Polyphase Boost Converter for Automotive and UPF Applications

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Abstract

Boost dc-to-dc converters have very good source interface properties. The input inductor makes the source current smooth and hence these converters provide very good EMI performance. On account of this good property, the boost converter is also the preferred converter for off-line UPF rectifiers. One of the issues of concern in these converters is the large size of the storage capacitor on the dc link. The boost converter suffers from the disadvantage of discontinuous current injected to the load. The size of the capacitor is therefore large. Further, the ripple current in the capacitor is as much as the load current; hence the ESR specification of the tank capacitor is quite demanding. This is specially so in the emerging application areas of automotive power conversion, where the input voltage is low (typically 12V) and large voltage boost (4 to 5) are desired. In the UPF rectifier applications, the input voltage varies from zero to maximum value twice in every cycle of the ac input voltage. The duty cycle therefore varies in the full range of zero to one. The inductor current varies from zero to rated current twice in every ac cycle of the input current. On account of these wide operating point variations, the design of the power circuit as well as the closed loop controller is a demanding task. This paper suggests polyphase operation of boost converter to overcome the disadvantages of large size storage capacitor in boost converter and off-line UPF rectifiers and a small signal analysis of N converters in parallel to an equivalent second order system in such converters, which have not been considered in previous works.

I. INTRODUCTION

In designing DC converters, parameters such as ratio of energy stored in inductor and capacitor to energy delivered to load in one period, maximum current in the switch and the value of the rms current in the output capacitor have great importance and it is necessary to be considered.

The motivation for this work is expressed through consideration of the above parameters in per unit measures for the two basic converters namely the buck and the boost converter [1]. Consider the boost converter in Fig.1 with per unit values defined as

$$V_{dc}=1,\ D=0.5,\ T_s=1,\ E_o=1,\ P_o=1,\ \Delta I_L/I_L=20\%,\ \Delta V_o/V_o=1\%$$

Table I gives the reactive elements and their energy storage capacity. From the table it is obvious that the boost converter requires total energy storage far in excess of buck converter.

Table. I: Comparing Buck & Boost Converters in terms of per unit values

	L (per unit)	C (per unit)	V_o (per unit)	E_L/P_oT_s	E_c/P_oT_s
Boost Converter	2.5	12.5	2	1.25	25
Buck Converter	1.25	10	0.5	1.25	1.25

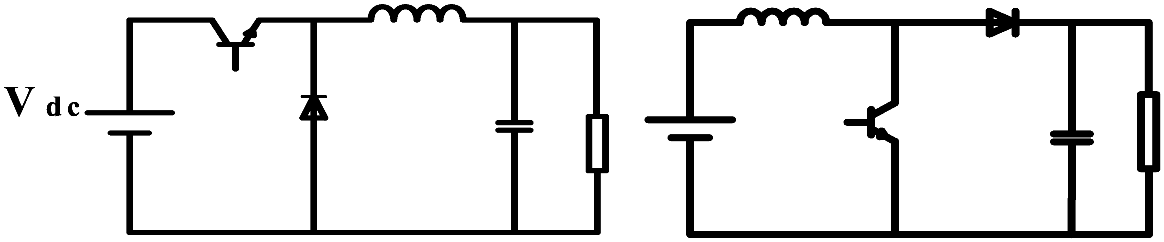

Fig. 1: Buck & Boost Converter respectively

One-way of reducing the storage requirement is increasing the switching frequency however this is not practicable in all instances. During the on state of the switch, the capacitor has to supply the load current in the boost converter and this discontinuity of current in the capacitor increases the rms value of current and also increases the amount of capacitor which is needed for correct operation of the circuit and therefore it results in more dissipation due to ESR of capacitor. In standard designs it is not uncommon to see tank capacitors one or two orders of magnitude higher than the ideally required capacitance A way to overcome this problem is using polyphase operation with appropriate phase shift in the control circuit of main switches [2,3]. This is done in such a way that at anytime one of the inductors is supplying the load current. The frequency of ripple current in the output capacitor is n times compared to the single stage and therefore the value of the capacitor required can be reduced.

Extracting continuous current from the input for applications such as unity power factor is usually the reason for choosing boost converter configuration. An advantage of using polyphase converter is in light load operation. In a single stage converter, the current will go to DCM operation while in this configuration one can switch off one, two or three switches according to the level of current so that CCM operation is extended. In section II, main points of analyzing small signal averaging of polyphase boost converter are presented. A comparison between the specifications of a single and polyphase boost converter is performed in section III. Section IV deals with experimental results and conclusions are summarized in section V.

II. Small signal averaging

The polyphase boost converter for N equals to four has been shown in Fig. 2. Each converter operates in continuous current mode (CCM) and these converters are controlled in current control mode through a series of synchronizing pulses, which are equally phase shifted in time [4,5].

If we have N converters in parallel then the total states in each cycle are 2N subintervals and the waveforms of input ripple current, output capacitor ripple current are at frequency which is N times the frequency of a single boost converter. If T_s denotes the period of switching frequency and T'_s denotes the period of input current then $T'_s = \frac{Ts}{N}$ (1)

On account of the above, the time interval each switch conducts is dependent on D and is given by

On Time $T_1 = \frac{T_s}{N}\{ND - floor(ND)\} = T'_s D'$ (2) Where $D' = ND - floor(ND)$ (3)

May be defined as duty ratio referred to T'_s and ***floor*( )** is the integer part function

And similarly Off Time $T_2 = \frac{T_s}{N}\{floor(ND+1) - ND\} = T'_s(1-D')$ (4)

Such that $T_1 + T_2 = T'_s$

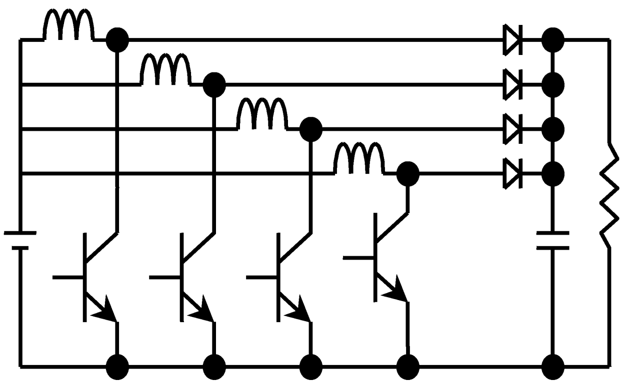

Fig. 2: Polyphase Boost Converters In Parallel

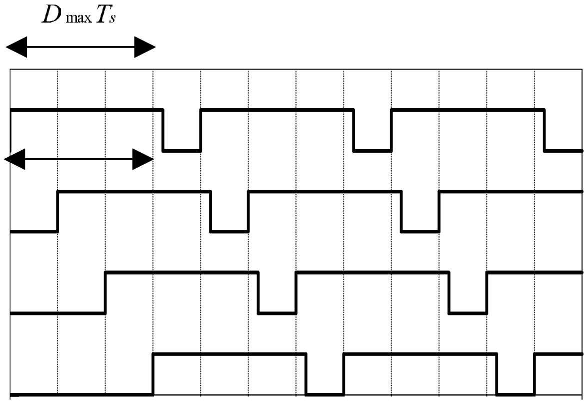

Fig. 3: The Case which $D_{\max} \ge 1 - \frac{1}{4}$

For example for N=4 and dependent on the value of D, table II gives the patterns of conduction of different switches. In practice as we have said before to prevent the output capacitor current to become large at least one inductor must supply the current to the load at any instant. From Fig.3 the phase shift of the last gate pulse with respect to the first one is $\frac{(N-1)}{N}T_s$

And this must be greater than $(T_{on})_{\max} = D_{\max} T_s$

i.e. $N \ge \frac{1}{1 - D_{\max}}$ or $D_{\max} \le 1 - \frac{1}{N}$ (5)

This gives a useful lower bound on the highest duty ratio that may be employed.

Table II: Different Possibilities of On & Off Situations of Switches for N=4

	Intervals	Switch	$0 \le D \le \frac{1}{4}$ (m = 4)	$\frac{1}{4} \le D \le \frac{1}{2}$ (m = 3)	$\frac{1}{2} \le D \le \frac{3}{4}$ (m= 2)	$\frac{3}{4} \le D \le 1$ (m= 1)
1	$0 \le t \le \frac{Ts}{4}$	ON	1 0 0 0	1 0 0 1	1 0 1 1	1 1 1 1
2		OFF	0 0 0 0	1 0 0 0	1 0 0 1	1 0 1 1
3	$\frac{Ts}{4} \le t \le \frac{Ts}{2}$	ON	0 1 0 0	1 1 0 0	1 1 0 1	1 1 1 1
4		OFF	0 0 0 0	0 1 0 0	1 1 0 0	1 1 0 1
5	$\frac{Ts}{2} \le t \le \frac{3Ts}{4}$	ON	0 0 1 0	0 1 1 0	1 1 1 0	1 1 1 1
6		OFF	0 0 0 0	0 0 1 0	0 1 1 0	1 1 1 0
7	$\frac{3Ts}{4} \le t \le Ts$	ON	0 0 0 1	0 0 1 1	0 1 1 1	1 1 1 1
8		OFF	0 0 0 0	0 0 0 1	0 0 1 1	0 1 1 1

Table II shows that D > 0.75 results in full ripple on the output capacitor (All four switches are on simultaneously). In writing dynamical equations and averaging function for $\dot{X} = AX + BU$ & N=4 If A_1, $A_2 \dots A_8$ represent system matrices correspondingly to above states then after averaging we will have: $\dot{X} = A_{av}X + B_{av}U$ Where

$$A_{av} = \frac{1}{N}\{D'\sum_{k=0} A_{2k+1} + (1-D')\sum_{k=1} A_{2k}\} \qquad (6) \qquad \text{And} \qquad B_{av} = \frac{1}{N}\{D'\sum_{k=0} B_{2k+1} + (1-D')\sum_{k=1} B_{2k}\} \qquad (7)$$

Whenever the **ith** inductor is connected to source the system matrix will be A_{2i+1} and similarly whenever the **ith** inductor is connected to load the corresponding system matrix will be A_{2i}.

For example for $N = 4$ & $0.5 \le D \le 0.75$
The A_{av} is:

$$\begin{bmatrix}
-\frac{4rl + (2-D')(rc \| R)}{4L1} & -\frac{(1-D')(rc \| R)}{4L1} & 0 & -\frac{(1-D')(rc \| R)}{4L1} & -\frac{(2-D')R}{4L1(R+rc)} \\
-\frac{(1-D')(rc \| R)}{4L2} & -\frac{4rl + (2-D')(rc \| R)}{4L2} & -\frac{(1-D')(rc \| R)}{4L2} & 0 & -\frac{(2-D')R}{4L2(R+rc)} \\
0 & -\frac{(1-D')(rc \| R)}{4L3} & -\frac{4rl + (2-D')(rc \| R)}{4L3} & -\frac{(1-D')(rc \| R)}{4L3} & -\frac{(2-D')R}{4L3(R+rc)} \\
-\frac{(1-D')(rc \| R)}{4L4} & 0 & -\frac{(1-D')(rc \| R)}{4L4} & -\frac{4rl + (2-D')(rc \| R)}{4L4} & -\frac{(2-D')R}{4L4(R+rc)} \\
\frac{(2-D')R}{4C(R+rc)} & \frac{(2-D')R}{4C(R+rc)} & \frac{(2-D')R}{4C(R+rc)} & \frac{(2-D')R}{4C(R+rc)} & -\frac{1}{C(R+rc)}
\end{bmatrix}$$

The above relationship is quite messy. We may simplify the dynamic model by defining the average inductor current as an equivalent state $i = (i_{L1} + i_{L2} + i_{L3} + i_{L4})$ (or input current) and V_C=V_C, the above matrix is simplified as

$$\begin{bmatrix}
-\frac{r_L}{L} - \frac{(4-3d')(R \| r_c)}{4L} & -\frac{(2-d')R}{4(R+r_c)L} \\
\frac{(2-d')R}{(R+r_c)C} & -\frac{1}{(R+r_c)C}
\end{bmatrix}$$

Provided that $L_1 = L_2 = L_3 = L_4 = L$ and $r_{L1} = r_{L2} = r_{L3} = r_{L4} = r_L$.

The simplifying rule is very easy and can be expressed as follows.Let us consider the matrix of a single stage converter with no non-idealities as $\begin{bmatrix} 0 & -\frac{1-d}{L} \\ \frac{1-d}{C^*} & -\frac{1}{RC} \end{bmatrix}$

Now using the relation (3) and substituting $\frac{C}{N}$ for C^* the equivalent matrix for N converters in parallel is obtained .It means that we have split the capacitor in Fig.2 to N parallel capacitor such that one of the converters is considered equivalently for average analysis. In case of non-idealities and in general case the above matrix is given by $\begin{bmatrix} -\frac{r_L}{L}-\frac{(m^2-(2m-1)d')(R \| r_c)}{\mathrm{NL}} & -\frac{(m-d')R}{N(R+r_c)L} \\ \frac{(m-d')R}{(R+r_c)C} & -\frac{1}{(R+r_c)C} \end{bmatrix}$ Where m=N,…,2,1 as indicated in table II.

III. Comparing Specifications

A. Inductor Size

By comparing two converters at the same condition i.e. same output power, same output voltage and same amount of ripple current in the inductors it is easy to show that for single stage

$$I_L = \frac{I_{load}}{1-D} \tag{8}$$

And for N converters in parallel

$$I_L = \frac{I_{load}}{N(1-D)} \tag{9}$$

Hence for same value of $\frac{\Delta I_L}{I_L}$ it is required that

$$L_n = NL \tag{10}$$

As the product of $A_c A_w$ remains same value in both single stage and N-parallel connected boost converters it means in practice the total size of N inductors are almost equal to the size of single stage inductor.

B. RHP

From the dynamical equation one can derive output to control transfer function. The results show that the RHP does not change because in the single stage the RHP equals to

$$\frac{V_g}{LI_L} \tag{11}$$

And for N converters in parallel it equals to

$$\frac{Vg}{(NL)(\frac{I_L}{N})} \tag{12}$$

From the dynamical equation and with the aid of matrices in relations (6), (7) the equality of the above result can be numerically justified.

C. Output Resistance

Again from the dynamical equation the output impedance or incase of output DC current the output resistance of the N parallel converter comes down substantially so that for N=4, $L_1 = L_2 = L_3 = L_4 = 50\mu H \quad C = 1000\mu F$, $r_c = 0.02$, $r_l = 0.01$, $V_0 = 32$ and $D = 0.5$ $\quad \dfrac{Ro_4}{R_{o1}} = 0.1244$

The curve for the different values of D has been sketched in Fig.4
Where $Rbased = Ro1$ (Output resistance of single stage)

D. RMS Current of Capacitor

One of the interesting point is the rms current of the output capacitor. Fig. 5 shows the waveform of the current. $Iload = 10, \Delta IL = 20\%, D = 0.7$ With a good approximation the rms current of the output Capacitor can be expressed by

$$I_{rms} = I_L\sqrt{(ND - K + 1)(K - ND)} \qquad (13)$$

Where $K = floor(ND) + 1$ & $I_L = \dfrac{I_{load}}{N(1-D)}$ and "$floor$" shows the integer part function

Or equivalently it can be written as

$$I_{rms} = I_L\sqrt{D'(1 - D')} \qquad (14)$$

Where $\quad D' = ND - floor(ND)$

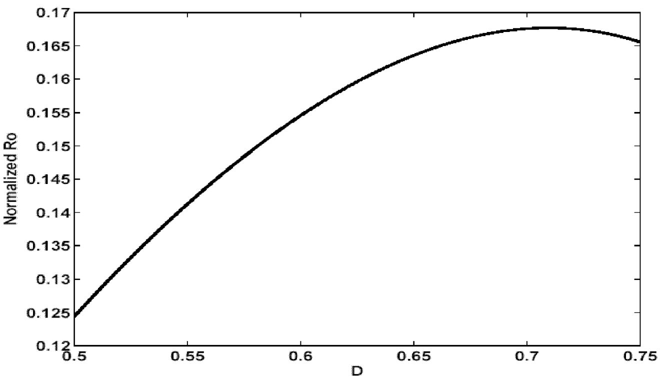

Fig. 4: The Normalized Output Resistance for N=4

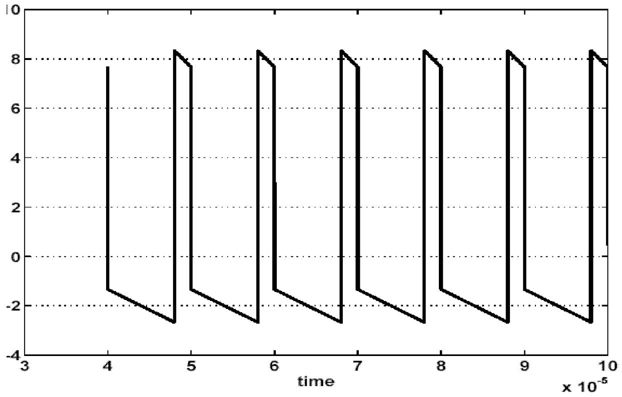

Fig.5: The Current of the Output Capacitor (D=0.7)

For $Iload = 16$, $Vg = 8$ Fig. 6 shows the rms current for different values of N. In Fig. 6 the converters are under similar conditions i.e. same output power; same output voltage and same amount of ripple current in the inductors. Fig. 7 shows the normalized rms capacitor current where the rms base current is the capacitor current in a single stage. It may be seen that the capacitor ripple current fall at least by a factor of N, but can be even smaller for certain duty ratios, such as 0.25,0.5,0.75, etc.

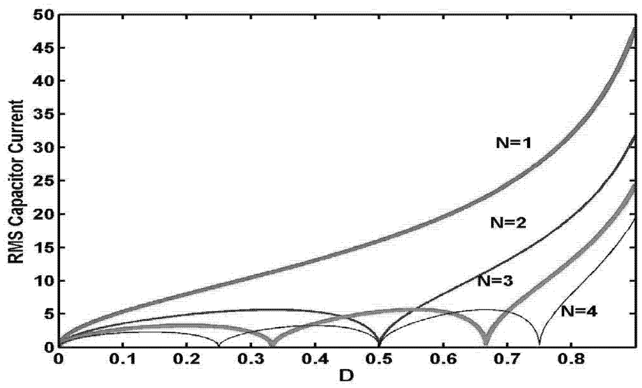

Fig. 6: The rms Capacitor Current for different N

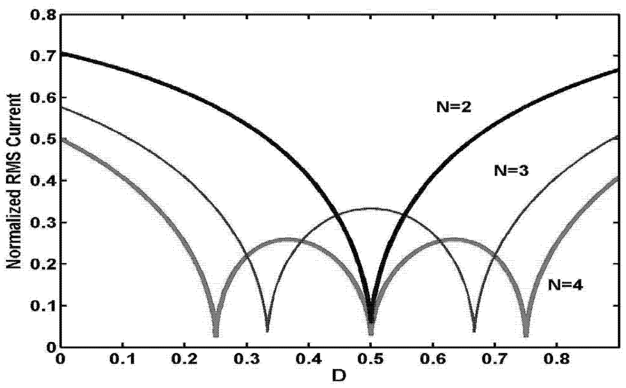

Fig. 7 : The Normalized rms Capacitor Current

E. Input Ripple Current

The input current is at higher frequency than the switching frequency and hence the EMI filter will be lighter and as it can be seen from Fig. 8 the peak to peak-input ripple current is substantially reduced and the magnitude of this current ripple for polyphase boost converter can be derived as [6]

$$\Delta Iin,N = \frac{(K-ND)(ND-K+1)}{ND(1-D)}\Delta Iin,1 \qquad (15) \qquad \text{Where, } K=1,2,\ldots,N\,, \quad \frac{K-1}{N}\le D\le\frac{K}{N}$$

Or equivalently

$$\Delta Iin,N = \frac{D'(1-D')}{ND(1-D)}\Delta Iin,1 \qquad (16)$$

And

$$\Delta Iin,1 = \frac{Vg}{L}DTs \qquad (17)$$

The normalized magnitude of this ripple as a function of D and for different values of N has been sketched in Fig. 9. Again the improvement is at least by a factor of N.

F. Output Ripple Voltage

Based on the previous analysis and the relation for output ripple voltage

that is $\frac{\Delta Vo}{Vo} = \frac{DTs}{RloadC}$ (18)

The output ripple voltage for N converters is given by $\left(\frac{\Delta Vo}{Vo}\right)_N = \frac{D'(1-D')}{N^2D(1-D)}\left(\frac{\Delta Vo}{Vo}\right)_1$ (19)

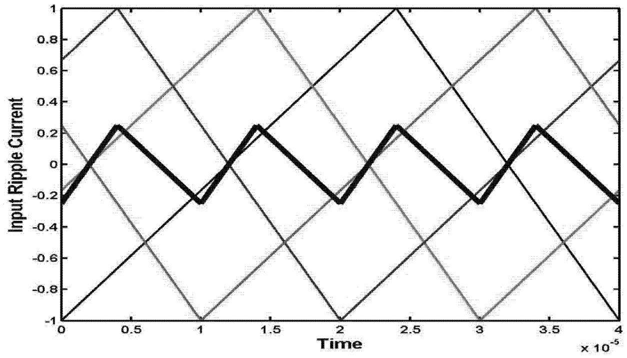

Fig. 8: Input ripple Current for N=4 &D= 0.6

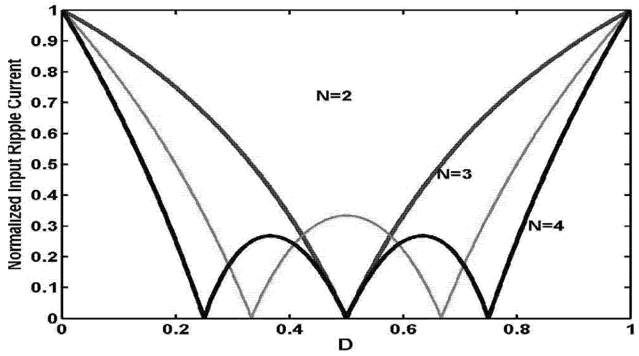

Fig.9: Normalized Magnitude of Input Current

Or in other words for the same amount of ripple in the output the size of capacitor reduces by a factor, which is equal to $\frac{D'(1-D')}{N^2 D(1-D)}$ and it has been sketched for different values of N and D in Fig. 10

IV. Experimental Results

A prototype with four boost converters in parallel has been tested. Each converter has a specification of 35W and they are operating under 90-degree phase shift in control circuit. The waveforms have been shown in Figs. 11,12 and 13. As it can be seen from Figs. 11 and 12 the rms output capacitor current for N=4 and N=1 are 0.463A and 1.803A respectively and the ratio is equal to 0.257 which matches with simulation results from Fig. 7 and similarly the ratio of input ripple current for N=4 and N=1 is equal to 0.297A/0.9A=0.33 which is almost matches with simulation results of Fig 9.

V. Conclusion

This paper discusses basic analysis method for polyphase boost converter. A novel method for defining the duty cycle corresponds to the N paralleled converters and an equivalent second order system for small signal averaging which are operating under phase shift, and current control have been developed. The size of N boost converters in parallel is almost same as a single boost converter of the same total power because the size of main parts-inductors-almost remains same.

Both simulations and measurements show that polyphase boost converter has several advantages over single boost converter. Smaller rms current in the energy-storage capacitor, lower input ripple current and lower output ripple voltage or smaller size of the tank capacitor are those important points, which have been considered.

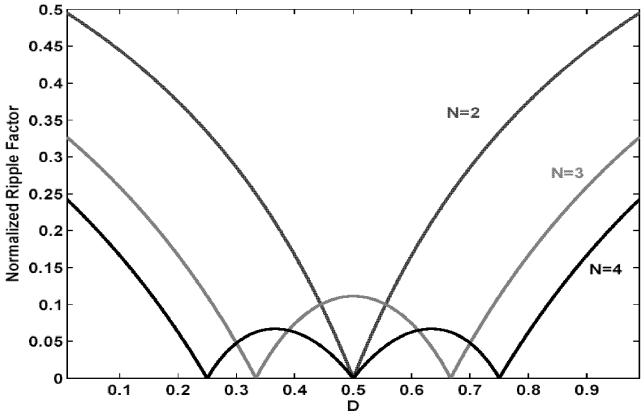

Fig. 10: Ripple Factor of Output voltage

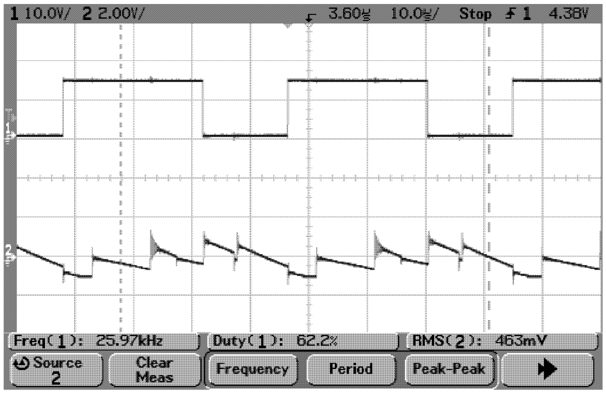

Fig. 11: The Output Capacitor Current for N=4

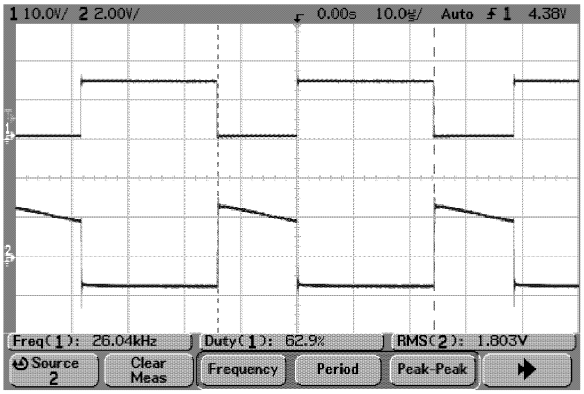

Fig. 12: The Output Capacitor Current for N=1

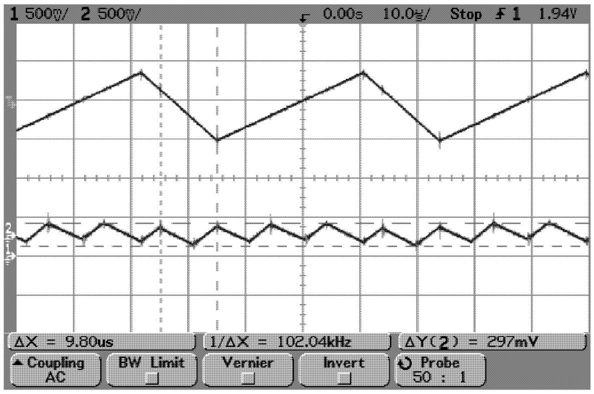

Fig. 13: Input ripple Current for N=4

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