Control strategy for efficient operation of super-resonant SLSR (contactless) converters

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Abstract: – **A new faster control method is presented, in attempt to achieve stable operation and higher efficiency of any Series Loaded Series Resonant (SLSR) power converter, and especially when the application requires contactless energy transfer. This instantaneously reacting control method is based on calculated individual energy portions delivered to the resonant circuit. Its viability is demonstrated by simulation of an analogue circuit implementation.**

Keywords: – **Series Loaded Series Resonant Converter, Contactless Energy Transfer, Energy Portions Control.**

I. Introduction

The resonant power conversion is reluctantly but gradually accepted by the industry especially in military and space applications. The operation of the SLSR power converter is already analysed, e.g. in [1], [2], [3] and many more but the fast response of its control is a problem. The existence of internal energy tanks (resonant inductance and capacity) makes the task of controlling the processes quite difficult, especially when the circuit elements are not ideal (contactless energy transfer). Many works are published, aimed at resonant converters control. Those usually include calculation of normalized phase-plane trajectory as V. Nguen et al. [5], and L. Rossetto [6]. The calculations are complicated and need multiple points of real-time measurements in order to recalculate the moments of necessary commutations, e.g. in [7] of B. Souesme and [8] of M. Kim. A work was initiated in 1991 at TU Delft to find a control method, suitable for the SLSR case [4], controlling the energy flux by heavy calculations.

In the recent years the SLSR converter is successfully applied for a contactless power transfer, e.g. in [9] of S. Valtchev. This makes more important the research for a new, simplified but fast method to control the resonant processes in the power converter, fixing its operation points in the zone with best efficiency. The work presented here is not a final solution but most of all a demonstration of the possibility to achieve the required operation by simple calculation (also analogue).

II. Classic Methods (Frequency Mode)

The application of frequency mode (FM) regulation is the simplest way to guarantee the zero voltage switching (ZVS) of the power devices when the switching frequency is kept higher than the resonance. Because of this narrow range of achievable output power regulation, combinations with the less efficient pulse-width mode (PWM) of switching are also applied [2]. In this article the solution is given to guarantee the minimum of switching losses and minimum risk of instability, i.e. PWM must be used only when the capacitor voltage amplitude is lower than the supply voltage.

The simplest FM regulation can be achieved by a PLL circuit as in Fig.1. It guarantees the zero-voltage switching (ZVS) but it is not fast and may present problems when fast changes occur (as in contactless power transfer converters).

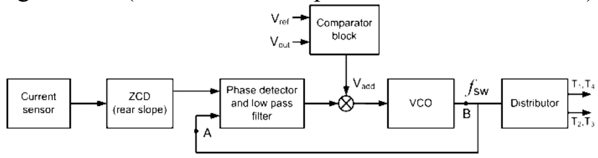

Fig.1. Simplified feedback loop of a PLL control circuit

The circuit in Fig.1 is a closed PLL which maintains the required phase shift between the zero-crossing of the rear slope of the resonant current and the end of the control pulse for the corresponding diagonal of switches. The block named ZCD is detecting the instant when the rear slope of the resonant current pulse is crossing the zero. This zero crossing marks the end of the half-period of the resonant current pulse. This point (designated by x_o) is shown in Fig.2. The output of VCO produces a train of pulses, each ending at the instant x_k. The switching frequency f_{sw} produced by the VCO will assure the required phase shift between x_k and x_o. The corresponding transistor receives the VCO signal as a command pulse applied to its gate. This pulse finishes at the moment x_k. Identical pulses (finishing at x_k) are applied to both transistors in each diagonal of the power bridge.

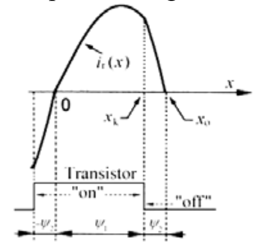

Fig.2. Resonant current i_r intervals and power switch command.

The minimum phase shift between the rear slope zero crossing of the current (x_o) and the end of the transistor conduction interval (x_k) corresponds to a minimum length of the diode conduction interval ψ_2. The zero phase difference (ψ_2=0) should be avoided (i.e. the switching frequency is dangerously close to the resonant frequency). This danger is prevented by applying the voltage V_{add} produced in the Comparator block presented in Fig.1. The voltage V_{add} must have a guaranteed minimum value in order to produce the necessary phase shift for a zero voltage switching (ZVS): the transistor is to be switched on always before its diode cuts off

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its current. The regulation of the output is accomplished by further increasing of the voltage V_{add} and thus increasing the width of the normalized time interval ψ_2. The output current is obtained in function of the normalized output voltage $q = \frac{Vo}{Vin}$ and the ψ_2 interval:

$$I_o = \frac{2\dfrac{(1+q)(1-\cos\psi_2)}{\cos\psi_2 - q}}{\psi_2 + \arccos\left(\dfrac{1-q-q\dfrac{(1+q)(1-\cos\psi_2)}{\cos\psi_2 - q}}{1-q+\dfrac{(1+q)(1-\cos\psi_2)}{\cos\psi_2 - q}}\right)} \tag{1}$$

The regulation function is presented graphically as $q=f(I_o, \psi_2)$ Fig.3 and can be achieved by integrated circuits providing fixed pulse width with regulated distance between the pulses. The examples can be UC1860 produced by UNITRODE (TI), MC33066 by MOTOROLA, etc.

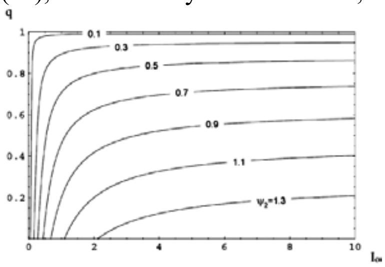

Fig.3. Output voltage in function of output current at fixed ψ_2

The graphics in Fig.3 show a voltage source character of this regulation. A better regulated FM/PWM operation converter is already presented in [2]. It shows in most of the operation area a current source character which is convenient for safer parallel operation of many stages. One version of a mixed PWM/FM control is implemented by the integrated circuit UC1825. The circuit is a PWM regulator which internal oscillator frequency is controlled externally, as presented in Fig.4.

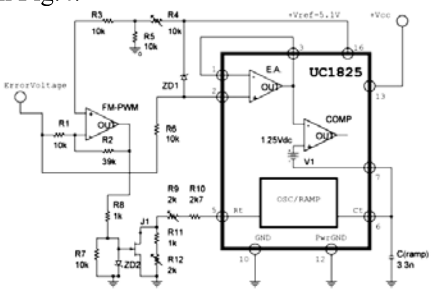

Fig.4. Example of mixed operation (FM/PWM) regulation circuit

The integrated circuit UC1825 is allowed to operate as a normal PWM regulator for the lowest error voltages (this is the proper mode of operation when the demand of power is low). At high demand, the operation is FM.

III. Instantaneous Control of SLSR Converter

A. Energy Balance

The excitation voltage V_{LC} applied across the LC tank is $V_{LC1}>0$ during the interval ψ_1, i.e. t_1 and $V_{LC2}<0$ in the interval ψ_2, i.e. t_2. The energy of the resonant circuit repeats its value after each half period of switching and in a steady-state operation no increase or decrease is observed:

$$\Delta E_{LC} = \int_0^{t_1} V_{LC1} i_r dt + \int_{t_1}^{\frac{T_{sw}}{2}} V_{LC2} i_r dt = 0 \tag{2}$$

When ΔE_{LC} remains zero, the energy portion extracted from the input voltage source E_s is exactly corresponding to the energy portion delivered to the output voltage source V_o. If it is necessary to increase the energy in the tank, then the circuit will be commanded by a longer conduction interval ψ_1 (longer than the necessary for its steady state), and hence ΔE_{LC} will be positive. When the energy must diminish, then ψ_1 is decreased and ΔE_{LC} will be negative.

At the zero crossing points of the current (Fig.2) the total resonant tank energy consists only of resonant capacitor energy $\frac{C_r v_{c\max}^2}{2}$, so the capacitor voltage maximums v_{cmax} can be used as a measure for that energy. Integrating the expression (2), it can be rewritten:

$$\Delta E_{LC} = V_{LC1}[v_c(t_1) - v_c(0)]C_r + V_{LC2}[v_c(\frac{T_{sw}}{2}) - v_c(t_1)]C_r \tag{3}$$

If the absolute value of the voltage v_{cmax} is marked as v_{cmax1} at the x=0 and v_{cmax2} at $\frac{T_{sw}}{2}$ then it can be written:

$$\Delta E_{LC} = C_r[V_{LC1}(v_{c\max 1} + v_c(t_1)) + V_{LC2}(v_{c\max 2} - v_c(t_1))] \tag{4}$$

The excitation values can be written in (4) from their normalized forms [1]: V_{LC1}=1-q and V_{LC2}=-1-q. The normalized output voltage q is supposed to be constant during one switching period and then the capacitor energy change (in normalized form) is:

$$\Delta E_C^{\,N}(0, x_o) = 2\left[v_c(t_1) - q\frac{(v_{c\max 1} + v_{c\max 2})}{2}\right] \tag{5}$$

Expression (5) predicts the capacitor voltage $v_c(t_1)$ at which it will be necessary to switch off the transistor (t_1 time point corresponds to the angular point x_k). The commutation command will be produced when the calculated value $v_c(t_1)$ becomes equal to the measured capacitor voltage. The prediction has to take into account the previously measured amplitude v_{cmax} and the necessity for the change (if necessary) of the energy portion.

In the case of contactless power transfer, the converter operation is described by the same expression (5) where v_{cmax1} and v_{cmax2} correspond to the resonant capacitor voltage amplitudes in the primary [9]. In that case, the normalized output voltage q in (5) will be substituted by the corrected value $q^T = Kq$, where K is the magnetic coupling factor of the loosely coupled magnetic link.

B. Simplified strategy for regulation

It is important to choose a convenient variable which level can serve as the indicator for the energy transferred each half period in the SLSR converter. This variable (v_c) must reflect proportionally the size of the energy portions transferred in each half period. During the transient process by the value of this variable, it must be possible to predict and change the size of the energy portions circulating in the resonant tank.

In the hard switching converter with Current Mode regulation, the inductor current transferring the periodically repeated portions of energy is the instantaneous power indicator. Considering that the resonant processes are not so simple, the Current Mode regulation is not directly applicable in the SLSR converter. For example, the peak value of the resonant current does not correspond to the maximum stored energy in the resonant loop. In contrary, the SLSR converter shows output characteristics similar to a current source, with curves that fall to limited current values as shown in [1], [2]. In this case, the internal variable is better to be the voltage, proportional to the energy portion.

From the regulation point of view, the only controllable variable that the power switches can commutate is the current through the transistors. In the same time, by controlling the current conduction, the transistors control the charge of the resonant capacitor. The interchange of inductive and capacitive energy does not easy allow to measure the total energy in every time point, but the total energy can be easily measured in the points of resonant current zero crossings.

The SLSR converter may keep its steady-state operation during enough long time (i.e. many switching periods) if there is no variation of the load and no change of the input power source parameters, as well. In that case the v_{cmax1} and v_{cmax2} are equal and transistor must switch off at:

$$v_c(t_1) = qv_{c\max} \tag{6}$$

In Fig.5 the time point at which the transistor conduction is turned off is defined by the previous amplitude v_{cmax1} and the desired next amplitude of the capacitor voltage v_{cmax2}.

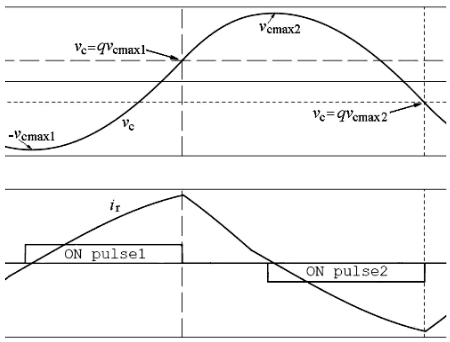

Fig.5. Switching off of the resonant current in SLSR converter

Fig.5 proves the steady-state equation (6). To keep the amplitudes of the resonant capacitor voltage v_{cmax} unchanged, it is necessary to keep the switching off at the defined level (6). This will require a calculation by multiplying the normalized output voltage q and the last measured amplitude voltage of the resonant capacitor.

The transition process from a lower output power consumed in an initial steady-state, to a higher level of output power, will require several portions of additional energy to be supplied by the resonant tank. The amplified error voltage (proportional to the difference between the really obtained and the required output voltage) will define the necessary portions of energy for the transition (similarly to the Current Mode Control).

The requirement for a positive increment ΔE_{LC} in the energy portions corresponds to a higher consumption of power at the output. The control reaction will be to produce the turn-off of the transistor diagonal at a higher level of resonant capacitor voltage expressed in (7.9):

$$v_c(t_1)_{new} = qv_{c\max 1} + q\Delta v_{c\max 1} = q\left(v_{c\max 1} + \Delta v_{c\max 1}\right) \tag{7}$$

The expression (7) is approximated but it is more practical for the resonant regulation process. It expresses the energy increment ΔE_{LC} as a linear proportion to the capacitor voltage amplitude change Δv_{cmax1}. To apply a more accurate method to determine the necessary energy portion will be more difficult to implement.

The calculation of (7) is implemented by memorizing the initial value of the resonant capacitor voltage v_{cmax1}. After that the voltage Δv_{cmax} (proportional to the output voltage error signal) is added, then the sum is multiplied by q in order to correspond to the expression (7). The calculated new level $v_c(t_1)_{new}$ is delivered to the comparator. When the resonant capacitor voltage reaches the reference $v_c(t_1)_{new}$ the transistor current is switched off similarly to the illustrated by Fig.5. After the transistors are switched off, the capacitor voltage is expected to reach the new amplitude value v_{cmax2}. The process is limited by the already running v_c waveform, so the maximum required next reference level $v_c(t_1)_{new}$ cannot be higher than the initial value v_{cmax1}. This means that the full transition process will need several half periods.

C. Implementation in block diagrams

The implementation of the idea described requires several electronic blocks in order to memorize, calculate, compare and limit the values of the variables that the internal sensors will measure. The simplified block diagram of the whole power converter is presented in Fig.6.

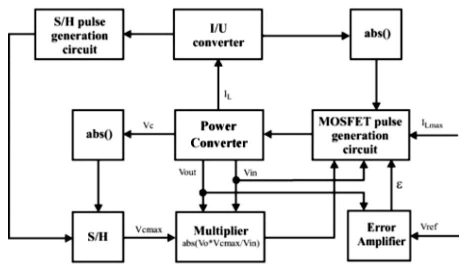

Fig.6. Block diagram illustrating the method for SLSR control

A more detailed block diagram is presented in Fig.7.

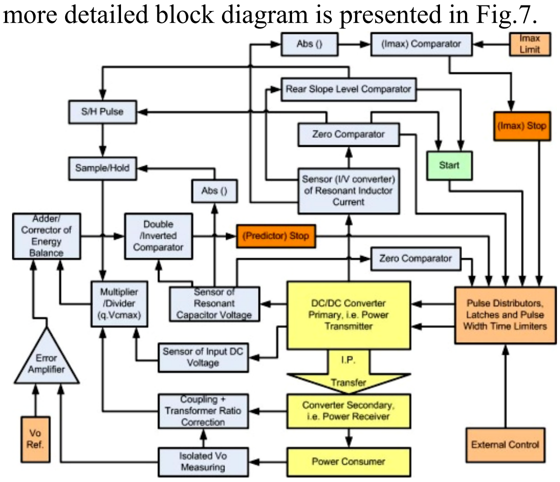

Fig.7. Detailed functional diagram of SLSR control

C. Realistic simulation

Analogue circuits (combined with some logic ICs) are chosen to demonstrate the described idea in more realistic operation (Fig.8).

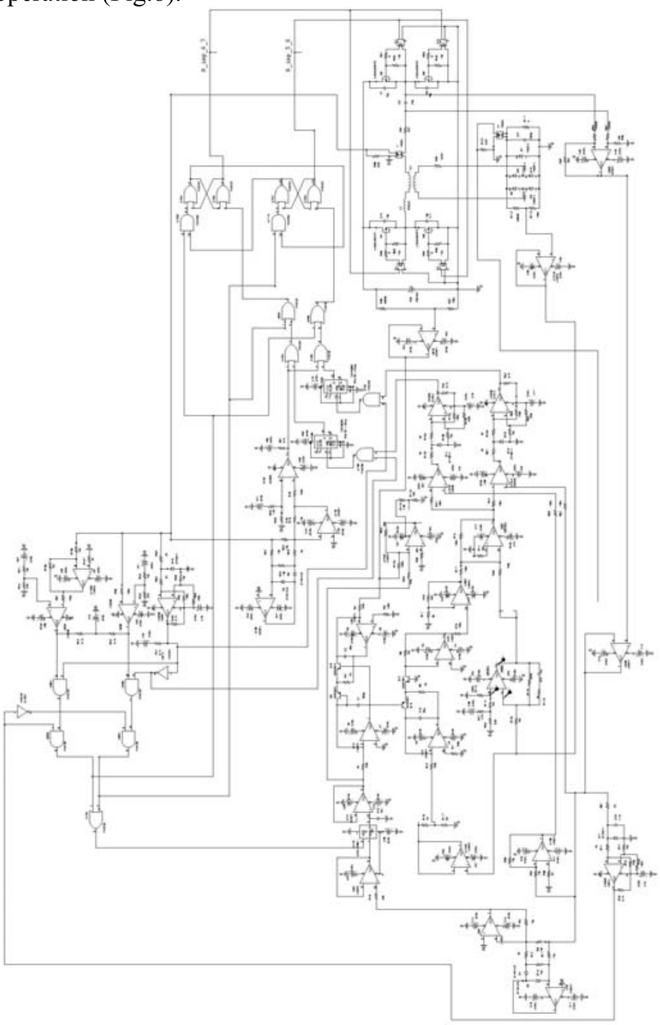

Fig.8. The complete simulated circuit.

Some simulation results are presented in Figs.9 and 10.

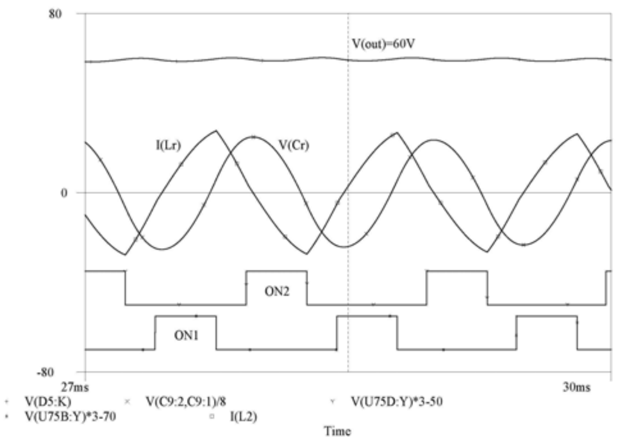

Fig.9. Simulation results of circuit in Fig.8.

The process of fast increase of the output voltage is represented more clearly in a state variable plane (v_c, i_r) plot, as it is shown in Fig.10. The consequent switching-off points are marked as "T_{off}" and show a fast increase in the state variables values.

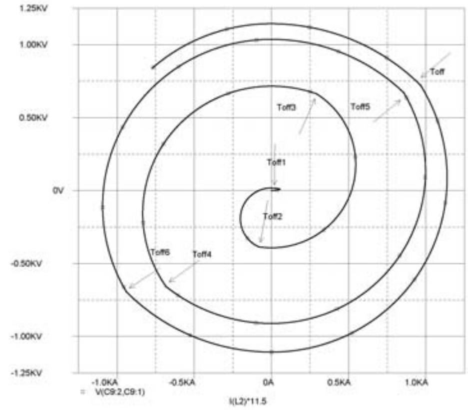

Fig.10. State variable diagram (v_c and i_r).

IV. CONCLUSION

This instantaneous control method permits to regulate the internal processes faster and with higher precision. In case of contactless energy converter the control permits applying the value of the magnetic coupling coefficient as another variable which is also observed and included into the calculation. The presented schematic implementation is mainly analogue and its viability is demonstrated by simulation results. By applying digital circuits, the control will be further improved, becoming more flexible, functional and reliable.

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