A 1 MHz wide bandgap power amplifier for high-precision applications

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A 1 MHz Wide Bandgap Power Amplifier for High-Precision Applications

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Keywords

«Harmonics», «Wide bandgap devices», «High frequency power converter», «Industrial application».

Abstract

The goal of this research is to optimize the total inductor volume of the extra bias inductor opposed current converter. Wide bandgap semiconductors are used to raise the switching frequency to 1 MHz for a further decrease of the inductor volume. The research is validated with a high-precision current amplifier prototype.

Introduction

Switched-mode power converters (SMPCs) are widely used in industry. For some applications, such as lithography and magnetic resonance imaging (MRI), their precision and linearity are important. Taking lithography for example, the position error of the moving stage is determined by the non-linearity error of the current amplifier [1]. In order to obtain a nm position accuracy, lithography requires less than -100 dB current distortion for the high-precision motor control. With the development of industry, those applications are raising their demands for higher bandwidth and smaller volume [2].

Increasing the switching frequency of the power converter is a solution to meet those demands since it increases bandwidth and reduces the volume of the filter components. However, higher switching frequencies cause larger power dissipation and temperature rise on the switches, which is a challenge for the widely-used Si-based MOSFETs. Besides, due to the finite turn-on and turn-off time of the switches, a dead-time is required in half-bridge (HB) or full-bridge (FB) converter. When the switching frequency becomes higher, the dead-time results in larger distortion, which is the most significant source of distortion in this topology [3].

Wide bandgap semiconductors, such as Gallium Nitride (GaN) transistors are an alternative to the Si-based MOSFETs. Since the switching speed of GaN transistors is faster than Si-based MOSFETs, the switching loss is largely reduced thus making it possible to switch at a higher frequency. To overcome the dead-time distortion, a different topology named by opposed current converter (OCC) is proposed in [4], which fully eliminates the dead-time and results in high linearity.

In [5] and [6], it is proved that the OCC topology has good linearity and is an appropriate substitute of HB or FB converter. A comparison of different topologies combined with the GaN transistors is done

based on simulations in [7]. It shows that GaN transistors are promising to use in OCC topology but it's not verified by experimental results yet. One of the disadvantages of the OCC topology is that the total inductor volume is increased due to the extra bias current to maintain the continuous current mode [8]. In [9], an extra bias inductor opposed current converter (ELOCC) is proposed to reduce the total inductor volume but the volume is not optimized yet.

In this research, the total inductor volume of ELOCC topology is modeled based on [9] but the accuracy of the model is improved by taking the rms value of the variable inductor ripple current into consideration. The total inductor volume is optimized by adapting the inductor current ripple ratio. GaN transistors are used to raise the switching frequency to 1 MHz for a further decrease of the inductor volume. The bandwidth of the converter is also improved. The research is verified by a high-precision current amplifier prototype with 320 V output voltage and 12.5 A output current.

The Extra Bias Inductor Opposed Current Converter

Compared to the equivalent HB converter, the OCC has two additional blocking diodes and one additional inductor, which makes it possible for the two switches to conduct at the same time. Consequently, no dead-time is required thus the dead-time distortion is eliminated. The two blocking diodes also force the filter inductor current going through each leg in one direction. If a bias current i_{bias} is added to each leg, the current in the filter inductor becomes continuous thus eliminating the distortion caused by the discontinuous current mode (DCM). Due to the doubled number of the filter inductors and the extra bias current through the filter inductors, the total filter inductor volume of OCC becomes much larger compared to the equivalent HB converter [8].

A volume reduction method by adding an extra bias filter inductor L_b , which is referred to the ELOCC, is shown in Fig. 1. The dc bias current mainly goes through the bias inductor L_b , which decreases the peak current and rms current of the filter inductors thus reducing the total inductor volume. The waveform of the bias inductor, filter inductor and output current of ELOCC is shown in Fig. 2.

Fig. 1: The extra bias inductor opposed current converter (ELOCC) with output filter.

Fig. 2: The waveform of filter inductor and output current of OCC topology.

The area-product method is used to estimate the volume of the inductor as described in [10] and defined

by

$$A_P = \frac{L}{\hat{B}JK_u} \hat{I}, \quad (1)$$

where L is the inductance, $\hat{B}$ is the peak magnetic flux density, J is the rms current density, K_u is the window utilization, $\hat{I}$ is the peak current and I is the rms current. To make volume comparison for the same core material, $\hat{B}$, J and K_u are assumed to be equal for all inductors. The volume of the inductor is estimated by

$$V_L = K_{\text{vol}} A_P^{0.75}, \quad (2)$$

where K_{vol} is the core's geometrical constant and can be regarded as a constant for the inductors discussed in this section. The filter current ripple Δi_{Lf} during each switching cycle is given by

$$\Delta i_{Lf} = \frac{V_{\text{dc}}}{L_f} \frac{1}{8} T_{\text{sw}} (1 - m^2) = k_r \hat{i}_{\text{out}} (1 - m^2), \quad (3)$$

where T_{sw} is the switching period, m is the modulation index, $\hat{i}_{\text{out}}$ is the maximum output current and k_r is the filter current ripple ratio. For a sinusoidal output, the ripple current is variable in each switching cycle therefore the rms value of the ripple current is also variable. The rms ripple current during a full output period is governed by

$$I_{\text{rms,ripple}} = \hat{i}_{\text{out}} \sqrt{\frac{1}{3} k_r^2 \left(\frac{3}{8} M_m^4 - M_m^2 + 1 \right)}, \quad (4)$$

where M_m is the modulation depth and assumed as 0.9 in this research. The calculated rms ripple current is about 66% of the estimation in [9] with a fixed ripple current assumption. Therefore, a smaller filter inductor volume is expected. The area-product of the each filter inductor is calculated as

$$A_{P,L_f} = \frac{V_{\text{dc}} T_{\text{sw}} \hat{i}_{\text{out}}}{8 \hat{B} J K_u k_r} \left[\frac{1}{2} + k_r + Q_{Lf} \left(\frac{1}{2} + k_0 \right) \right] \sqrt{\frac{1}{8} + Q_{Lf}^2 \left(\frac{1}{2} + k_0 \right)^2 + \frac{1}{3} k_r^2 \left(\frac{3}{8} M_m^4 - M_m^2 + 1 \right)}, \quad (5)$$

where k_0 is the offset current coefficient factor and Q_{Lf} is one of the current distribution coefficients and defined by

$$Q_{Lf} = \max \left(\frac{R_{Lb}}{R_{Lb} + 2R_{Lf}}, \frac{L_b}{L_b + 2L_f} \right), \quad (6)$$

where R_{Lb} and R_{Lf} are the parasitic series resistance of the bias inductor and filter inductor respectively. The area-product of the bias inductor is given by

$$A_{P,L_b} = \frac{V_{\text{dc}} T_{\text{sw}} k_{Lb} \hat{i}_{\text{out}}}{8 \hat{B} J K_u} \left[Q_{Lb} \left(\frac{1}{2} + k_0 \right) + \frac{m_{\text{bias}} k_r}{k_{Lb}} (1 + \hat{m}'_{\text{out}}) \right] \times \sqrt{Q_{Lb}^2 \left(\frac{1}{2} + k_0 \right)^2 + \frac{1}{3} \left[1 + \frac{1}{2} \hat{m}_{\text{out}}'^2 (1 + m_{\text{bias}}) \right] \left(\frac{m_{\text{bias}} k_r}{k_{Lb}} \right)^2}, \quad (7)$$

where k_{Lb} is the value of L_b relative to L_f , m_{bias} is the bias voltage modulation index, $\hat{m}'_{\text{out}}$ is the maximum value of the relative modulation index and is equivalent to

$$\hat{m}'_{\text{out}} = \frac{M_m}{1 - \frac{1}{2} m_{\text{bias}}} \quad (8)$$

and Q_{Lb} is the other current distribution coefficient expressed as

$$Q_{Lb} = \max \left(\frac{2R_{Lf}}{R_{Lb} + 2R_{Lf}}, \frac{2L_f}{L_b + 2L_f} \right). \quad (9)$$

The total inductor volume is estimated as

$$V_{L,EL} = K_{vol} \left(2A_{PLf,EL}^{0.75} + A_{PLb}^{0.75} \right). \quad (10)$$

For an equivalent HB, the area-product is governed by

$$A_{P,LHB} = \frac{V_{dc} T_{sw}}{16\hat{B}JK_u} \hat{i}_{out} \frac{1+2k_r}{k_r} \sqrt{\frac{1}{2} + \frac{4}{3}k_r^2 M_m}. \quad (11)$$

The volume of the filter inductor in a HB is estimated as

$$V_{L,HB} = K_{vol} A_{P,LHB}^{0.75}. \quad (12)$$

The total inductor volume is highly dependent on the current distribution coefficients Q_{Lb} and Q_{Lf} . For different k_r and k_o , there are different optimized value of Q_{Lb} and Q_{Lf} . To make a fair comparison to [9], the relative total inductor volume compared to the equivalent HB is calculated and the result is shown in Fig. 3. $V_{L,HB}$ is the inductor volume of HB with $k_r = 0.2$ and used as a reference voltage for comparison. The minimum total inductor volume is achieved when $k_{Lb} = 0.0147$, which is about a factor of 1.32 of an equivalent HB and is about 8% smaller than the expected value in [9]. A theoretical volume reduction of up to 66% can be achieved by using the ELOCC topology.

Fig. 3: The relative inductor volume of ELOCC to HB $V_{L,EL}/V_{L,HB}$ as a relation to k_{Lb} .

Fig. 4: The relative inductor volume of ELOCC as a function of k_r .

The total inductor volume can be optimized by k_r with a fixed k_{Lb} , which is depicted in Fig. 4. There is an optimized $k_r = 0.45$, which result in a factor of 0.82 of $V_{L,HB}$.

Prototype Design

A current amplifier prototype is built to verify the performance and functionality of the full-bridge equivalent ELOCC topology using wide bandgap semiconductors. The power-stage is depicted in Fig. 5. This current amplifier is designed for a high-precision motor drive application. The requirement of the prototype is depicted in Table I. The prototype power stage is built with GS66508T 650V enhancement mode GaN transistors and C3D10065E SiC Schottky diodes to achieve a switching frequency of 1 MHz.

Decoupled output and bias current controllers C_{out} and C_{bias} are used in the prototype. An overview of the control diagram is shown in Fig. 6. The output current is sampled at 500 kHz sampling frequency and the bias current is sampled at 1 MHz sampling frequency. All the controllers are updated at 2 MHz frequency and aligned with the asymmetrically sampled triangular carrier wave PWM.

Fig. 5: The power-stage of the prototype.

Fig. 6: ELOCC proposed control diagram.

Table I: The requirement of the current amplifier

Parameter	Value	Unit	Description	Parameter	Value	Unit	Description
V_{dc}	360	V	Dc voltage	f_{sw}	1	MHz	Switching frequency
$\hat{v}_{out}$	320	V	Output voltage	$\hat{i}_{out}$	12.5	A	Output current
L_x	2.5	mH	Load inductor	R	4	Ω	Load resistor

The inductor design is optimized according to the calculation in previous section. With the existing magnetic core, the filter inductor is selected as $36 \mu\text{H}$ and the current ripple k_r is selected as 0.2 for the full-bridge equivalent ELOCC considering the core loss and saturation effect. The bias inductor is selected as $6.8 \mu\text{H}$ considering the di/dt limitation. Therefore, k_{Lb} is 0.189, which results in a total inductor volume a factor of 1.61 of an equivalent HB converter theoretically. The total inductor volume is shown in Fig. 5, which matches with the theoretical estimation.

Simulation and Experimental Results

Simulated current waveforms with closed-loop configuration is shown in Fig. 7. The simulation model is based on the selected components and conducted in Simulink/PLECS. The parasitics of all the components, the on-resistance variance and inductor saturation effect are neglected. The output current magnitude $\hat{i}_{out}$ is 12.5 A_{pk} and the bias inductor current i_{Lb} is set as 5 A_{dc}. The frequency spectrum of the output current is shown in Fig. 8.

Fig. 7: The simulated waveforms of filter inductors, bias inductor, output current of OCC topology. A constant bias current i_{bias} is added to the two legs.

Fig. 8: The simulated frequency spectrum of the 12.5 A_{pk} 160 Hz sinusoidal output current at 360 V.

As can be seen, the bias current and the output current are fully decoupled in the closed-loop control. The waveform of filter inductor current i_{Lf1} and i_{Lf2} are both sinusoidal while i_{Lf1} is larger than i_{Lf2} because the switch node voltage u_{sn1} is larger than u_{sn2} in order to create a positive bias current. The expected relative magnitude (dB relative to the first order harmonic) of all the harmonic distortion is below -135 dB and the higher order harmonics are always lower.

Measurements are conducted to verify the design. The bias current of the P phase and the output current are measured with current probes which are connected to a Tektronix MSO54 oscilloscope. The measured waveforms are post-processed in MATLAB for better visualization. Waveforms of 12.5 A_{pk} output current and 5 A_{dc} bias current at 360 V dc supply voltage are shown in Fig. 9. Because of the relatively low sampling frequency compared to the frequency of the bias current, the bias current is not controlled well and coupled with the output current.

Fig. 9: Waveforms at 12.5 A_{pk} 160 Hz output current and 360 V dc supply voltage.

Fig. 10: The frequency spectrum of the 12.5 A_{pk} 160 Hz sinusoidal output current at 360 V.

The linearity of the current amplifier is tested by measuring the distortion harmonics in the output current under the same working condition. The output current is measured with a high linearity current sensor and analyzed in SR785 spectrum analyzer. A frequency spectrum of the output current is shown in Fig. 10.

The relative amplitude of the noise floor for this current amplifier is as low as -115 dB. The relative amplitude of the third order harmonic is -88 dB. The highest harmonic is the 15th order harmonic, the relative amplitude of which is -77 dB. Other harmonics have lower amplitude. The measured distortion is higher than the simulation because of all the non-linear factors, including PWM distortion, transistor on-resistance variation and inductor saturation effect.

The distortion of the existing product, which is a similar amplifier based on the FB topology and working at 187.5 kHz switching frequency, is -53 dB under the same output condition. Compared to the product, the prototype based on the ELOCC topology improves the third order harmonic rejection by 35 dB and gives a much better linearity. The bandwidth of the prototype is larger than 20 kHz tested with a 500 μ H load inductor and 4.7 Ω load resistor, which is already a significant improvement. The volume of the power stage is also decreased by more than 40% compared to the existing product.

Conclusion

In this paper, the total inductor volume of ELOCC topology is modeled with an improved accuracy and optimized by selecting the optimized current ripple ratio. GaN transistors are used to raise the switching frequency to 1 MHz in order to reduce the total inductor volume further. As a result, the volume of the power stage is reduced by 40% compared to the previous product. The bandwidth and the linearity are also improved. It is shown that the GaN transistors are effective in reducing the switching loss and promising to use in high-precision converters.

Further research could be focused on a better decoupled control of the bias current and output current in order to get better rejection for higher order harmonics.

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