SIMPLE RESONANT FREQUENCY TRACKER AND ESTIMATOR

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Abstract: One of the main difficulties of designing resonant power supplies is ensuring the switching frequency stays the same as resonant tank's main resonant frequency or deterministically shifted. The only way to achieve this is to track the switching frequency. Usual approach is to utilize a PLL. This article describes simple resonant frequency tracker, providing elegant and simple solution to the conventional PLL drawbacks and furthermore allowing the controller to simply estimate resonant current amplitude on half-period basis.

Keywords: resonant frequency tracking ZCS DC/DC LLC

1 INTRODUCTION

One of the most evident tendency in the field of DC/DC power conversion is since-the-beginnings lasting switching frequency increase. It enables designers to reduce size of converters and therefore manufacturing costs as well. Starting at a few hundreds of Hz decades ago, at times of the first SCRs and bulky low power density converters, up to tens of MHz in today's bleeding edge low power DC/DC converters. As the overall switching losses are proportional to the switching frequency, special attention is nowadays paid to lower these losses. One way of lowering them is to engage snubbers, effectively moving part of the tranistors'/diodes' switching losses onto the snubbers. Other way is to utilize the soft switching technique, therefore reducing the switch current/voltage during switch-on and switch-off. This is called a Zero Current Switching (ZCS) or Zero Voltage Switching (ZVS). An advantage of the resonant power supplies is therefore mainly improving conversion efficiency at the cost of increased control complexity.

The increase of a control subsystem complexity is mainly caused by the fact resonant converters are best suited for specific load conditions operation instead of no-to-full load operation, supported by ordinary hard switching converters. This leads to various different control approaches, including the novel Multiperiod Modulation (MpM) described in [1], [2] and [3].

In most resonant converter control approaches the ability to track resonant tank (RT) resonant frequency f_0 or to maintain the switching frequency f_{sw} to be strictly higher or lower then the non-constant f_{sw} is vital or at least allows further loss reduction. The common approach to track f_0 is to use a Phase-Locked Loop (PLL). The use of an analogue PLL is however usually very problematic and leads to huge settling times, orders of magnitude longer then the resonant period $T_0 = 1/f_0$ is. This could be enough if there would have been only long-term changes like temperature dependencies or components aging, but it is not the real situation. In the case of MpM (see [2] for details) the use of traditional PLL is even more complicated, making it practically unusable. There are several reasons for this to be true:

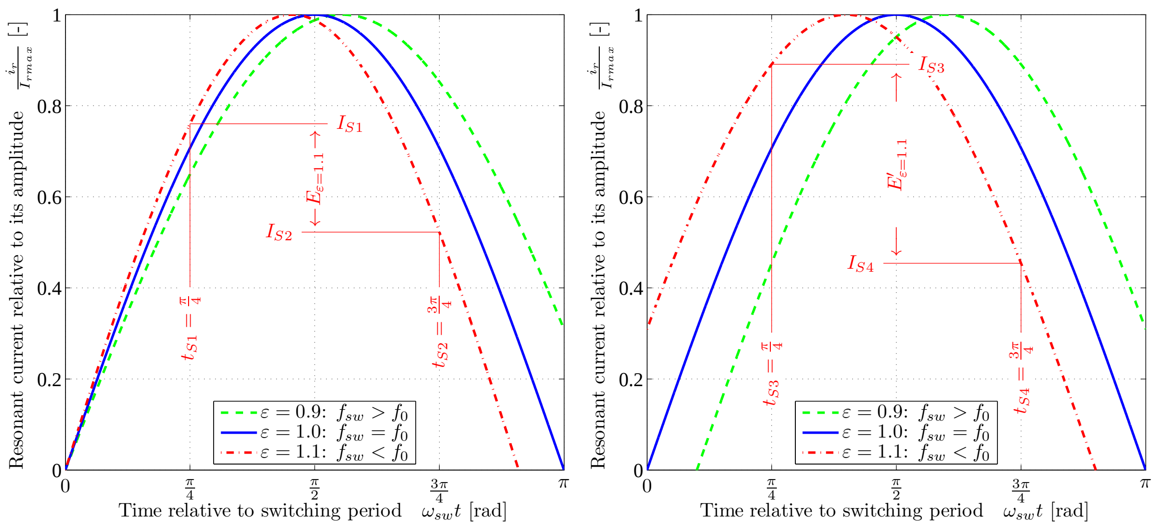
1. The resonant current $i_r(t)$ amplitude is by design often (cycle by cycle) lowering down to negligible values.

2. The resonant frequency f_0 changes rapidly due to skin and proximity effects in resonant capacitors and (although to far less extent) to a resonant induktance nonlinearity.

These problems can be partially solved using software PLL, while ignoring the low-amplitude i_r portions, but still it will not track reliably those rapid f_0 changes. To solve this problem, the resonant frequency estimation technique (published originally in [4]) was extended to fit in the MpM driven LLC resonant converter described in [2] and [3].

2 RESONANT FREQUENCY TRACKING

The resonant frequency tracking method is based on just sampling two values during every resonant half-period and comparing the results. The sample times of both samples (t_{S1} and t_{S2}) are placed symmetrically around the peak value at half of the resonant half-period. The situation is shown on the figure 1. Just like several other figures down this article, the figure is divided in two parts. The left one (figure 1a) depicts the situation just after oscillations were excited (in time $t = 0$). Right figure depicts the situation when the resonant current i_r is oscillating and any possible switching frequency error $\varepsilon \neq 1$ causes resonant current phase shift and out of resonance operation.



(a) First cycle after excitation. **(b)** Continuous resonance.

Figure 1: Resonant current $i_r(\omega_{sw}t)$ half-period with plotted sample times. The sampled values I_{S1} and I_{S2} are plotted for the case of $f_0 = 1.1 f_{sw}$.

The sampling times t_{S1} and t_{S2} could be selected differently, but the values of $\frac{\pi}{4}$ and $\frac{3\pi}{4}$ ensure both the widest work range and sensitivity. The relative switching frequency error

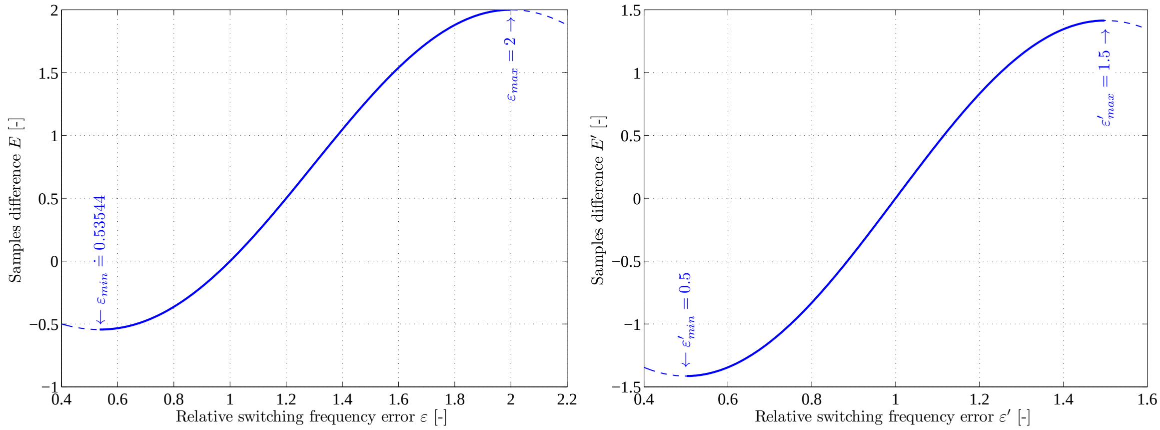
$$\varepsilon = \frac{f_0}{f_{sw}} \tag{1}$$

to the sampled current difference

$$E = I_{S1} - I_{S2} \tag{2}$$

transfer characteristic for the previously selected sampling times is shown on the figure 4. The important thing to note is the usability range of this method is restricted to the monotonous interval of

$$\varepsilon \in (0.53544; 2) \tag{3}$$



(a) First cycle after excitation. **(b)** Continuous resonance.

Figure 2: The relative switching frequency error ε to the sampled current difference E transfer characteristic. The usable range is emphasized using solid line.

during the first resonant half period and to

$$\varepsilon' \in (0.5; 1.5) \tag{4}$$

during subsequent half periods. Using (1), for resonant frequency f_0 the following is then true

$$f_0 \in (0.53544 f_{sw}; 1.5 f_{sw}). \tag{5}$$

At the first look the curve on figure 4a could be wrongly treaten as an offseted sinusoid, but it is not the case. The formula describing the transfer function is derived out of (2) as follows

$$E = I_{S1} - I_{S2} = \sin(\varepsilon\omega t_{S1}) - \sin(\varepsilon\omega t_{S2}) = \sin(\varepsilon\omega t_{S1}) - \sin(\varepsilon\omega(0.5 - t_{S1})). \tag{6}$$

For the figure 4b the sequence is similar, but the result is a sinusoid

$$\begin{aligned} E' &= I_{S3} - I_{S4} = \sin(\omega(t_{S1} + \varepsilon - 1)) - \sin(\omega(t_{S2} + \varepsilon - 1)) = \\ &= \sin(\omega(t_{S1} + \varepsilon - 1)) - \sin(\omega(0.5 - t_{S1} + \varepsilon)). \end{aligned} \tag{7}$$

The derived usability range must be fulfilled under all conditions, otherwise the algorithm becomes unstable and the switching frequency f_{sw} will settle at wrong value.

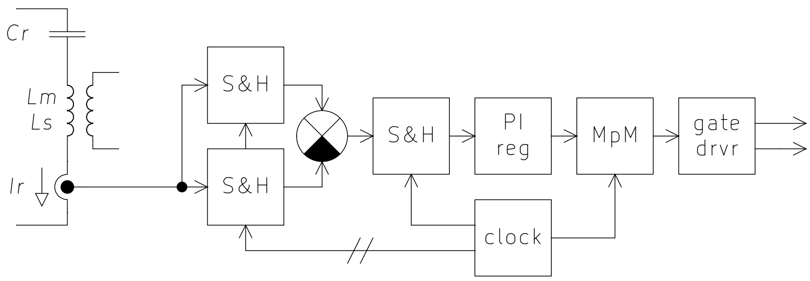


Figure 3: The resonant frequency tracker structure. (Block name abbreviatons: S&H – sample and hold module, PI reg – PI regulator, MpM – Multiperiod Modulation, gate drvr – gate drivers of the power stage transistors)

The difference between the two characteristics at figure 4 slightly complicates the switching frequency controller. As further investigation shows, around usual work point of $\varepsilon = 1$ the differences are negligible and the simple PI controller is sufficient.

3 RESONANT CURRENT ESTIMATION

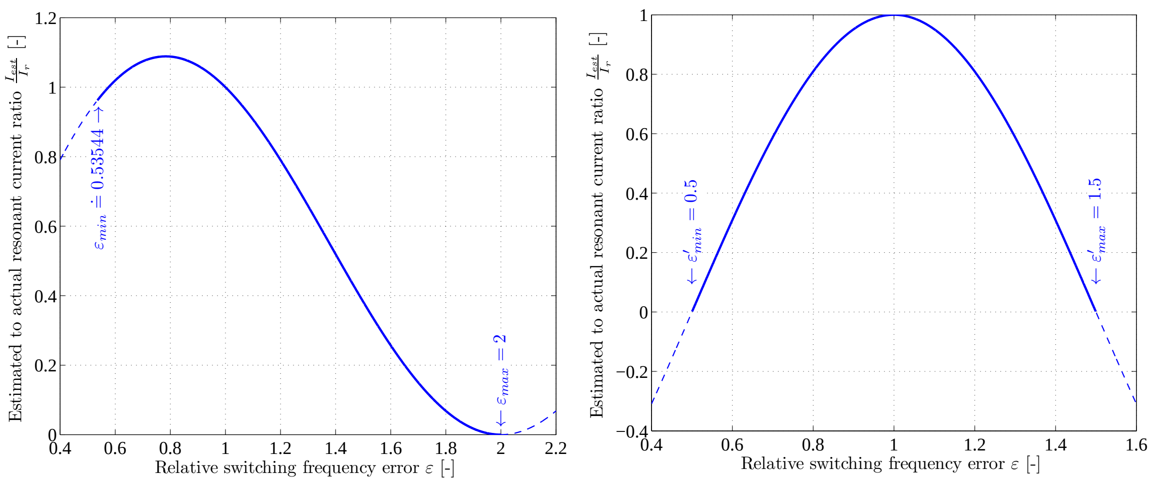
As for resonant frequency f_0 tracking we already have two samples of resonant current $i_r(t)$ amplitude at defined time, we can estimate actual resonant current amplitude I_{rmax} of this half-period directly by solving following formula

$$I_{rmax} = \frac{\sqrt{2}}{2}(I_{S1} + I_{S2}), \tag{8}$$

or RMS value of sinusoidal resonant current waveform as

$$I_r = \frac{I_{S1} + I_{S2}}{2}. \tag{9}$$

If we suppose resonant frequency f_0 is reasonably well tracked, the result will be precise enough for overcurrent detection and current control use. Following figures depict the current estimation accuracy on switching frequency error ε dependency.



(a) First cycle after excitation.

(b) Continuous resonance.

Figure 4: The estimated to actual current ratio $\frac{I_{est}}{I_r}$ on relative switching frequency error ε dependency. The usable range is emphasized using solid line.

Usually the converter control unit will be DSP driven so the use of oversampling by multiple A/D converter channels can be utilised as to gain better estimation accuracy and noise immunity. For precise resonant current estimation at $\varepsilon \neq 1$, results can be corrected applying inverse error functions or using more accurate I_r approximation than equation (9) is. As the estimator is supposed to work near the $\varepsilon = 1$ point, it was not dealt with during the current research.

4 PRACTICAL USE NOTICES

Utilizing a DSP as the control element greatly simplifies the design, resulting the structure on figure 3 to change significantly. As sample and hold modules are also included in the power electronics control dedicated DSPs (like Texas Instruments TMS320F28xxx family and others), all the blocks of figure 3 except gate drivers and resonant current sensor (usually current transformer) will become part of the DSP internal structures.

As was described in the previous section, it is very simple to extend the resonant frequency tracker by the resonant current estimator functionality, wasting only tiny computational resources of the DSP.

Various resonant converters types are supposed to run at switching frequency f_sw slightly lower or higher than the resonant f_0 is. This can be very easilly and deterministically achieved by simply adding another constant in the sum block of figure 3 and therefore introducing desired frequency shift. This method can be modified to another extent by using a variable instead of the constant and to precisely control switching versus resonant frequency shift. This may be used for variable frequency control or precise switching losses tuning and minimalization.

By using this (or similar) type of tracker and estimator with the multiperiod modulation, which ensures strict ZCS switching, most of the interferring switching noise pulses are lying far enough from the sampling moments. By this most of the switching noise related problems are effectively mitigated.

5 CONCLUSION

The article describes a novel method of tracking the resonant frequency of the resonant DC/DC converters. Its use is mainly focused on the Multiperiod Modulation LLC resonant converters, but is usable for other types as well. Several design limitations along with benefits were provided. The biggest benefit of using this method is a possibility to estimate resonant current amplitude on half period basis without any added complexity. This allows the designer to effectively implement fast current control loop and overcurrent protection.

Currently the proposed tracking and estimating strategy is being implemented in the prototype converter.

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