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Effects of Spread Spectrum on Output Filter of Buck Converter

J. Jankovskis[1], D. Stepins[1], N.Ponomarenko[1]
[1]*Department of Electronics and Telecommunications, Riga Technical University, Azenes St. 12–407, LV-1048 Riga, Latvia*
jankovsk@rsf.rtu.lv

***Abstract*—Consideration is being given to the use of spread spectrum technology (SST) as the one for noise suppression and the one with coexisting action on the components of output filter of buck converter. The analysis shows that the use of SST produces an additional stress on the output filter components such as power inductor and output capacitor. For both components switching modulation generates extra power loss (as opposed to unmodulated case) which increases with the growth of frequency deviation. During this more energy is dissipated just within frequencies which are below nominal switching frequency, than is saved within frequencies that are higher than the nominal switching frequency. But this effect is not strong; it is possible to compensate it by slight increase of nominal switching frequency and thus to maintain high efficiency of converter in the case of SST usage as well.**

***Index Terms*—Buck converter, electromagnetic interference, output filter, power losses, switching frequency modulation.**

I. INTRODUCTION

Electronics provides the means both for processing of signals and power. In typical electronic equipment operation within these two main domains brings to general net characteristics of system: size, weight, and efficiency. In the case of these metrics just the power supplying devices are the parts defining in large measure the total characteristics of electronics on a whole first of all because they are used in innumerable units thus taking part in conversion of considerable amounts of energy. In the past decades the power supply has been making significant progress through gradual replacement of linear regulators supply with switch mode one (SMPS). At the same time the progression of power supply comparatively is not wholly satisfactory. Indeed, information processing electronics based on microprocessors is developing truly exponentially (i.e., is following Moore's Law). Contrary to this power supply electronics based on the highest efficiency converters – SMPS ones is developing much more slowly: its power density is doubling only in every ten years [1] since 1970), but even this is obtained mainly through the possibilities to use higher switching frequencies, f_{sw} which in their turn are growing by a factor approaching to ten in recent decades [1]. Nevertheless the operation with as high as possible f_{sw} (following from the use of proper topology, available advanced components [2] and apt design) do more than just improve the power density and efficiency; in fact f_{sw} affects nearly every performance characteristic of the supply [3], [4]. The use of highest f_{sw} requires new attention placed on parasitic impedances, effects from printed circuit board (PCB) layout, potential sources of large output ripple switching spikes, and electromagnetic interference (EMI) – both in forms of conducted and radiated ones [5].

Just the suppression of EMI is as a rule the primary task not only with high f_{sw} converters but in fact with all ones realizing hard switching (i.e. high efficiency [6]). Decrease of EMI is possible to achieve both by traditional passive methods (filtering, shielding, grounding, proper PCB design, a.o.) and by active ones – spread spectrum technologies (SST) [7]. The latter one (known as dithering as well) allows the suppression of f_{sw} harmonics in the point of their generation – switching transistors without intervention into power parts of converter. The use of SST in concept is the frequency modulation (FM) of f_{sw} with the modulating signal f_m ($\mathrm{f_m} \ll \mathrm{f_{sw}}$). The modulation suppresses the amplitudes of each harmonic of naturally discrete harmonic spectrum of unmodulated f_{sw} by converting them in corresponding noise bands but with significantly less noise intensities. So the energy emitted by hard switching is not concentrated as yet in high amplitude discrete harmonics but is spreaded out more evenly over a larger frequency range with the top levels of EMI considerably reduced. Hence the appropriate maximum f_{sw} would be determined by EMC assurance methods used as well.

Thus the use of SST creates a multitude of high frequencies attacking the output LC-filter of converter and creating specific operational frequency effects. Power dissipated in period of switching modulation T_m by filter components is changing in time t in accordance with the change of most powerful component of spectrum $f_{sw}(t)$ within $\mathrm{f_{sw\,min}} \ldots \mathrm{f_{sw}(t)} \ldots \mathrm{f_{sw\,max}}$. In this research there are examined the additional power dissipation effects in output filter components generated by the use of SST in itself as well as the effects under different levels of load.

II. THE CONVERTER AND SPREAD SPECTRUM USED

In the research closed-loop buck converter (Fig. 1) operating in continuous current mode (CCM) is used. The

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buck converter can operate both as unmodulated and FM converter. The power stage of the converter consists of power MOSFET *Q1*, Schottky diode *VD1*, electrolytic output capacitor C_{out}, power inductor L with the ferrite core and output resistive load R_{load}. Nominal output current is 200mA and output DC voltage 4.5 V.

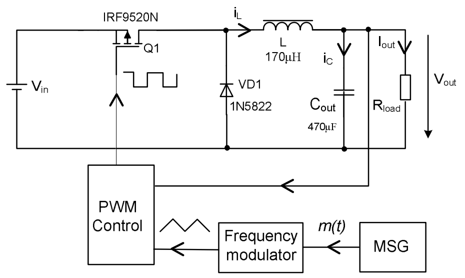

Fig. 1. Schematic of FM closed-loop buck converter used.

In order to spread the spectrum and reduce peak EMI levels, the control circuit consists not only of PWM controller SG2524 but also of frequency modulator and modulating signal generator (MSG). The triangular-like output signal of the frequency modulator is fed into PWM controller comparator input. If MSG is disconnected from the frequency modulator, then power MOSFET *Q1* f_{sw} is constant (unmodulated). If MSG is connected to the frequency modulator then f_{sw} is modulated by $m(t)$. The instantaneous switching frequency in this case is

$$f_{sw}(t) = f_{sw0} + \Delta f_{sw} m(t)\,, \tag{1}$$

where Δf_{sw} is the switching frequency deviation (in our experiments 35kHz); $m(t)$ is triangular modulating signal; f_{sw0} is the nominal switching frequency (in our experiments 100kHz).

In Fig. 2. simulated input current spectra are shown for both unmodulated and FM buck converter. The effect of FM to reduce EMI of the buck converter is clearly seen. After applying FM the noise energy concentrated in discrete harmonics is distributed over a wider range of frequencies and peak EMI levels noticeably reduced [8]–[12].

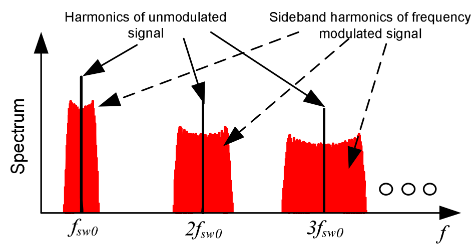

Fig. 2. Simulated spectra of unmodulated and frequency modulated input current.

III. GENERALITIES OF THE OUTPUT FILTER

To a first approximation the values of idealized, losses L and C comes from the principal analysis of the topology and allowed level of ripples (for the case of CCM) [8]:

– From the ripples Δi_L of inductor current i_L

$$\Delta i_L = Y_{out}(1-D)/f_{sw}L\,; \tag{2}$$

– From the ripples ΔV_{out} of output voltage V_{out}

$$\Delta V_{out} = (\pi^2 V_{out}/2)(1-D)(f_c/f_{sw})^2\,, \tag{3}$$

where D is the duty ratio, $D = V_{out}/V_{in}$ with V_{in} as the input voltage; f_C is the corner frequency of *LC*-formed low-pass filter, $f_C = 1/2\pi\sqrt{LC}$; $f_C \ll f_{sw}$. With the proviso that the values of L and C are already chosen, from Eqs.(1), (2) follows that in the case of SST more ripples are observed just at $f_{sw\,min}$.

In reality currents through L and C (respectively i_L and i_C) should produce the losses. To account for these losses (and for possible supplementary frequency effects at some particular high frequencies within spread spectrum) there are need to consider the parasitics of filter components. Specifically, for usual three elements equivalent circuit of components there are two parasitics for each: the equivalent series resistance (r_C, ESR) and inductance (L_C, ESL) for capacitor and the equivalent series resistance (R_L) and parallel self-capacitance (C_0) for inductor. There is no general relation allowing connecting L and C values with their resistances; at best it is possible to point on characteristic values, correlation, and tendencies within groups [9], [10].

In the following it is assumed that losses generated in the components equal to their resistance times squared root-mean square current through them. Besides, for evaluation of impact of SST on filter we are not interesting about absolute value of losses but on their change relative to unmodulated converter when there is only one value of $f_{sw} = f_{sw0}$. This allows, similar as in [13], [14], from different loss terms to account only for loss terms which are *const* or non-linear functions of f_{sw}, assuming that linear ones are compensating themselves in the case of symmetrical FM.

IV. SST CAUSED EFFECTS ON INDUCTOR

A. *Without effects of coil saturation*

To evaluate effect of FM on power inductor losses, first of all unmodulated buck converter power inductor loss will be analyzed. In general power inductor power loss can be divided into winding (P_w) and magnetic core (P_{mc}) losses. The winding losses for the unmodulated buck converter inductor

$$P_w = R_L I_L^2\,, \tag{4}$$

where R_L is the power inductor resistance. I_L is the rms current of inductor. Considering that inductor current consists of both AC I_{Lac} and DC I_{out} components (Fig. 3), I_L is

$$I_L = \sqrt{\frac{1}{T_{sw}}\int_0^{T_{sw}} (i_L(t))^2\,dt} = \sqrt{I_{out}^2 + I_{Lac}^2}\,, \tag{5}$$

where

$$I_{Lac} = \frac{(V_{in} - V_{out})DT_{sw}}{2\sqrt{3}L} \tag{6}$$

is the effective value of $i_{Lac}(t)$ (Fig. 3). From (4) – (6) it can be derived that

$$P_w = R_L I_L^2 = R_L I_{out}^2 + \frac{R_L D^2 (V_{in} - V_{out})^2}{12 L^2 f_{sw}^2}. \qquad (7)$$

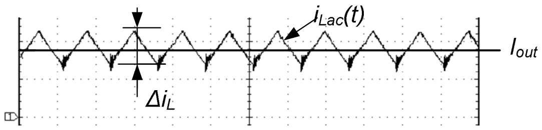

Fig. 3. Unmodulated buck converter power inductor current. (Scale: 10us/div; 100mA/div).

In order to derive magnetic core losses P_{mc} the modified Steinmetz equation [13] for nonsinusoidal waveforms can be used

$$P_{mc} = C_m f_{eq}^{\alpha-1} (\Delta B / 2)^{\beta} f_{sw} v_c , \qquad (8)$$

where C_m, α, β are Steinmetz equation empirical coefficients that are usually listed in magnetic core material datasheets; v_c is the magnetic core volume; ΔB is the difference between maximum and minimum magnetic induction; f_{eq} is the equivalent frequency according to [13]

$$f_{eq} = \frac{2}{\Delta B^2 \pi^2} \int_0^{T_{sw}} \left(\frac{dB}{dt} \right)^2 dt . \qquad (9)$$

Assuming that inductor current is linearly changing as shown in Fig. 3 it can be derived for unmodulated buck converter from (8) that

$$P_{mc} = C_m \left[\frac{2}{D(1-D)\pi^2} \right]^{\alpha-1} \left(\frac{(V_{in} - V_{out}) D}{2NS} \right)^{\beta} \frac{v_c}{f_{sw}^{\beta-\alpha}}, \qquad (10)$$

where S is the magnetic core cross-section area; N is the number of turns.

As it can be deduced from (7) and (10) inductor power loss for unmodulated converter consists of two terms: loss that is independent on f_{sw} and loss which is nonlinearly dependent on it. It is obvious that nonlinearly dependant on f_{sw} losses can increase due to FM even if the coil is not saturated.

In the following let's analyze the losses in the case of FM converter. In this situation the average power losses of the inductor can be calculated as follows

$$P_{loss\,\mathrm{mod}} = \frac{1}{T_m} \int_0^{T_m} p_{im}(t) dt , \qquad (11)$$

where $p_{im}(t)$ – instantaneous losses in power inductor. Taking into account that in the case of FM T_{sw} is not constant then the average power loss $P_{ind}(T_{sw,k})$ for each switching cycle $T_{sw,k}$ is function of it. So (11) can be rewritten as follows

$$P_{loss\,\mathrm{mod}} = \frac{1}{T_m} [\int_0^{t_1} p_{im}(t) dt + \int_{t_1}^{t_2} p_{im}(t) dt + ... + \int_{t_{k-1}}^{t_k} p_{im}(t) dt +$$

$$+ \int_{t_{N-1}}^{t_N} p_{im}(t) dt] = \frac{1}{T_m} \sum_{k=1}^{N} P_{ind}(T_{sw,k}) T_{sw,k} , \qquad (12)$$

where N is an integer number, $N = T_m / T_{sw}$. Considering that T_m is usually much higher that T_{sw} then instead of summation the integration can be used as follows

$$P_{loss\,\mathrm{mod}} = \frac{1}{T_m} \int_0^{T_m} P_{ind} (f_{sw}(t)) dt . \qquad (13)$$

So the average power losses of the inductor can be calculated using (1), (7), (10) and (13). After using these Eqs. it is concluded that FM can increase inductor loss, due to nonlinearly dependent on f_{sw} terms in (7) and (10).

B. With saturation effects of core

It is well known that magnetic core saturation is very problematic in SMPS. Therefore it is of importance to examine possible side effects of FM on power inductor when core is entering saturation. In order to examine these effects output current of the buck converter was increased of about three times nominal current (from 200mA to 600mA). Experimental inductor current is shown in Fig. 4 (when I_{out}=200mA) and Fig. 5 (when I_{out}=600mA). As it can be seen in Fig. 4 there is no saturation of magnetic core due to the use of FM. However sharp increase in the inductor current is observed when I_{out}=600 mA as it is seen in Fig. 5.

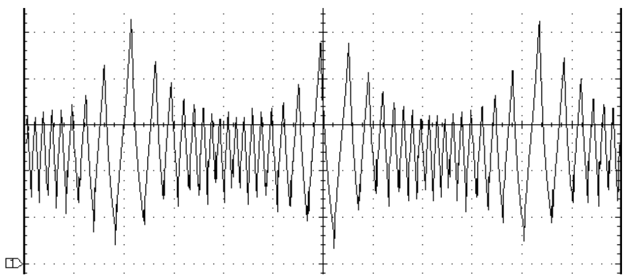

Fig. 4. FM buck converter power inductor current without core saturation effects. Parameters: f_{sw0}=100kHz; f_m=5kHz; Δf_{sw}=35kHz. (Scale: 50us/div; 100mA/div).

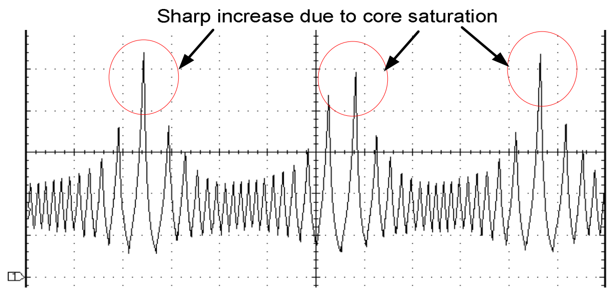

Fig. 5. FM buck converter power inductor current with core saturation effects. Parameters: f_{sw0}=100kHz; f_m=5kHz; Δf_{sw}=35kHz. (Scale: 50us/div; 400mA/div).

This is because the magnetic core materials saturates when f_{sw} is in the vicinity of f_{swmin}. For the frequencies which are noticeably higher than f_{swmin} there is no saturation. Obviously power inductor losses can drastically increase due to SST effects on power inductor with the saturation effects of core. Therefore when designing FM power converter it is important to properly choose f_{swmin} to avoid the magnetic core saturation.

V. SST Effects on Capacitor

The actuality of ESR in C_{out} is placing primary emphasis upon effects of operating frequency and current. In SST operating frequency takes the instant values $f_{sw}(t)$ within $f_{swmin} \dots f_{swmax}$ according to the modulation. The operating effective value of current I_C comes from (7) as

$$I_C = (V_{in} - V_{out})D / 2\sqrt{3}Lf_{sw}(f) . \tag{14}$$

To make an assessment about power dissipated by C_{out} there is need to evaluate its ESR. The value of ESR is complex function of characteristics of materials and construction techniques used and of frequency f. To a first approximation the value is possible to present as the sum of all resistances from conducting elements, R_{scon} and that of dielectric ones, $R_{sd} = \tan\delta(f)/2\pi fC$

$$ESR(f) \approx R_{scon}(f) + \tan\delta(f)/2\pi fC, \tag{15}$$

where $\tan\delta$ is the net loss tangent (dissipation factor) of the dielectrics used. Based on assumption that R_{scon} mainly is related with the metallic elements from this it follows that the value must grow with f (at least by a skin-effect at high f). In its turn $\tan\delta(f)$ of dielectrics is growing in opposite direction – towards low frequencies [9]. This means that in principle *ESR*(*f*), presented as (15), should be a function with minimum. Published experimental results really show on such a dependence on f: *ESR*(*f*) is function with clearly defined minimum for high frequency ceramic capacitors [15] and *ESR*(*f*) is with broad and indefinite minimum for typical capacitors (Al and Ta electrolytic, multilayer polymer, X7R ceramics) used for SMPS [11]. Since C_{out} in SMPS typically is used at frequencies near (or slightly left of) minimum, it can be predicted with certainty $\pm(10 \dots 20)\%$ that $\mathrm{ESR(f)} \approx \mathrm{ESR} = \mathrm{const}$ for f_{sw} changes within $\pm\Delta f_{sw}$ in SST. By account of Eq.(14) the loss power generated by ripples

$$P_{rip} \approx ESR \cdot I_C^2 = const / f_{sw}^2(t) . \tag{16}$$

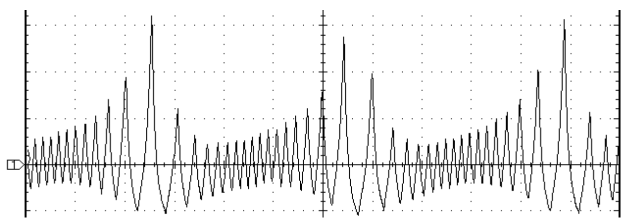

Fig. 6. FM buck converter output capacitor current when core enters saturation. Parameters: f_{sw0}=100kHz; f_m=5kHz; Δf_{sw}=35kHz. (Scale: 50us/div; 400mA/div).

This nonlinear relation shows that more losses are generated at lower frequencies; additionally it shows that there is greater increase of generated losses for $f_{sw}(t)$ within $-\Delta f_{sw}$ than that of decrease within $+\Delta f_{sw}$ as compared to ones at f_{sw0}. But this effect is not strong. Really, accounting that the average value of $\langle 1/f_{sw}^2(t) \rangle$ is $1/(f_{sw0}^2 - \Delta f_{sw}^2)$ this for our converter ($f_{sw0} = 100kHz$; $\Delta f_{sw} = 35kHz$) gives $\langle f_{sw} \rangle = 94kHz$. The overload of converter, as it is seen from Fig. 6, enhances this effect.

VI. Conclusions

The analysis carried out shows that the improvement in EMC of the converter by use of SST yet produces an additional stresses on the components L and C_{out} of output filter. For both components switching modulation generates extra power loss (as opposed to unmodulated case) which increases with the growth of frequency deviation $\pm\Delta f_{sw}$. During this more energy is dissipated just within $-\Delta f_{sw}$ than is saved (as against to f_{sw0}) within $+\Delta f_{sw}$. But this effect is not strong; it is possible to compensate it by slight increase of f_{sw0} and thus to maintain high efficiency of converter in the case of SST usage as well.

Magnetic core can enter saturation due to the use of SST. This reflects as sharp increase in the power inductor current and consequently increase in losses in the inductor. So when designing FM converter minimum switching frequency should be chosen properly to avoid core saturation.

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