

Article

Control Strategy Proposal for Modular Architecture of Power Supply Utilizing LCCT Converter

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Abstract: Following the invention of resonant power converters, lots of new topologies with significant improvements considering increase of efficiency and power density are arising. The main differences are related to the configuration of the resonant tank structure. In this paper, LCCT topology is proposed, while main aim is focused on the application of the modular architecture of power supply (MAPS) systems. Initially, principal analysis is given which describes basic features and components selection of LCCT resonant converter. After principal analysis, the application is oriented for above mentioned modular architecture of power supply, where more than one LCCT converter operated simultaneously is considered. The main asset within the presented system investigation is the proposal for the control strategy of the modular power supply system. Instead of the requirement on the most efficient operation within the whole operational power range of MAPS, the proposal for control strategy is given, while it is expected that each module of MAPS will be loaded evenly. The control strategy is based on the digital control, whereby sequential switching of the individual power stages is autonomous based on the information's of the actual power load. Presented solution gives improvements of the parameters where extended range of input voltage, extended range of output power, flat characteristic of efficiency, and lower ripple current and/or voltage will be required.

Keywords: resonant converter; efficiency; redundancy; reliability; control strategy

1. Introduction

The concept of resonant converters has greatly expanded into various industrial and consumer applications, such as power supplies for distributed systems, laptops, LCD televisions, aerospace, automotive, and telecom sectors as well. To reach the satisfactory electrical parameters and behavior of converter, it is necessary to utilize new concepts of its main circuit [1–6]. Focusing on the expectation of next days the perspective topologies must withstand crucial requirements on the power density, efficiency, reliability, and operational life for various specific or industrial applications. Instead of that, the quality of electrical variables (voltage, current) and extended regulation range are also required to be above standard normative [7–12]. Furthermore, very narrow tolerances are necessary even during load transient at the same time while guaranteeing high reliability and efficiency [13–19]. To satisfy load requirements, modular architecture of switched mode power supply and suitable control strategies must be used. As paralleled modules must share the same load current, the problem of detecting the load sharing between the modules arises [20–23]. The common approach is to use a resistive shunt, but this solution decreases efficiency, particularly in high-current low-voltage applications [24–27]. Different modular topologies have been explored in the past aimed mainly to reduce the need of having long strings of PV panels [28,29]. However, for the case of other applications, long stacks

of cells are not avoidable. Thus, a different converter structure, control scheme, and modified stack structure become necessary if for example automotive or telecom application is considered [30,31].

The main contribution of this paper is the presentation of control strategy proposal for modular architecture of power supplies, while the main aim is achievement of the even distribution of power delivery between individual modules. Opposed to the previous studies the proposal is simple, whereby it utilizes precise lossless current sensing (output values through hall sensors) and current control that is independent for individual modules. Voltage control is also independent and is ensured by the MASTER module, which can be any of the modular concept. Through this approach, where voltage control is superior oppose to current control, and thanks to independent frequency control of individual modules, even current sharing is secured, and frequency influencing is eliminated (beat frequencies etc.). Instead of that, the focus was oriented on the most valuable efficiency operation of the modular power supply. The advantages from the presented approach are strongly coupled to the reliability and lifetime issues of switched mode power supplies. Given proposal is practically implemented at fourth order DC–DC resonant converter with LCCT network within modular architecture of power supply, operated at parallel-input and parallel-output configuration. It is expected that three modules are operated simultaneously, while redundancy, overloading capability, short-circuit prevention, and no-load are expected applications [32,33].

The structure of the paper is as follows:

- LCCT converter topology and calculation of circuit components for proposed application
- Transfer characteristic of proposed converter and description of operational regions
- Modular architecture and possible configurations for the use within telecom data centers
- Parallel cooperation of converters without sequential switching and autonomous power sharing

Implementation of control strategy master-voltage and current-slave regulator within modular architecture

2. LCCT Converter for Modular Architecture of Power Supply

Based on the study of several papers, and based on the industrial standards, which are already known, the definition of nominal parameters for LCCT converter (Figure 1), which may be used within modular architecture of power supply, is defined first.

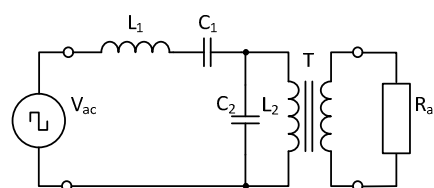


Figure 1. Simplified equivalent circuit of resonant tank for LCCT converter.

Industrial standard for distributed systems is defined with 400 Vdc of bus voltage, and with 48 Vdc of output voltage. For these reasons the definition of the basic input–output parameters of proposed LCCT converter are as follows:

- Input voltage = 400 Vdc
- Output voltage = 48 Vdc
- Output power = 500 W
- Switching frequency = 100 kHz

Based on the nominal input–output parameters, it is possible to estimate main circuit components, i.e., resonant inductors and capacitors (Figure 1). Calculation of resonant components for proposed converter is given within simplified procedure by Equations (1)–(6). The mathematical model of LCCT

converter that is derived from LCLC converter is given in Appendix A. General formula for nominal input impedance of resonant converter is given by (1).

$$L \cdot \omega_{res} = \frac{1}{\omega_{res} \cdot C} = |Z_N|; |Z_N| = \frac{V_{outN}^2}{P_{outN}} \quad (1)$$

where V_{outN} is nominal output voltage at the output of resonant tank, P_{outN} —is nominal output power of converter, L is resonant inductor, C is resonant capacitor and ω_{res} —is resonant frequency.

Definition of quality factor of resonant circuit under nominal load is given by (2).

$$q_N = \frac{L \cdot \omega_{res}}{|Z_N|} = \frac{1}{\omega_{res} \cdot C \cdot |Z_N|} \quad (2)$$

We can calculate resonant element values by using Equations (3)–(6).

$$L_1 = \frac{V_{outN}^2 \cdot q_N}{\omega_{res} \cdot P_{outN}} \quad (3)$$

$$L_2 = \frac{V_{outN}^2}{\omega_{res} \cdot P_{outN} \cdot q_N} \quad (4)$$

$$C_1 = \frac{P_{outN}}{\omega_{res} \cdot V_{outN}^2 \cdot q_N} \quad (5)$$

$$C_2 = \frac{P_{outN} \cdot q_N}{\omega_{res} \cdot V_{outN}^2} \quad (6)$$

The previous simple analysis is valid for LCLC converter [34], while the analogy between those equations for computation of resonant elements for LCCT converter is valid.

For previously given nominal input–output parameters the computation of main circuit components was provided. Resonant components are calculated regarding the fact that nominal output voltage at the output of resonant tank is $V_{outN} = 200$ Vdc. This voltage is transformed by high frequency transformer to the required output voltage value, which for target application shall be 50 Vdc. Based on that, the voltage transfer ratio of transformer is $n = 4$. Nominal power of one converter module was defined to $P_{outN} = 500$ W, what is relevant for industrial standards. Resonant frequency of resonant tank is $f_{res} = 100$ kHz and nominal quality factor is $q_N = 2$. The value of q_N was selected according to the requirement related with the steeper character of transfer characteristic of proposed resonant tank. With the use of (1)–(6) following values of components: $L_1 = 254.78$ μ H, $L_2 = 63.69$ μ H, $C_1 = 9.95$ nF, $C_2 = 39.8$ nF were calculated, where L_2 shall be considered as leakage inductance of the primary side of transformer.

Within the design of the transformer core, RM12 shape is selected. Table 1 summarizes basic parameters of the transformer.

Table 1. Basic transformer parameters.

Component	Value	Units
Primary inductance	254.78	μ H
Leakage inductance	63.69	μ H
Air gap	168	μ m
Number of primary turns N1	8	-
Number of secondary turns N2	4	-

3. Transfer Characteristic of Proposed LCCT Converter and Its Operational Regions

AC sweep analysis of the proposed resonant tank (LCCT) was performed for the purposes of operational regions identification through investigation of voltage-gain characteristic (Figure 2).

It can be seen, that resonant point is located at main resonant frequency $f_{res} = 100$ kHz. The gain characteristics is designed for nominal load (100% of P_{out}), overload (200% P_{out}) and for light-load (20% P_{out}) condition. From Figure 2 can be deduced that load change will not have any effect on the value of output voltage when switching frequency is equal to f_{res} , because voltage gain for resonant point is equal to unity for any value of the load. This ability is comparable to well-known LLC converter, and such property acts as advantage from regulation point of view [35]. Instead of that, shape of characteristic is mirrored, thus high gain can be achieved for loads below 100 % in both areas (above and below resonance).

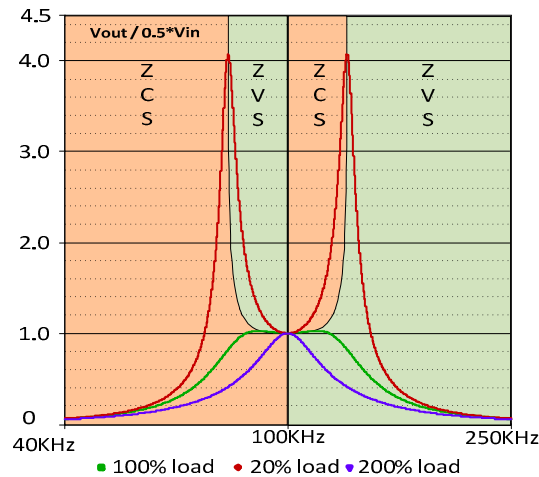


Figure 2. Voltage gain characteristic with operating regions.

Another advantage is that ZVS (Zero Voltage Switching) regions, which are preferred for operation, are visible both on left and right side from resonant frequency. Basically, it can be said that LCCT topology of resonant circuit gives possibility for wider regulation of output variables, almost because of the shape of transfer characteristic and ZVS region that is presented through the wide range of switching frequency.

Furthermore, for verification of transfer specifics of proposed LCCT converter, two variations were investigated (Figure 3). The first one has transformer voltage ratio equal to one, second has voltage ratio 1:0.25 what is required to achieve $V_{out} = 50$ Vdc within nominal control point of resonant converter.

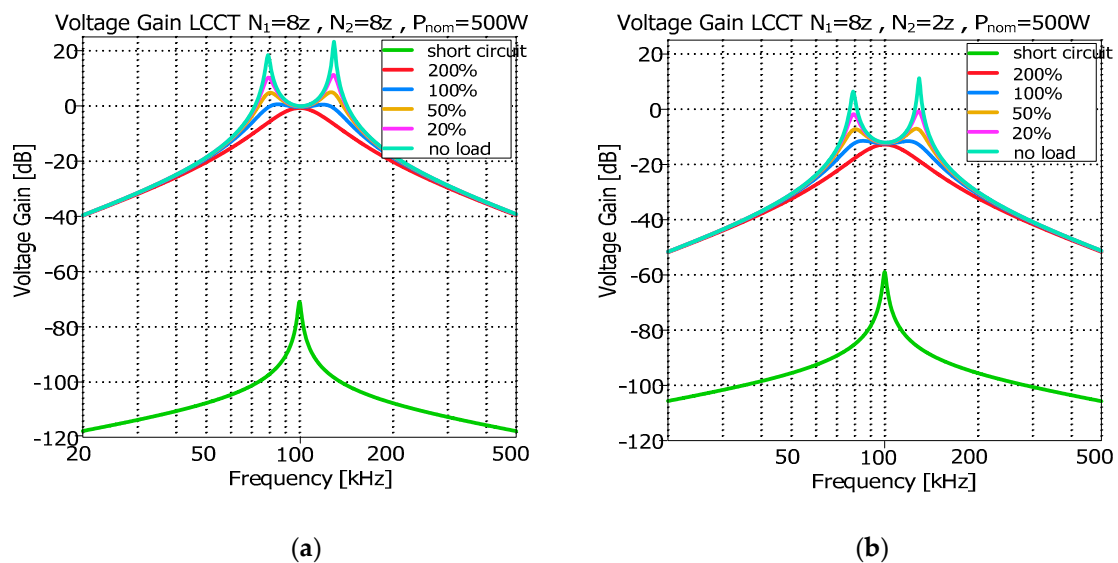


Figure 3. Impact of turn ratio on the voltage gain of resonant tank, (a) ratio 1:1, (b) ratio 1:0.25.

Shape of voltage transfer characteristic is similar comparing both situations, near the same if we exclude trace for state of short circuit (Figure 3). The only difference is naturally the value of the voltage gain, which is proportional to voltage ratio of transformer. For resonant point of operation, the value of voltage gain for second alternative is -12.4314 dB, what means that the voltage on secondary winding of transformer is given by Equation (7). This is relevant for the required value of the output voltage which shall be 50 V.

$$V_{N2} = \frac{V_{in}}{2} * 10^{\frac{-12.4314}{20}} = \frac{V_{in}}{2} * 0.239 \quad (7)$$

On the other side, visible difference can be seen for critical mode of operation—short circuit of the output. It is seen, that converter, whose voltage ratio is 1:0.25 does not have as high attenuation as is for ratio 1:1. Based on this fact, it can be said that the higher the voltage ratio is, the worse the short-circuit abilities converter has.

Based on previous analysis, several investigations regarding steady-state operation as well as dynamic properties identification have been realized within selected operating conditions and operating regions [36]. As was found, the proposed converter has expected abilities i.e., overload possibilities, short-circuit immunity, and fast dynamic response as well as wide operation range considering input voltage. Those properties are valuable within modular architectures of the power supplies where high level of redundancy is required.

4. Modular Architecture of Power Supply

At this point the use of LCCT converter within the architecture of modular power supply is described. As was already mentioned, configuration of modular power supply is suitable for various applications. Within this paper, we have focused on the use for telecom data storage servers (Figure 4a). Several specifications are given for selected application and are related with input supply voltage, output voltage and power of individual modules, while efficiency must relate to given industrial standards.

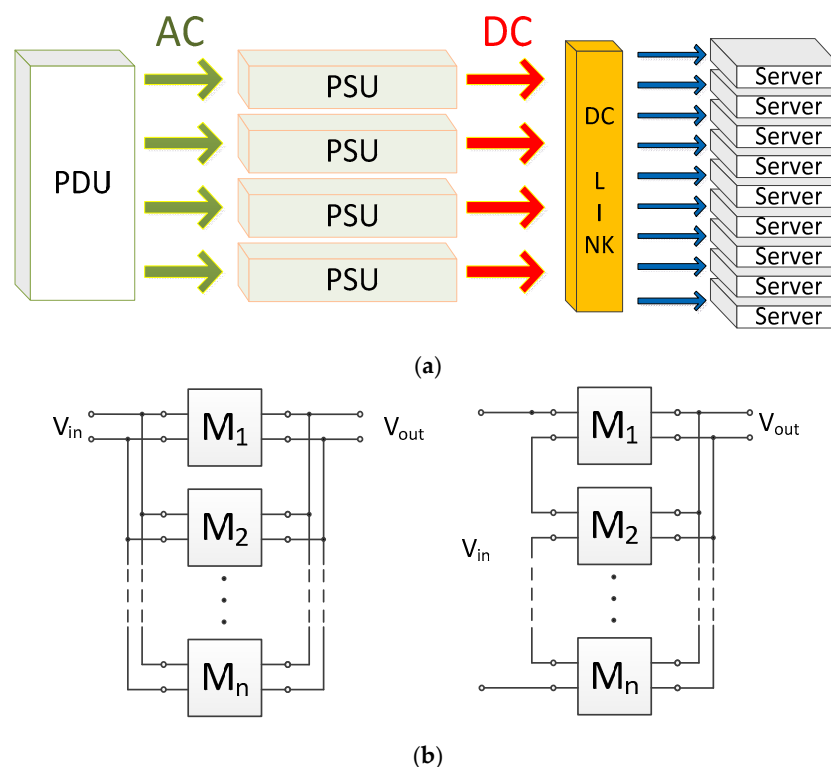


Figure 4. Modular architecture of power supply unit (a) and possible configurations (b).

For given configuration (Figure 4a) several input–output configurations of PSU can be considered (Figure 4b). Most common use is parallel combination at the input and output of individual stages (PIPO), or serial–parallel combination SIPO–serial input, parallel output (Figure 4b). With the use of these possibilities, it is possible to supply required value of output voltage, while power delivery can be extended by connection of individual power stages based on the application requirements.

Mutual control of modular architecture is important task due to several issues like, uniform power delivery, uniform thermal stress, the most efficient operational point, etc. Standard single DC/DC converter has principal efficiency characteristic that is shown on Figure 5. When parallel combination of DC/DC stages will be used, the operational region with the peak efficiency can be extended for wide power range of modular power supply system (Figure 6). The proper control of individual modules therefore must be focused on the achievement of the highest efficiency operational point. Other advantages of modular architecture oppose to single stage converters are:

- Low output ripple of current and voltage
- High-efficiency operation for various power deliveries
- Individual outputs of several modules, or common output DC link
- Redundancy and higher reliability oppose to one power supply

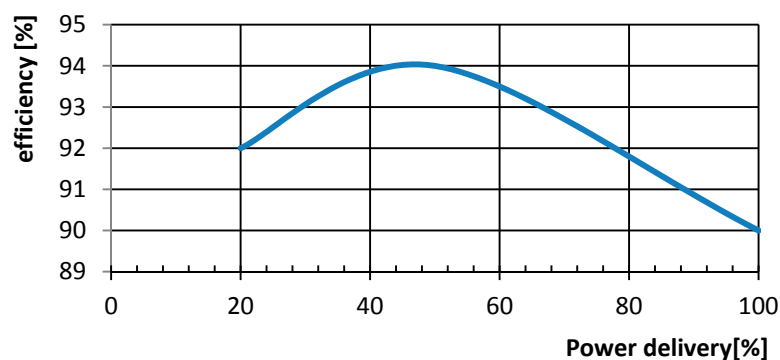


Figure 5. Efficiency curve of standard DC/DC converter stage.

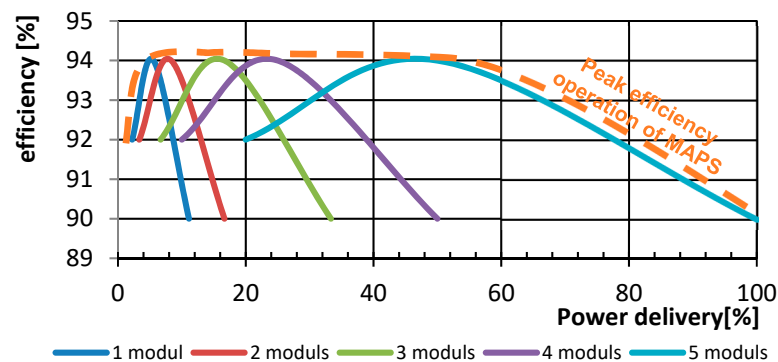


Figure 6. Efficiency curves of parallel configuration of five DC/DC stages.

For the experimental investigation of the modular power supply system properties, the physical sample of proposed LCCT converters have been realized in the configuration PIPO (Figure 7a). The main circuit parameters of the one module are listed in Table 1, while the input–output configuration of the entire modular system (Figure 7b,c) are:

- Input voltage = 400 Vdc
- Output voltage = 48 Vdc
- Output modular power = 1500 W

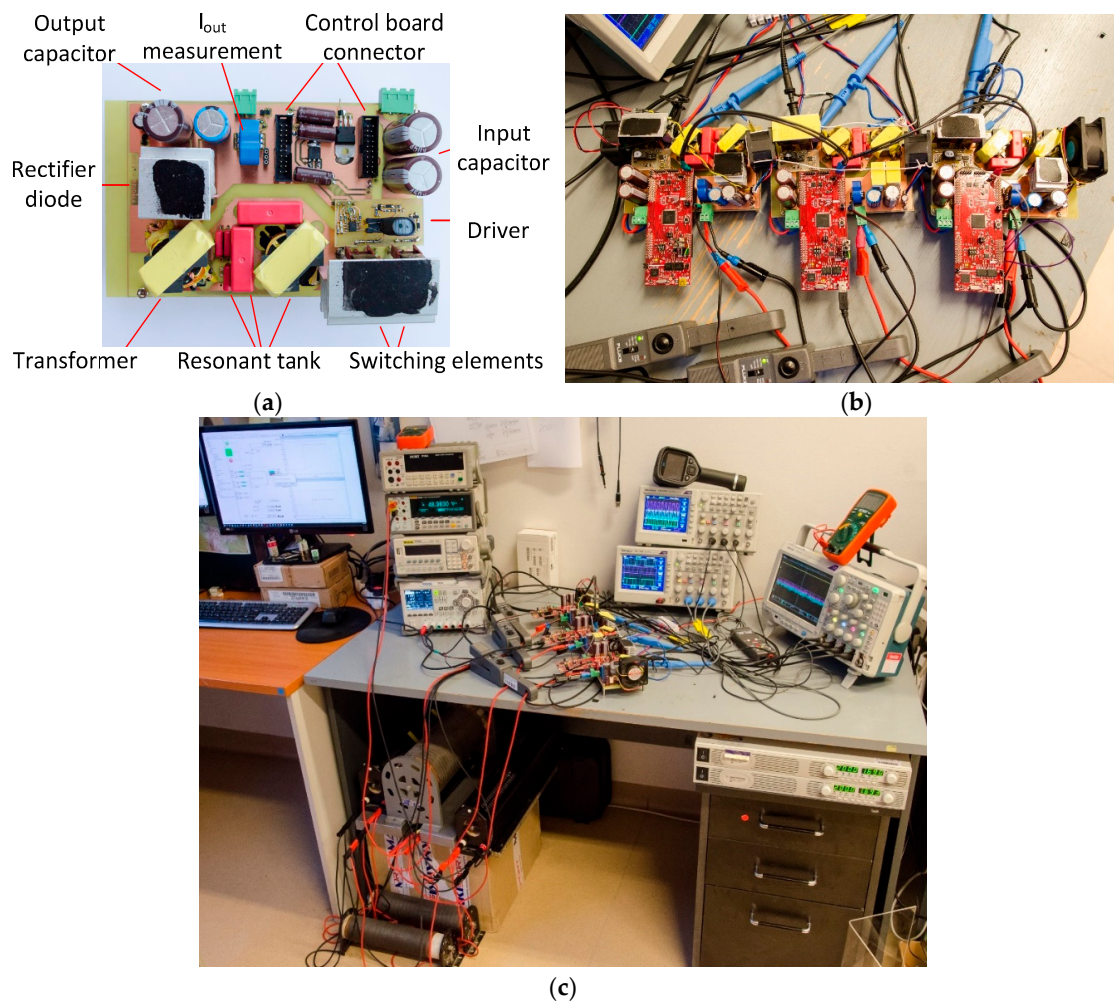


Figure 7. Physical prototype of proposed LCCT converter (a), and its modular concept (b), within experimental set-up (c).

5. Control Strategies for the Proposed Modular Power Supply System

For the control proposal of the modular architecture of power supply system, several approaches can be utilized. The first one can be described as non-synchronized, while each converter is identical to others, thus common pulsed-width modulation is applied for each converter. This solution is not proper, because qualitative indicators of the power supply system is harmed through this approach, i.e.,:

- Not optimized voltage and current ripples at the output of converter
- Existence of beat frequencies
- Wrong voltage or current control due to component intolerances
- Not optimized efficiency performance of whole system

Another solution is decentralized control what means that individual modules do not have information about the power loading of the others. Such strategy is optimal within applications, where repeated or frequent dynamic load changes are not presented. The mentioned approach is mostly used within current sourced power supplies [37,38]. In this paper, two improved control strategies are presented. They are considered as centralized control strategies, thus supervising system (MASTER module) is being utilized, while control commands for each SLAVE module are managed by the MASTER module.

5.1. Control Strategy of MAPS Based on Frequency Follower Technique

The first control proposal is based on the frequency follower (Figure 8). The MASTER module can be any of the individual converters, whereby its control loop is different in comparison with SLAVE modules. Each module contains own output current regulator and the regulator for compensation of the output voltage drop. The computation of the required output current value at the output of each module is based on (8).

$$I_{ref} = \frac{V_{ref} - V_{out}}{R_{droop}} \quad (8)$$

where R_{droop} is 0.1Ω , I_{ref} is compared with actual output current value. Because each of the module is independent from the current regulator point of view, the condition related to the system redundancy is achieved.

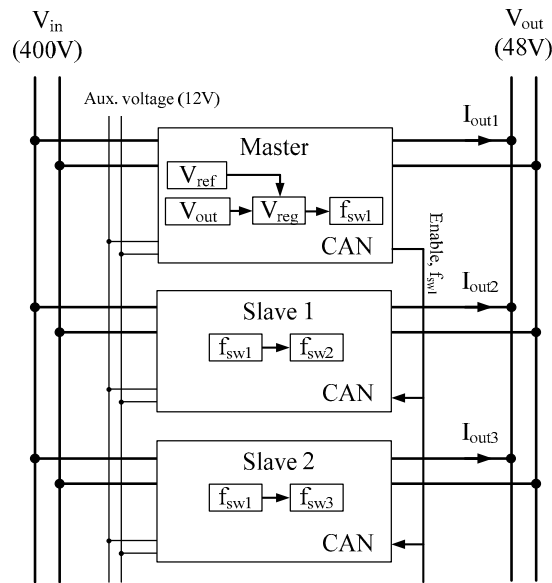


Figure 8. Block diagram of the control strategy proposal for MAPS based on the frequency follower.

For the communication between individual subsystems, CAN bus is utilized. This approach was selected due to availability of this function on the control system which was based on the LaunchpadXL 28069M (Texas Instruments, Dallas, Texas, TX, USA). The MAPS is regulated on the constant output voltage (48 Vdc), while the current has limited maximal allowable value.

MASTER module is monitoring the value of output voltage, value of the output current, and contains the PI regulator for the control of constant output voltage. The PWM signals are then transmitted through the CAN bus to the other SLAVE modules, based on the information from the regulator of the MASTER module.

The dynamic performance was investigated, i.e., start-up of the system for full load (Figure 9) and for 20% of P_{OUT} (Figure 10). Also, dynamic change from initial 100% of P_{OUT} to 20% of P_{OUT} was investigated (Figure 11).

It is seen on Figure 9 that the output voltage waveform (blue line) is common for each module, while the differences within the output currents (blue, cyan, brown lines) are visible. These differences are related to the differences of the transfer characteristics of individual modules (component tolerances), thus current unbalance of MAPS is evident. The more visible current unbalance is for the start-up sequence of MAPS when 20% of P_{OUT} is applied at the output.

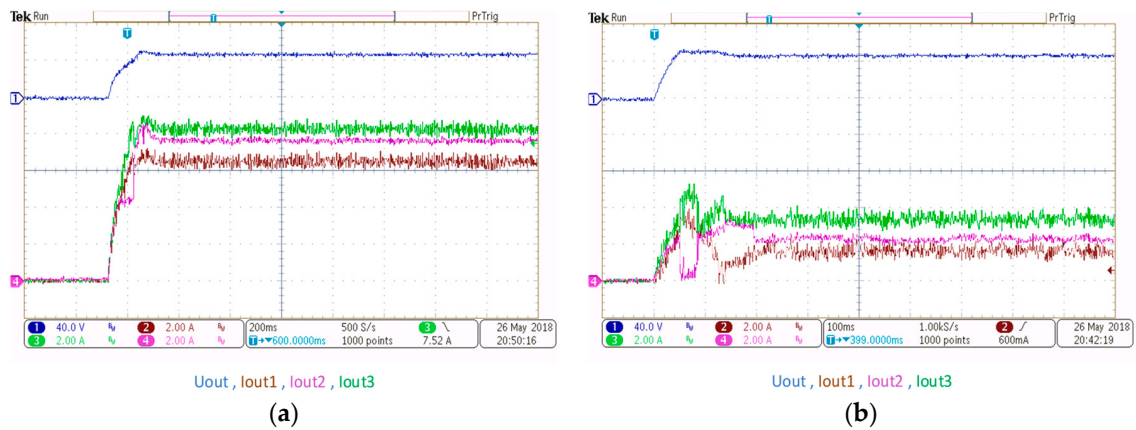


Figure 9. MAPS start-up sequence for frequency follower control strategy. (a) ($P_{OUT} = 100\%$), (b) ($P_{OUT} = 20\%$).

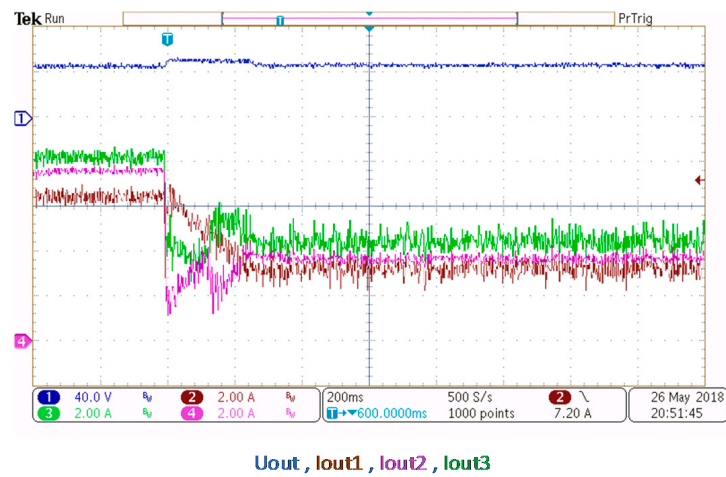


Figure 10. Dynamic load change from 100% to 20% of P_{OUT} for frequency follower control strategy.

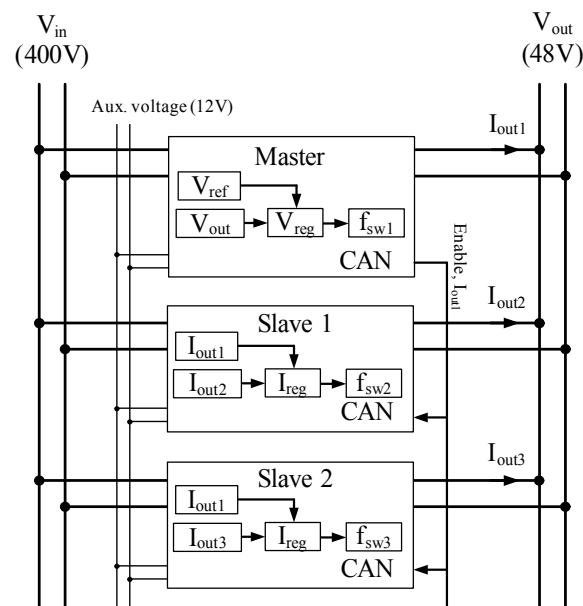


Figure 11. Block diagram of the control strategy proposal for MAPS based on the MASTER-voltage, SLAVE-current.

The dynamic load change experiment was realized (Figure 10), while the MAPS responds to the requirement on the output power reduction from initial 100% to 20%. It is again seen, that the current unbalance is again valid, thus the converters are unevenly loaded. The consequence of this fact is higher depreciation of the module which has the highest load, which can be negatively reflected within lifetime of the individual components. Another negative issue is uneven thermal performance of individual modules.

To eliminate these unbalances, the second control strategy is proposed and is based on the uniform power share at any condition of the MAPS.

5.2. Control Strategy of MAPS Based on MASTER-Voltage and SLAVE-Current Technique

The optimized solution for the control of MAPS is represented by the control technique, which considers problematics related to the current sharing of the individual modules (Figure 11). This approach eliminates the lack of the previous centralized control technique.

The main MASTER module contains PI regulator of the output voltage and sensor for the output current I_{OUT1} . The data relevant to the value of I_{OUT1} are then transmitted through the CAN bus to other SLAVE modules. This information acts as the current reference value, which is processed by the current regulators of SLAVE modules. The procedure can be then described as follows, i.e., the more current the SLAVE modules will supply the more output voltage rise will be evident on the common output. This will cause reaction on the MASTER module, which starts to initialize the I_{OUT1} to be reduced, what will stabilize the output voltage as well as output currents (I_{OUT1} , I_{OUT2} , I_{OUT3}).

The switching frequency of each module is independent, because each regulator defines its value for individual modules, thus the value of the I_{OUT1} , I_{OUT2} , I_{OUT3} must be even (Figure 12). Presented approach is valuable due to existence of the component tolerances.

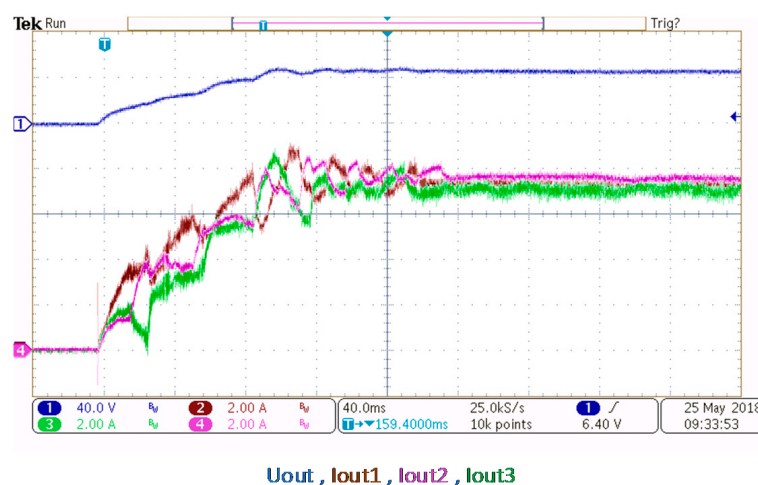


Figure 12. MAPS start-up sequence for MASTER-voltage, SLAVE-current control strategy ($P_{OUT} = 100\%$).

According to the operational characteristic at various output power of the MAPS and considering MASTER-voltage and SLAVE-current control strategy, the optimization of the sequential switching of the individual modules based on the actual load is realized. This approach is done because of the requirements on the efficiency performance of the MAPS (Figure 6) targeting the most efficient operation at any point.

The approach for the proposed system is that for P_{OUT} between 0–25% just MASTER module operates (Figure 13). SLAVE 1 module enters to the operation if the value of the I_{OUT} exceeds 8A (80% of output power of single module). From this point two modules operate, while for the elimination of the negative influences from the start-up of SLAVE 1, the operational area of MASTER + SLAVE1 modules is limited by the hysteresis of the I_{OUT} . From Figure 13 is seen, that if I_{OUT} drops below 4A, SLAVE1 is

disconnected from to operation. MAPS full operation comes to account, when the I_{OUT} exceeds 16A (50% of total P_{OUT}). From this point up to full load operation all modules operate, while each module is loaded by the same value of the power, thus the system is balanced. The sequential switching itself can be influenced also by the thermal performance, whereby if the allowable temperature of a single module will be exceeded during its long-term operation, then other can be switched to balance the actual state. Presented approach can be applied on the various number of the modules.

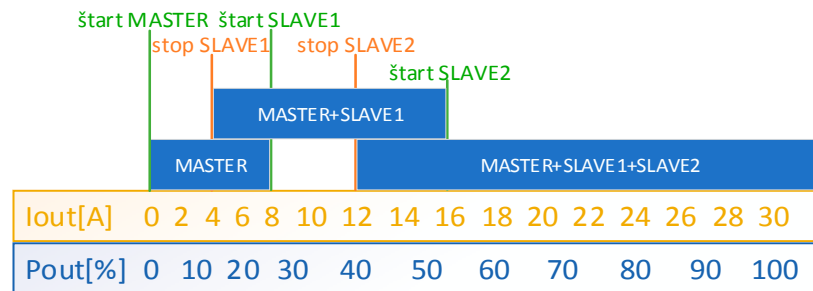


Figure 13. Technique of the sequential switching of MAPS within MASTER-voltage, SLAVE-current control strategy.

Figure 14 shows experimental results from the sequential switching of MAPS, while the output voltage and output currents waveforms are shown. Initial dynamic change from 25% to 35% is shown, whereby after I_{OUT1} (MASTER) exceeds 8A SLAVE 1 is being switched. Result is that I_{OUT1} is reduced and both I_{OUT1} and I_{OUT2} have the same value.

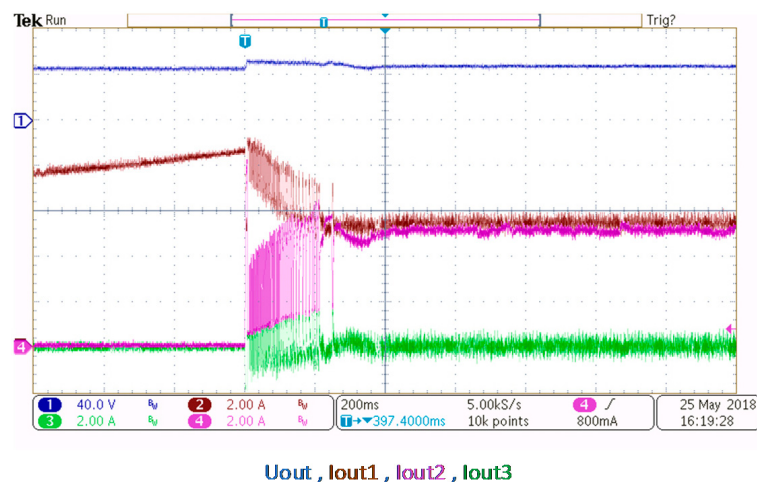


Figure 14. Experimental results from sequential switching of MASTER-voltage SLAVE-current control strategy ($P_{OUT} = 25\%$ to 35%).

Reverse experiment was realized for the situation when MASTER and SLAVE 1 have been operating and the requirement on the P_{OUT} reduction to 15% was initialized. From Figure 15 is seen, that the I_{OUT2} of SLAVE 1 has decreased to zero, and just MASTER module continues the operation. In this way, the requirement on the efficient operation of MAPS is guaranteed (Figure 6).

Final confirmation of the sequential switching operation was realized for the P_{OUT} increase from 40% (two module operation) up to 60% (three module operation). Figure 16 is showing that during load increase above 16A, SLAVE 2 initialized its operation and all output currents I_{OUT1} – I_{OUT3} reached even value, thus the balanced operation is guaranteed.

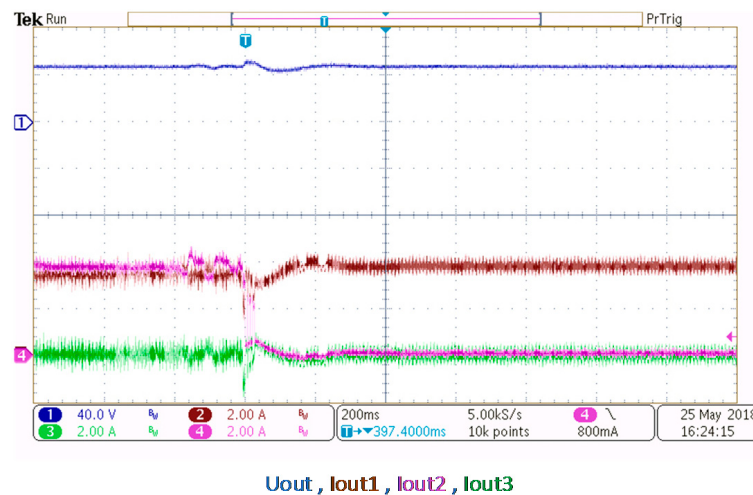


Figure 15. Experimental results from sequential switching of MASTER-voltage SLAVE-current control strategy ($P_{OUT} = 35\%$ to 15%).

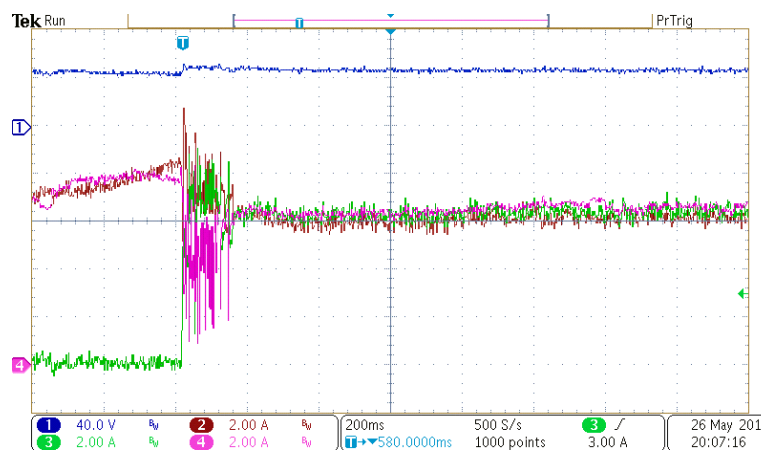


Figure 16. Experimental results from sequential switching of MASTER-voltage SLAVE-current control strategy ($P_{OUT} = 40\%$ to 60%).

5.3. Efficiency Performance of Proposed Control Strategy

The evaluation of proposed control strategy was verified also by the efficiency investigation. The experiment set to the relevancy of the expectations, which were described on Figure 6. Initially, the operation of individual modules was done, thus one module was operated up to its nominal power (500 W). Then two modules in PIPO configuration have been investigated, while the PWM signals of both converters have been mutually phase-shifted by 180° . The nominal power of this two-module configuration was 1000 W. After it the three-module operation was investigated from efficiency point of view, while again the modules were connected in PIPO and the nominal output power was the same as for the proposed MAPS system (1500 W). For three-module operation the PWM signals were also mutually phase-shifted by 120° . Finally, the proposed control strategy MASTER-voltage, SLAVE-current for proposed MAPS was tested (Figure 17). It is seen, that through the use of proposed control strategy it is possible to achieve the most efficient operation considering parallel operation of converter modules. At this point it shall be noticed, that proposed control strategy is suitable for various converter topologies where constant output voltage is required. It means that the efficiency improvement can also be done through utilization of perspective devices, materials, and variations of converter's main circuit.

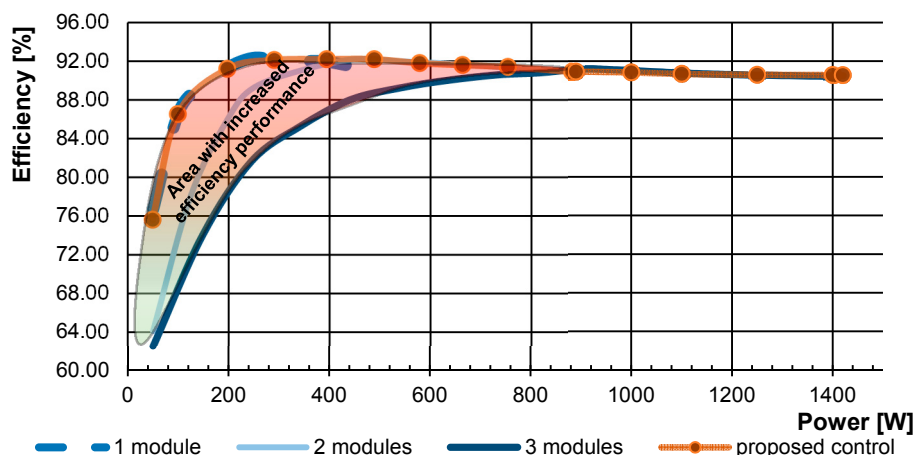


Figure 17. Efficiency performance of individual module operation (frequency follower) in comparison with MASTER-voltage, SLAVE-current technique.

6. Conclusions

In this paper, the LCCT converter was discussed as possible solution for modular architecture of power supply (MAPS) systems. Initially design of main circuit of proposed converter was briefly described following description of its operational regions. Consequently, experimental samples of LCCT converter with 48 V/500 W parameters have been prepared, while solution for three-module MAPS was considered during investigation of the control strategy of such system. For the flexible modification and user comfort, digital control was used based on the LaunchpadXL 28069M. For the comparison purposes, standard control technique based on frequency follower was initially verified for the MAPS operation. It was found that within this method the converters are operating unevenly from the power delivery point of view. Such an approach may negatively affect thermal performance of the system as well as life-time and reliability. Therefore, it was necessary to improve control strategy in the way of the even power sharing of individual modules within MAPS system operation. For this purpose, MASTER-voltage, SLAVE-current technique has been proposed and experimentally verified, while it was found that during various operational conditions, the converters evenly share delivered power. Also, sequential switching was proposed based on the value of the delivered power in order to utilize individual modules at the most proper operational point considering efficiency. The final verification of proposed control strategy was done in the way of the investigation of efficiency of MAPS system, where frequency follower technique and MASTER-voltage, SLAVE-current have been compared. It was found that proposed control improves efficiency performance of the system within low-power range of the system, i.e., 0%–50% of P_{OUT} . Future works will be focused on the utilization of proposed technique on multi-element converter systems (above three modules) and on the efficiency optimization through the use of perspective devices and main circuit topologies.

Author Contributions: M.F. prepared the paper structure and has organized individual works within the paper preparations, made experimental approach and finally he wrote the paper. J.M. has designed experimental prototype of proposed converter has organized laboratory measurements. P.S. is responsible for the state of the art review within the area of the proposed converter. B.D. has performed all the mathematical support for the investigated converter topology.

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Conflicts of Interest: The authors declare no conflict of interest.

Appendix A Complex Mathematical Model of Proposed LCCT Converter

The design process of LC resonant tanks should be considered from three different points of view or criteria:

- Considering the nominal voltage and current stresses at steady-state operation
- Considering minimum voltage and current stresses during transient responses
- Considering the required value of total harmonic distortion of the output voltage

Note: Using real transformer, its parasitic like resistances of transformer ferromagnetic, the capacitance between winding, etc. can be infiltrated within the resonant LCLC or LCCT filter. The main difference between LCLC and LCCT configuration is that L_2 from LCLC topology is assumed to be magnetizing inductance of the transformer (Figure A1).

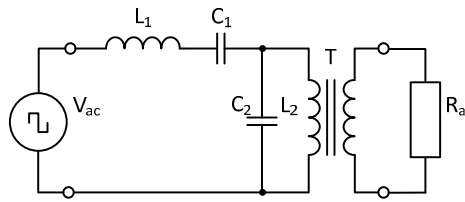


Figure A1. A Numerical solution using the state-variable method.

A. Numerical solution using the state-variable method

The numerical expression of the output voltage of the switching part of single-phase inverter i.e., input voltage of the resonant LCCT filter

$$v_{IN}(k) = \sqrt{2} V_{DC} \sin \left[\text{integer} \left(\frac{4}{T} \Delta T k \right) \frac{\pi}{2} + \frac{\pi}{4} \right] \quad (\text{A1})$$

where U_{DC} —the input voltage of inverter; T —time period; $\Delta T k$ —discrete time.

Continuous dynamic model of resonant LCCT filter with R load

$$\begin{aligned} \frac{di_{L1}}{dt} &= -\frac{r_1}{L_1} i_{L1} - \frac{1}{L_1} v_{C1} - \frac{1}{L_1} v_{C2} + \frac{1}{L_1} v(t) \\ \frac{dv_{C1}}{dt} &= \frac{1}{C_1} i_{L1} \\ \frac{dv_{C2}}{dt} &= \frac{1}{C_2} i_{L1} - \frac{1}{C_2} i_{L2} - \left(\frac{1}{r_2} + \frac{1}{R} \right) \frac{1}{C_2} v_{C2} \\ \frac{di_{L2}}{dt} &= \frac{1}{L_2} v_{C2}. \end{aligned} \quad (\text{A2})$$

Numerical dynamic model of resonant LCLC filter using Euler explicit method

$$\begin{aligned} i_{L1}(k+1) &= i_{L1}(k) + \Delta T \left[-\frac{r_1}{L_1} i_{L1}(k) - \frac{1}{L_1} v_{C1}(k) - \frac{1}{L_1} v_{C2}(k) + \frac{1}{L_1} v_{IN}(k) \right] \\ v_{C1}(k+1) &= v_{C1}(k) + \Delta T \left[\frac{1}{C_1} i_{L1}(k) \right] \quad (\text{A3a, b, c, d}) \\ v_{C2}(k+1) &= v_{C2}(k) + \Delta T \left[\frac{1}{C_2} i_{L1}(k) - \frac{1}{C_2} i_{L2}(k) - \left(\frac{1}{r_2} + \frac{1}{R} \right) \frac{1}{C_2} v_{C2}(k) \right] \\ i_{L2}(k+1) &= i_{L2}(k) + \Delta T \left[\frac{1}{L_2} v_{C2}(k) \right] \end{aligned} \quad (\text{A3})$$

where v_{C2} is the output voltage of the resonant LCCCT filter ($v_{C2} = v_{out}$).

Using successive approximation method, we can solve as steady states as a transient dynamic state. Resulted values of state variable as i_{L1} , v_{C1} , v_{C2} , i_{L2} and their rms values (V_{outN}) can be used for determining of resonant element values in Equations (3)–(6) in the text Section 2.

B. Analysis in frequency domain (AC sweep)

Nominal impedance and nominal admittance resonant *LLTC* filter with a resistive load

$$|Z_N| = \frac{V_{outN}^2}{P_{outN}}; |Y_N| = \frac{P_{outN}}{V_{outN}^2} \quad (A4)$$

The impedance of the series and the parallel part of the resonant *LCCT* filter is defined by the following equations

$$\begin{aligned} Z_1(\omega) &= r_1 + j\left(\omega L_1 - \frac{1}{\omega C_1}\right) = r_{1N}|Z_N| + j|Z_N|q_N\left(f_{rel} - \frac{1}{f_{rel}}\right) \\ &= |Z_N|\left[r_1 + jq_N\left(f_{rel} - \frac{1}{f_{rel}}\right)\right] \end{aligned} \quad (A5)$$

where: $r_{1N} = \frac{r_1}{|Z_N|}$; —the parasitic resistance of the series part of the filter; $f_{rel} = \frac{\omega}{\omega_{res}}$ —is relative variable frequency (practically, in real apparatus, $\omega \equiv \omega_{switch}$ i.e., of the inverter); q_N —nominal quality factor; Z_N —nominal impedance.

Thus

$$\frac{|Z_1(\omega)|}{|Z_N(\omega)|} = \sqrt{r_1^2 + \left[q_N\left(f_{rel} - \frac{1}{f_{rel}}\right)\right]^2}, \quad (A6)$$

and similarly, we can calculate the impedance of the parallel part of the parallel part of the filter. For the admittance of this part valid

$$\begin{aligned} Y_2(\omega) &= \frac{1}{r_2} + \frac{1}{R} + j\left(\omega C_2 - \frac{1}{\omega L_2}\right) = \frac{1}{r_{2N}}|Y_N| + \frac{1}{r_N}|Y_N| + j|Y_N|q_N\left(f_{rel} - \frac{1}{f_{rel}}\right) = \\ &= |Y_N|\left[\left(\frac{1}{r_2} + \frac{1}{r}\right) + jq_N\left(f_{rel} - \frac{1}{f_{rel}}\right)\right] \end{aligned} \quad (A7)$$

where: $\frac{1}{r_{2N}} = \frac{1}{r_2|Y_N|}$ —parasitic conductance of parallel part of the filter; $\frac{1}{r_N} = \frac{1}{R|Y_N|}$ —conductance of the load resistance; $f_{rel} = \frac{\omega}{\omega_{res}}$ —is relative variable frequency (practically, in real apparatus, $\omega \equiv \omega_{switch}$ i.e., of the inverter); q_N —nominal quality factor; Y_N —nominal admittance.

Thus

$$\frac{Y_2(\omega)}{Y_N(\omega)} = \left(\frac{1}{r_2} + \frac{1}{r}\right) + jq_N\left(f_{rel} - \frac{1}{f_{rel}}\right), \text{ and } \frac{|Y_2(\omega)|}{|Y_N(\omega)|} = \sqrt{\left(\frac{1}{r_2} + \frac{1}{r}\right)^2 + \left[q_N\left(f_{rel} - \frac{1}{f_{rel}}\right)\right]^2}, \quad (A8)$$

hence

$$\frac{Z_2(\omega)}{Z_N(\omega)} = \frac{1}{\left(\frac{1}{r_2} + \frac{1}{r}\right) + jq_N\left(f_{rel} - \frac{1}{f_{rel}}\right)} = \frac{\left(\frac{1}{r_2} + \frac{1}{r}\right) - jq_N\left(f_{rel} - \frac{1}{f_{rel}}\right)}{\left(\frac{1}{r_2} + \frac{1}{r}\right)^2 + \left[q_N\left(f_{rel} - \frac{1}{f_{rel}}\right)\right]^2} \quad (A9)$$

Since we denote the denominator as

$$\begin{aligned} DEN &= \left(\frac{1}{r_2} + \frac{1}{r}\right)^2 + \left[q_N\left(f_{rel} - \frac{1}{f_{rel}}\right)\right]^2 \\ \frac{|Z_2(\omega)|}{|Z_N(\omega)|} &= \sqrt{\frac{\left(\frac{1}{r_2} + \frac{1}{r}\right)^2 + \left[q_N\left(f_{rel} - \frac{1}{f_{rel}}\right)\right]^2}{DEN^2}} = \sqrt{\frac{1}{DEN}} \end{aligned} \quad (A10)$$

and relative input impedance

$$\frac{Z_{in}(\omega)}{Z_N(\omega)} = \frac{Z_1(\omega) + Z_2(\omega)}{|Z_N(\omega)|} = \left[r_1 + \frac{\left(\frac{1}{r_2} + \frac{1}{r}\right)}{DEN}\right] + j\left[\left(f_{rel} - \frac{1}{f_{rel}}\right)\left(q_N - \frac{q_N}{DEN}\right)\right] \quad (A11)$$

The module of the relative input impedance $Z_{in}(\omega)$ of a resonant LCCT filter in the relation of the nominal impedance $Z_N(\omega)$

$$\frac{|Z_{in}(\omega)|}{|Z_N(\omega)|} = \sqrt{\left[r_1 + \frac{\left(\frac{1}{r_2} + \frac{1}{r}\right)}{DEN}\right]^2 + \left[\left(f_{rel} - \frac{1}{f_{rel}}\right)\left(q_N - \frac{q_N}{DEN}\right)\right]^2} \quad (A12)$$

The voltage-frequency transfer (voltage gain) will be

$$G(\omega) = \frac{|V_{out}(\omega)|}{|V_{in}(\omega)|} = \frac{|Z_2(\omega)|}{|Z_{in}(\omega)|} = \sqrt{\frac{1}{DEN \left\{ \left[r_1 + \left(\frac{1}{r_2} + \frac{1}{r} \right) \frac{1}{DEN} \right]^2 + \left[\left(f_{rel} - \frac{1}{f_{rel}} \right) \left(q_N - \frac{q_N}{DEN} \right) \right]^2 \right\}}} \quad (A13)$$

Graphical interpretation of $G(\omega)$ is given in Figures 2 and 3 in the text of the paper.

It is important to note, that for f_{rel} equal one (switching frequency = resonant frequency) the voltage gain does not practically depend on the load. Consequently, by the change of switching frequency is possible partially to control the output voltage of the resonant LCCT filter. Another way for how to control the output voltage is the use of classical pulse-width control (or phase-shift control, respectively) when the amplitude of fundamental harmonic is

$$V_{1RMS} = \frac{2\sqrt{2}}{\pi} V_{DC} \sin\left(\frac{\beta}{2}\right)$$

where β is the width of voltage pulses.

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