread spectrum techniques is very effective solution to reduce EMI because the main characteristics of SPC such as the output voltage ripples, the efficiency and the input power quality (mainly power factor and THD) are not virtually affected due to FM, if the recommendations proposed in the doctoral thesis are considered. The results obtained in the research also show that the use of FM has to play a large role in reducing EMI of SPC. This is because apart from its advantages it has also relatively smaller drawbacks, than the other EMI suppression techniques because, firstly, FM is sufficiently easy to implement and it practically do not increase the size and cost of SPC (especially if it is implemented on a PWM integrated circuit), secondly, it has relatively small negative effects on important SPC quality characteristics and sometimes the adverse effects can be even neutralized completely (e.g. it is proved in the thesis that from the normative point of view the negative effect of FM can be completely neutralized when f_m is slightly higher than maximum power analyzer frequency F_{max}).

The effectiveness of the use of FM mainly defined by the first harmonic attenuation, depends not only on modulating signal parameters (as it was assumed in the majority of publications analyzed) but also on the switching frequency f_{sw} and power stage parameters. The attenuation can be appreciably improved by properly choosing f_{sw} or using the optimized double-slope modulating waveform for which optimal parameter t_0 value is to be found using the procedure developed in the doctoral thesis.

The LF ripples that have been considered as the main problem of the use of FM, in principle are not significant problem as it also is proved in the thesis. For example, the LF ripples can be significantly mitigated in CCM by reducing the difference between the switching delays which is the main cause of the problem. This can be achieved, for example, by using proposed artificial delay circuit to get small t_d. The LF ripples in DCM can be reduced only if f_m is lower than f_{cut} at least several times or by lowering the ratio $\Delta f_{sw}/f_{sw}$. From the HF switching ripples point of view the boost or related topologies should be used for FM SPC.

The input current LF components caused by FM are responsible for worsening the input power quality. The causes of the problems are the same as for the output voltage LF ripples.

The input power quality of FM SPC can be appreciably improved taking in account the similar recommendations as for reducing the LF output voltage ripples.

The results obtained in the thesis also show that for higher f_{sw}, the negative effect of FM on the output voltage ripples and the input power quality is lower. In fact, this coincides with the modern SPC development trend to increase SPC switching frequencies.

The results presented and the calculation procedures developed could be used for designing and developing high-quality FM SPC. In general, we can conclude that FM is very effective solution to reduce EMI in SPC with small negative effect on FM SPC main quality characteristics if the results and the recommendations proposed are considered.

List of Main Abbreviations

SPC	Switching Power Converter
EMI	Electromagnetic Interference
FM	Frequency Modulation; Frequency Modulated
CCM	Continuous Conduction Mode
DCM	Discontinuous Conduction Mode
MOSFET	Metal-Oxide-Semiconductor Field Effect Transistor
LISN	Line Impedance Stabilization Network
AM	Amplitude Modulation
PFC	Power Factor Corrector
THD	Total Harmonic Distortion
LF	Low Frequency
HF	High Frequency
RBW	Resolution Bandwidth
ESR	Equivalent Series Resistance

List of Main Symbols

V_{in}	Input DC voltage	D	Average duty ratio
V_{out}	Output DC voltage	A	Attenuation of the1st harmonic amplitude of periodic rectangular pulse train
η	Efficiency	A_{EMT}	Attenuation of the 1st harmonic amplitude of LISN radio frequency (RF) output voltage
C_{out}	Output capacitance	t_d	Difference of the switching delays
C_{in}	Input capacitance	d	Instantaneous duty ratio
L	Inductance of a power inductor	$H_{co}(s)$	Control-to-output transfer function
R_{out}	SPC load resistance	$\langle\rangle$	Voltage/current averaged to the switching period
f_{sw}	Unmodulated SPC switching frequency; FM SPC central switching frequency	f_{cut}	Open-loop gain cut-off frequency
f_m	Modulation frequency	$\sim$	Small AC component of averaged voltage/current
$m(t)$	Modulating signal with unitary amplitude	$T(s)$	Open-loop gain
Δf_{sw}	Switching frequency deviation	$A_{HF}(t)$	Envelope of HF output voltage ripples
T_m	Modulating period	$V_{op\text{-}p}$	Unmodulated SPC peak-to-peak output voltage ripples
r_{cin}, r_{cout}	Input/Output capacitor ESR	$V_{ofmp\text{-}p}$	FM SPC peak-to-peak output voltage ripples

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