A Resonant Local Power Supply with Turn off Snubbing Features

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***Abstract* - A local power supply circuit which is driven by the main switch of a PWM converter is described and analyzed. The operation of the circuit is based on the charge pump principle with a resonant reset of the charge pump capacitor. The charge pump capacitor also serves as a turn off snubber of the main switch. The analytical derivations were verified by experimental results.**

I. INTRODUCTION

Switch mode inverters and converters need a local power supply to feed the control circuitry and switch drivers. Many methods have been used in the past to obtain this auxiliary supply. They range from a stand alone local supply connected to main power line [1-3], to extra winding of transformers and inductors [4]. The consideration for choosing one approach over the others are numerous: cost, power level, the need for isolation, interfering noise and others. Clearly, there is no one optimal solution that fits all applications. Here we propose an additional approach that may have a merit in some applications. The proposed approach is a lossless charge pump built around the main switch. The method is similar to the one proposed earlier [5] which applies a hard switched capacitor charge pump. The present approach differs from the one described in [5] in several aspects. Among them is the soft switching that is obtained throughout, the lossless nature of the operation and the rather high power level that can be easily reached. This could be an advantage in systems that include DC fans that require substantial power. Furthermore, the proposed circuit also acts as a lossless turn-off snubber of the main switch and in some practical cases may be able to compete with known lossless turn-off snubbers [6-8].

II. THE TOPOLOGY

The basic topology of the proposed Local Power Supply (LPS) and its connection in a boost converter is shown in Fig. 1. Capacitor C_1 serves as a charge pump that delivers a fixed charge quanta each time the switch Q is turned off. The charge is transferred to the output side which includes a clamping Zener diode D_Z, a storage capacitor C_2 and the load, depicted as a resistor R_s. When the switch Q is turned on, capacitor C_1 is reset via the resonant inductor L. Excess energy of L, over what is required to reset C_1, is transferred to the output. It should be noted that the pump capacitor C_1 serves in fact as a lossless turn off snubber to the main switch. The larger the capacitor the better is the snubbing action. However, as detailed below, if the main load of LPS is the driver of the main MOSFET, C_1 will be rather small as compared to the parasitic capacitances of the transistor. If higher loads are expected the snubbing effect will be more significant.

III. ANALYSIS

Main assumptions:

1. Transistor Q and all diodes are ideal, but parasitic capacitances of Q are taken in account. It is assumed that these capacitances are linear.

2. Inductance of the input inductor of the converter L_{in}, capacitance of the output capacitor of the converter C_o and capacitance C_2 of LPS are infinitely high. Therefore the input current of the converter I_{in}, the output voltage of the converter V_o and the output voltage of LPS V_s do not include an ac component:

$$I_{in}=\text{const}, \qquad V_o=\text{const}, \qquad V_s=\text{const}$$

The modes of operation of the proposed LPS will be discussed in relation to the timing diagram obtained by PSPICE simulation (Fig. 2).

A. Time intervals

Interval t_0-t_1 begins at t_0 when the transistor Q is turned off. Equivalent circuit for this interval is given in Fig. 3 where C_{Qout} is the output capacitance of the transistor. At t_0 the voltage across C_{Qout} is zero and the voltage across the capacitor C_1 is $-V_s$ (because the voltage across C_2 is $+V_s$).

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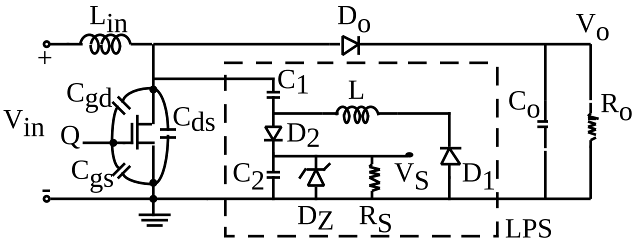

Fig. 1. Proposed Local Power Supply (LPS) connected in a boost converter: R_S - load resistance.

During the time interval t_0-t_1 capacitors C_{Qout} and C_1 are charging under action of the input current I_{in} flowing through the main inductor L_{in}. At t_1 the high terminals of the two capacitors reach V_o. The voltage of diode D_o reverses polarity and it turns on. As a result, the charging process of the capacitors stops and the current I_c of the circuit C_1-D_2 is interrupted. The voltage across the capacitor C_1 equals V_o-V_s at this instance. Duration of the charging interval $t_{0-1}=t_{ch}$ is found from the following equation:

$$t_{0-1} = t_{ch} = \frac{C_1 + C_{Qout}}{I_{in}} V_o \qquad (1)$$

Note that $t_{0-1}=t_{ch}$ is the minimum value of turn-off interval of the main switch ($t_{off\ min}$) required for proper operation of the LPS and the converter. If $t_{off}<t_{off\ min}$ the output diode D_o will not conduct.

During the time *interval t_1-t_2* (Fig. 4) there is no interconnection between the processes in the LPS and in the converter. This is true only under the above assumption that C_2 is infinitely large. For a finite value of C_2, a small current will flow into the LPS through C_1-D_2 due to the drop in V_s.

Interval t_2-t_3 (Fig. 5) begins at t_2 when the transistor Q is turned on. As a result a negative voltage -(v_{C1}-V_s) is applied to diode D_2 at t_2 that blocks its conduction. Diode D_1 turns on at the same instant under the action of the voltage across the capacitor C_1 ($v_{C1}=V_o-V_s$). This capacitor begins to discharge through the transistor Q, the diode D_1 and inductor L. The current of this resonant circuit is:

$$i_L = -i_{c1} = I_{Lm} \sin(\omega_r t) \qquad (2)$$

where

$$I_{Lm} = \frac{V_o - V_s}{Z_r}; \qquad Z_r = \sqrt{\frac{L}{C_1}}; \qquad \omega_r = \frac{1}{\sqrt{LC_1}}$$

The peak current through the transistor will be:

$$I_{Qm} = I_{in} + I_{Lm} = I_{in} + \frac{V_o - V_s}{Z_r} \qquad (3)$$

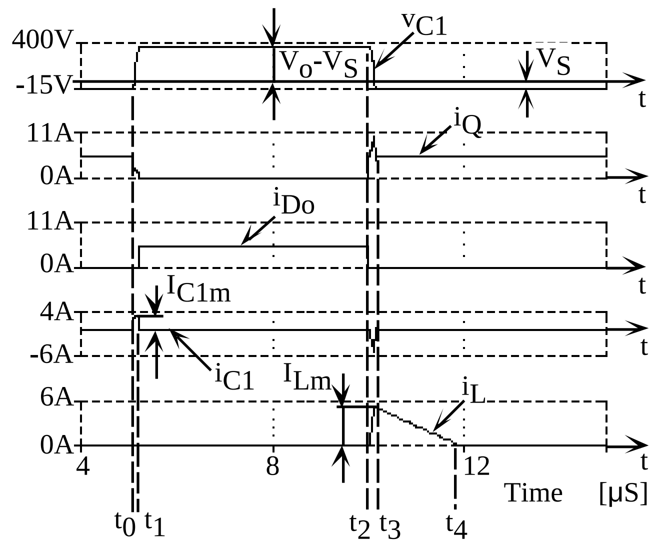

Fig. 2. Current and voltage waveforms.

The interval ends at t_3 when v_{C1} reaches -V_s and therefore the diode D_2 turns on. From this condition the duration of the interval t_{2-3} was found to be:

$$t_{2-3} = \frac{1}{\omega_r} \cos^{-1}\left(-\frac{V_s}{V_o - V_s}\right) \qquad (4)$$

For the case $V_s \ll V_o$ (4) can be approximated to:

$$t_{2-3} \approx \frac{\pi}{2}\sqrt{LC_1} \qquad (4a)$$

Note that t_{2-3} defines the required minimum value of turn-on interval of the main switch ($t_{on\ min}$). The minimum value of the duty cycle is thus found to be:

$$D_{min} = \frac{t_{onmin}}{T_s} \approx f_s \frac{\pi}{2}\sqrt{LC_1} \qquad (5)$$

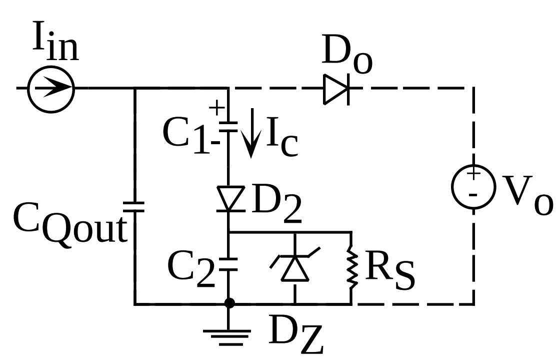

Fig. 3. Equivalent circuit for the t_0-t_1 time interval.

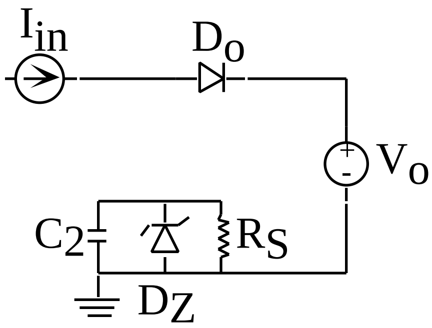

Fig. 4. Equivalent circuit for the t_1-t_2 time interval.

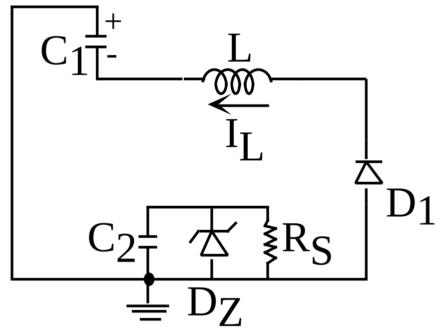

Fig. 5. Equivalent circuit for the t_2-t_3 time interval.

where T_s is the switching period and $f_S=1/T_s$ is the switching frequency.

Interval t_3-t_4 (Fig. 6) begins when the diode D_2 turns on. The current of the inductor L (i_L) flows now through diodes D_1 and D_2 and the parallel circuit D_Z, C_2, R_s. Initial condition of the inductor current is evaluated from (2) and (4). The interval ends when $i_L=0$.

Duration of the interval t_3-t_4 was found to be:

$$t_{3\text{-}4}=\frac{V_o - V_s}{\omega_r V_s}\sqrt{1-\frac{V_s^2}{(V_o - V_s)^2}} \tag{6}$$

For the case $V_s \ll V_o$ (6) can be approximated to:

$$t_{3-4} \approx \frac{V_o - V_s}{\omega_r V_s} \tag{6a}$$

Interval $t_{3\text{-}4}$ can continue during the off period. Hence approximately

$$t_{3-4} \approx T_s \tag{7}$$

B. Energy transfer

The energy E transferred to the power supply during one switching period and consumed by the load includes two components: E_1 and E_2

$$E = E_1 + E_2 \tag{8}$$

The component E_1 is the energy transferred directly into the parallel circuit C_2-D_Z-R_s during the interval t_0-t_1=t_{ch}

$$E_1 = V_S I_C t_{ch} \tag{9}$$

where I_C is the part of the input current of the converter which flows through C_1 during the interval t_0-t_1=t_{ch}:

$$I_C = I_{in}\frac{C_1}{C_1 + C_{Qout}} \tag{10}$$

Applying (1), (9) and (10) we obtain:

$$E_1 = C_1 V_s V_o \tag{11}$$

This relationship can also be derived directly by considering the total charge C_1V_o delivered to the output (V_s).

The component E_2 is the energy transferred at first into the capacitor C_1 (interval t_0-t_1=t_{ch}). Next, this energy is removed from the capacitor C_1 and put into the magnetic field of the inductor L (interval t_2-t_3). Then the energy is transferred into the parallel circuit C_2-D_Z-R_s (interval t_3-t_4)

$$E_2 = \frac{(V_o - V_s)^2 - V_s^2}{2}C_1 = \frac{V_o^2 - 2V_oV_s}{2}C_1 \tag{12}$$

From (11) and (12) we obtain:

$$E = \frac{C_1 V_o^2}{2} \tag{13}$$

The energy E injected into the power supply during one switching period T_S can also be described by following equation:

$$E = V_S(I_S + I_Z)T_S \tag{14}$$

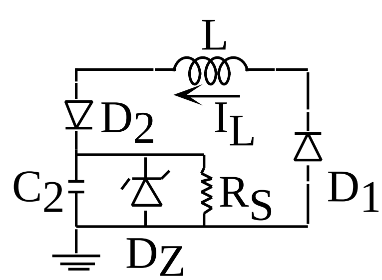

Fig. 6. Equivalent circuit for the t_3-t_4 time interval.

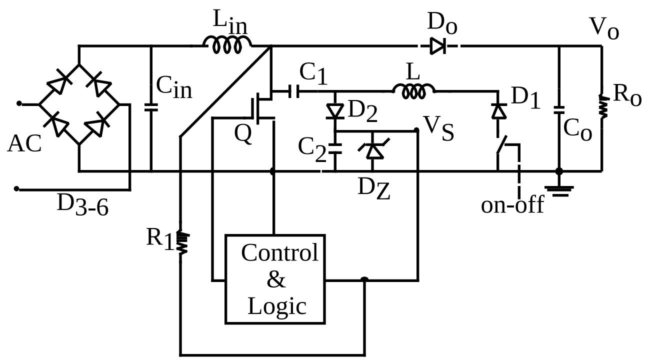

Fig. 7. Proposed local power supply connected in a boost converter and loaded by the MOSFET's controller/driver.

where I_s is the load current and I_Z is the current of the Zener diode.
From (13) and (14)

$$I_S + I_Z = f_S C_1 \frac{V_o^2}{2V_S} \tag{15}$$

The Zener diode current will thus be:

$$I_Z = f_S C_1 \frac{V_o^2}{2V_S} - I_S \tag{16}$$

The maximum load current is when $I_Z \approx 0$:

$$I_S = I_{S\ max} \approx f_S C_1 \frac{V_o^2}{2V_S} \tag{17}$$

C. Sizing C_1

We consider now the case when the main load of the LPS is the MOSFET's driver (Fig. 7). The average positive gate input current of the transistor is found from the following equation:

$$I_{g\ av} = C_{Qin} V_{gs} f_S \tag{18}$$

where V_{gs} is the gate source voltage of the transistor and

$$C_{Qin} = C_{gd}\left(1 + \frac{V_o}{V_{gs}}\right) + C_{gs} \tag{19}$$

C_{gd} and C_{gs} are gate-drain and gate-source capacitances of the transistor. It should be noted again that the above equations are under assumption that the transistor capacitances are linear behaviored.

We further assume that the gate current is a certain fraction of the power supply current:

$$I_{g\ av} = kI_{S\ max} \tag{20}$$

(k<1). Hence from (17):

$$C_{Qin} V_{gs} f_S = k f_S C_1 \frac{V_o^2}{2V_S} \tag{21}$$

Taking into account that $V_{gs}=V_S$, applying (19) and (21), the necessary value of the capacitance C_1 can be found:

$$C_1 = \frac{2}{k}\frac{V_{gs}}{V_o}\left[C_{gd}\left(\frac{V_{gs}}{V_o} + 1\right) + C_{gs}\frac{V_{gs}}{V_o}\right] \tag{22}$$

If $V_{gs} << V_o$

$$C_1 \approx \frac{2}{k}\frac{V_{gs}}{V_o} C_{gd} \tag{22a}$$

The two last equation imply that the value of C_1 is in the same order of magnitude as the parasitic capacitances of the transistor. In this case the contribution of C_1 to lower dv/dt might be insignificant. However, if additional power is required (e.g. for DC fans) C_1 will be larger and its contribution to turn off snubbing will be significant.

IV. EXPERIMENTAL RESULTS

The experimental boost converter with LPS (Fig. 1) had the following parameters: Q=IRFP460, D_o=MUR460, D_1=1N5819, D_2=MUR160, L_{in}=1mH, L=24.2µH, C_o=1mF, C_1=1.0 - 5.5nF, C_2=100µF, R_s=10 - 64 Ω. The experimental conditions were as follows: P_o=85W, V_s=15V, f_s=100kHz.

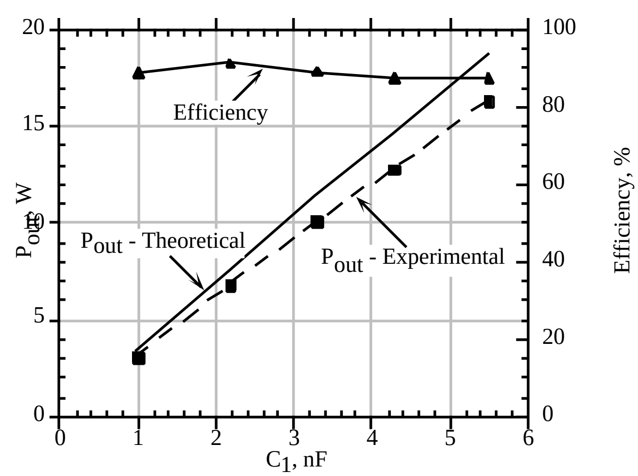

Fig. 8. Output power and efficiency as functions of the charge pump capacitance C_1.

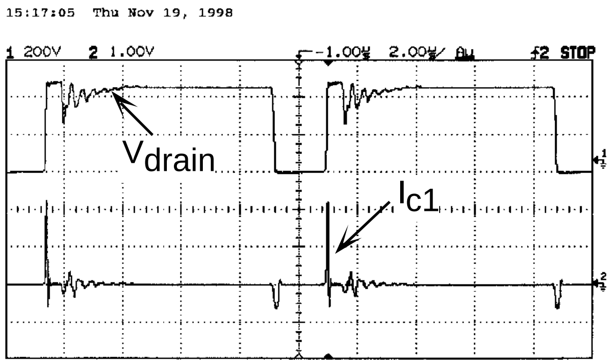

Fig. 9. Experimental waveforms of main switch (Q) drain voltage (upper) and charge pump capacitor (C_1) current (lower). Horizontal scale: 2μS/div. Vertical scales: 200V/div (upper) and 1A/div (lower).

The output power $P_{out} = V_s I_s$ of the experimental LPS was determinated for different values of capacitor C_1 with no Zener diode. The load resistance R_s was selected to obtain a constant V_s (=15V). Experimental results were compared with the power transmitted through the capacitor C_1. The latter was calculated by the equation:

$$P_{out} = f_S E = \frac{f_S C_1 V_o^2}{2} \tag{23}$$

The discrepancy between measured and calculated powers is mainly due to losses in the diodes and passive elements. Therefore the ratio between experimental and theoretical values of P_{out} can be considered as the efficiency of LPS (Fig. 8). The overall efficiency was found to approach 90%.

Experimental waveforms of the LPS operating in a soft switched Active Power Factor Correction circuit [9] in which $C_1 \approx 220pF$, $V_{in} = 220V_{rms}$, $V_o = 380V$, $V_s = 12.4V$ and $P_o \approx 1W$ are given in Fig. 9. The plots correspond to the peak of the ac input current.

V. DISCUSSION AND CONCLUSIONS

The main features of the proposed 'piggyback' local power supply are simplicity and high efficiency. This is correct as long as all the energy transferred through C_1 is actually consumed by the circuit. In reality, one will have to allow some bleeding through D_Z. However, in cases of an LPS with a large variable load (e.g. DC fans with speed control) considerable power can be wasted when the load is light. The remedy that can be proposed is an extra switch operating at low frequency (Fig. 7). This can be used to regulate the power level of the LPS. In this case, the snubbing action is active only when the extra switch is 'on'. The power that can be obtained from proposed LPS is rather high, limited only by the energy stored in the main inductor.

Like most local power supplies, the proposed LPS requires a start up circuitry. Before pulses can be supplied to the main switch there is a needed for an initial supply voltage to feed the PWM controller and driver. However, before switching begins, the LPS is inoperative and can not supply the auxiliary circuit.

This problem can be solved by any one of the methods applied in other local power supply designs. For example, a bypass resistor (R_1, Fig. 7) can be connected from the output (in case of a boost converter) to charge an electrolytic capacitor to the minimum operating voltage of the PWM controller/driver. Once the main transistor starts switching the bypass resistor could be disconnected.

Protection circuits (like overcurrent and overvoltage protection) that are normally present in each system will interrupt gate pulses in case of abnormal operation. With no main switch pulses the LPS voltage will drop and a recovery sequence will be required.

Aside from its prime function as a local power supply the circuit also serves as a turn off snubber. The effectiveness of the snubber increases with the power level of the local power supply when larger C_1 are required.

REFERENCES

[1] Power Integrations, Inc., *"Data book and design guide"*, 1996-97.

[2] B. Andreycak, "Unique "cheap and dirty" converter for low power bias supplies", *Unitrode Applications Handbook*, IC# 1051/1997, pp.4-14 - 4-15.

[3] B. Andreycak, "UCC3889 bias supply controller evaluation kit - schematic and list of materials", *Unitrode Applications Handbook*, IC# 1051/1997, pp. 4-68 - 4-69.

[4] B. Andreycak, "Optimizing performance in UC3854 power correction applications", *Unitrode Applications Handbook*, IC# 1051/1997, pp.4-16 - 4-20.

[5] B. Andreycak, "Inductorless bias supply design for synchronous rectification and highside drive applications", *Unitrode Applications Handbook*, IC# 1051/1997, pp.4-82 - 4-84.

[6] B. W. Williams, *"Power electronics. Devices, drivers, applications, and passive components"*, Second edition, McGraw-Hill, Inc., 1992.

[7] Ph. C. Todd, "Snubber circuits: theory, design and application", *Unitrode Corporation*, May 1993, pp. 2-1 - 2-17.

[8] L. D. Salazar, P. D. Ziogas, G. Joos, "On the minimization of switching losses in dc-dc boost converters", *Proceedings APEC '92*, pp. 703-708.

[9] H. Levy, I. Zafrani, G. Ivensky and S. Ben-Yaakov, "Analysis and evaluation of a lossless turn-on snubber", *Proceedings APEC '97*, pp 757-763.