Direct Phase Digital Control Method in Power Inverters Based on Dumping Frequency Analysis

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***Abstract*—Direct phase or indirect frequency control methods are used in series-resonant bridge power inverters for induction heating to maintain maximum power transfer as the load equivalent electrical parameters change during the heating process. The paper present mathematical analysis of the output voltage and current phase angle dependence on the resonant circuit dumping frequency when it is excited with pulse voltage with a different frequency than the resonant one. Based on this analysis and using only measurements of the deviation of the phase angle from its reference value, an improved control method is derived that calculates the new bridge switching frequency. An algorithm for digital implementation is presented. The improved method is verified by simulations and experiments on a prototype.**

Keywords—series-resonant bridge inverter; direct phase control method; digital control algorithm; dumping frequency.

I. INTRODUCTION

Series-resonant bridge inverters are used in a variety of applications. Maximum energy is transferred to the load when the converter switching frequency is same to the resonant one. In some applications, like direct induction heating, the heated work-piece equivalent electrical parameters are part of the resonant circuit [1–4]. As the temperature of the work-piece is increased, the resonant tank inductance and resistance change, thus changing the circuit resonant frequency. To ensure maximum energy transfer, the inverter control circuitry must adjust the switching frequency so that it follows the change of the resonant frequency. Different control algorithms are used to adjust the switching to the resonant frequency. Several of them are based on direct frequency control [5–6] and other use indirect frequency control by controlling the phase angle φ between the inverter output voltage and current [7–15]. The last control type, instead of indirect frequency, is more often called direct phase control method. Phase control provides reliable drive of the resonant converter in the presence of large dynamic changes in the load impedance during star-up, natural tracking of component variations with temperature and time, simplified control to output dynamics and a more linear relationship between phase command and output current when compared to frequency control [10, 11].

In the analysis of serial resonant converters it is usual to use the resonant circuit frequency ω_0 for two reasons: 1) assuming that the value of the resistance of the resonant tank is very small the dumping is negligible and thus resonant ω_0 and damping ω_d angular frequencies have very close values; 2) active power is calculated using the phase angle between voltage and current first harmonics. However, in bridge resonant converters the voltage waveforms are pulse and the current has a dumped sinusoidal form. In such cases the phase angle is calculated in respect to the dumping frequency and derived expressions show qualitatively different behavior. Based on dumping frequency analysis of the dependence of the phase angle φ on the switching frequency ω_s we develop improved direct phase control method and verify its performance with simulations and measurements on a prototype of a full-bridge series resonant inverter for induction heating.

II. DIRECT PHASE CONTROL METHOD

Fig. 1 shows a block diagram of the feedback control circuitry used in the direct phase control of full-bridge series-resonant inverter. It comprises of a current transformer that measures the resonant circuit (output) current, a zero crossing detector that gives zero voltage when $i_{out}(t) < 0$ and positive voltage when $i_{out}(t) > 0$, a microcontroller that implements the control algorithm, optocoupler galvanic isolation, a driver circuit that supplies firing pulses to the IGBT switches and a feedback circuit for IGBT overload protection measuring collector-emitter voltages v_{CE} to limit currents through T1 to T4.

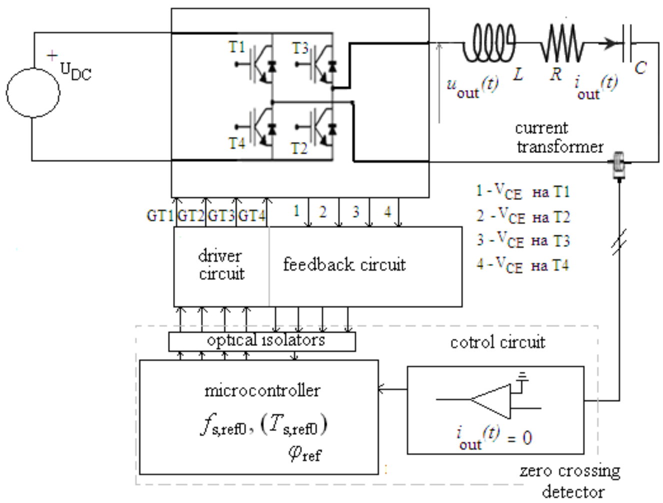

Fig. 1. Block diagram of the full-bridge series-resonant inverter.

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The microcontroller program has predefined values for the initial value of the switching frequency $f_{s,ref0}$ (or period $T_{s,ref0} = 1/f_{s,ref0}$) and the desired or the reference phase difference between the output voltage and current φ_{ref}. This phase difference would be zero or close to zero if maximum power transfer is needed, or have a specific value that corresponds to the desired output power when we like to control the power transfer.

Fig. 2 shows the output voltage and current waveforms in the more usual above-resonance mode of operation. In induction heating/melting and similar applications the heated work-piece equivalent electrical parameters are part of the resonant circuit. As the work-piece temperature increases, its equivalent resistance and inductance change, thus changing the circuit resonant frequency. Consequently, the deviation of the switching frequency from the resonant one is also changed, which results in undesired change of output power.

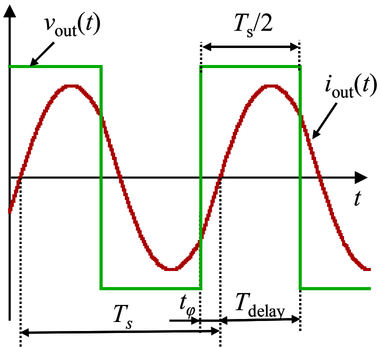


Fig. 2. Output voltage and current waveforms in above-resonance mode.

To maintain the desired output power the switching frequency needs to be adjusted. Control methods [10, 11] achieve this by adjusting the interval T_{delay} (after which T3 and T4 are switched off, and T1 and T2 switched on) according to (1).

$$T_{delay} = \frac{T_s}{2} - t_{\varphi,ref} = \frac{T_s}{2} - \varphi_{ref}\frac{T_s}{360°} \tag{1}$$

The new values of the positive half-period and consequently the new switching frequency are determined by the desired phase angle which makes this method a direct phase control one. This method has an advantage of being very simple and easy to implement with a low cost microcontroller. However, it has several disadvantages: the period (frequency) adjustments are made only in every positive half-wave; T_{delay} in (1) and the negative half-period are calculated using the period value T_s of the previous cycle; and as a consequence, the voltage pulses duty ratio is different than 0.5 during the adjustment period. For the phase angle time equivalent $t_{\varphi,ref}$ calculation in (1) the switching period T_s value is used as is in the analysis where the first harmonics of the output voltage and current are compared. In Fig. 2 the current waveform is comprised of pieces of damped oscillation and φ and t_φ depend on the circuit damping frequency. In order to make improvements to this method we first theoretically analyze the phase angle dependence on the deviation of the switching from the resonant/damping frequency.

III. Phase Angle Dependence Analysis

Analysis of the series-resonant converters usually use the voltage and current first harmonics to determine the circuit parameters and behavior. Assuming that the resistance in the circuit is small, the resonant frequency ω_0 is used in the calculations. To show the difference in these two approaches we first review the results with sine-wave voltage exaction and then elaborate the square-pulse excitation case.

A. Sine-wave excitation

When a series resonant circuit is excited by a sine wave voltage, all waveforms have the same shape and the current phase φ in respect to the voltage is a well known relation (2):

$$\varphi = \mathrm{arctg}[Q(\frac{\omega_s}{\omega_0} - \frac{\omega_0}{\omega_s})] \tag{2}$$

where ω_s is the switching, $\omega_0 = 1/(LC)^{1/2}$ the resonant angular frequency and the quality factor is $Q = \omega_0 L/R = 1/(\omega_0 RC)$.

The range of values for Q and ω_0 can be obtained using the real parameter values of the prototype resonant inverter for induction heating (used for method verification at the end of this paper) with rated power of 10 kW: $R = 0.24\ \Omega$, $L = 26.5\ \mu$H, $C = 26.6\ \mu$F. Using these parameters' values it is obtained that $\omega_0 = 37\ 664$ rad/s ($f_0 = 5\ 994$ Hz) and $Q = 4.16$. Table I summarizes ω_0, f_0 and Q values with typical R and L change of ±50% during metal-piece induction melting. Fig. 3 shows the phase φ change dependence on normalized switching frequency $x = \omega_s/\omega_0$ for three values for Q = 3, 4 and 5.

TABLE I. Rang of Change of Q with R and L Change ±50%

R (Ω)	$\Delta R/R$ (%)	L (μH)	$\Delta L/L$ (%)	ω_0 (rad/s)	f_0 (Hz)	Q	$\Delta\omega_0/\omega_0$ (%)	$\Delta Q/Q$ (%)
0.12	−50	13.25	−50	53 266	8 477	5.88	+41	+41
0.24	0	26.50	0	37 664	5 994	4.16	0	0
0.36	+50	39.75	+50	30 753	4 895	3.35	−18	−18

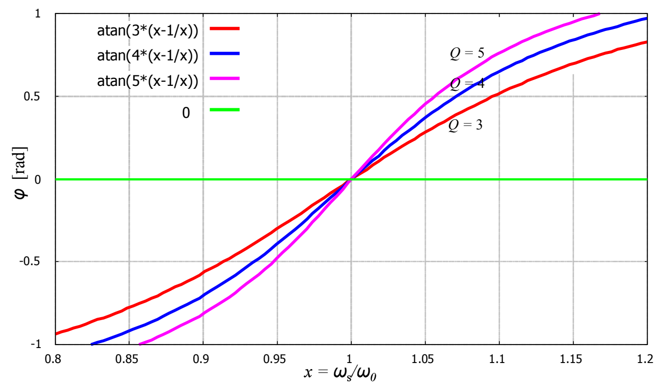


Fig. 3. Phase angle φ dependence on $x = \omega_s/\omega_0$ for Q = 3, 4 and 5, when the series resonant circuit is exited by a sinusoidal voltage.

B. Square pulse excitation

Let us now consider the series-resonant circuit excited by voltage pulses. If the voltage is in form of the Heaviside step function, then the current oscillates around zero with angular dumping frequency ω_d, as shown in Fig. 4. When the voltage has square pulses waveform (duty ratio D = 0.5) and amplitude $\pm V_{DC}$, then in every half-period the current is a piece of the dumped oscillation of Fig. 4 and looks like the waveform shown in Fig. 2. In the steady state the negative half-period waveform is symmetrical to the positive one in respect to the

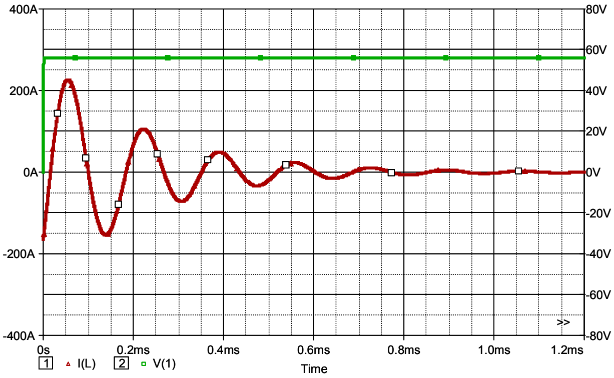


Fig. 4. Current waveform in the series-resonant circuit when excited by a Heaviside step voltage with amplitude V_{DC} = 56 V. Parameters' values are R = 0.24 Ω, L = 26.5 μH and C = 26.6 μF with initial values $i_L(0^+)$ = –165 A and $u_C(0^+)$ = –163 V to match the initial conditions in Fig. 2.

time axis. Harmonic analysis can be done in this case for the calculation of the active and reactive power, power factor etc. However, to determine the t_φ (and φ) as defined in Fig. 2 and its dependence on the deviation of ω_s from ω_d, the actual time waveforms from Fig. 4 have to be analyzed. Such analysis has not been carried in the literature to the best of our knowledge.

The series-resonant circuit current waveform for one half-period can be obtained from the second-order differential equation (4):

$$\frac{d^2 i(t)}{dt^2}+\frac{R}{L}\frac{di(t)}{dt}+\frac{i(t)}{CL}=\frac{1}{L}\frac{dv(t)}{dt} \tag{3}$$

The solution in this case is the under-dumped one ($\alpha < \omega_0$) since the current oscillates (Fig.4), it has no DC component since $v(t)$ is constant ($dv/dt = 0$) and has the form (4) or (5):

$$i(t)=e^{-\alpha t}(A_1\cos\omega_d t+A_2\sin\omega_d t) \tag{4}$$

$$i(t)=e^{-\alpha t}K\sin(\omega_d t-\varphi) \tag{5}$$

where $\alpha=\frac{R}{2L}=4528\mathrm{s}^{-1}$, $\omega_0=\frac{1}{\sqrt{LC}}=37665\frac{\mathrm{rad}}{\mathrm{s}}$, and

$$\omega_d=\sqrt{\omega_0^2-\alpha^2}=37392\frac{\mathrm{rad}}{\mathrm{s}} \tag{6}$$

Determination of the two constants K and φ for the steady-state solution can be done using two border conditions for this time interval, i.e. $i(0) = -I_0$ and $i(T_s/2) = +I_0$:

for $t = 0$
$$i(0)=K\sin(-\varphi)=-I_0 \tag{7}$$

for $t = T_s/2$
$$i(\frac{T_s}{2})=e^{-\alpha\frac{T_s}{2}}K\sin(\omega_d\frac{T_s}{2}-\varphi)=+I_0 \tag{8}$$

$$(8)\rightarrow\ e^{-\alpha\frac{T_s}{2}}K[\sin(\omega_d\frac{T_s}{2})\cos(-\varphi)+\cos(\omega_d\frac{T_s}{2})\sin(-\varphi)]=+I_0 \tag{9}$$

Dividing (9) by (7) we obtain:

$$\frac{e^{-\frac{\pi}{2Q}\frac{\omega_0}{\omega_s}}K[\sin(\pi\frac{\omega_d}{\omega_s})\cos(-\varphi)+\cos(\pi\frac{\omega_d}{\omega_s})\sin(-\varphi)]}{K\sin(-\varphi)}=-1 \tag{10}$$

$$e^{-\frac{\pi}{2Q}\frac{\omega_0}{\omega_s}}[\frac{\sin(\pi\frac{\omega_d}{\omega_s})}{\tan(\varphi)}-\cos(\pi\frac{\omega_d}{\omega_s})]=1 \tag{11}$$

and finally:

$$\varphi=\arctan\left(\frac{\sin(\pi\frac{\omega_d}{\omega_s})}{e^{+\frac{\pi}{2Q}\frac{\omega_0}{\omega_d}\frac{\omega_d}{\omega_s}}+\cos(\pi\frac{\omega_d}{\omega_s})}\right) \tag{12}$$

Relation (12) is graphed in Fig. 5 for Q = 3, 4 and 5 and shows considerably different behavior than (2) and Fig. 3. The function is not monotonous and it "oscillates" below resonance ($x = \omega_s/\omega_d < 1$) having negative, but also positive values for φ.

To verify this rather strange dependence, Fig. 6 gives PSpice simulation results of steady-state for several values of the switching frequency below and above resonance. Phase angles measured in these waveforms match and verify results obtained by (12). Also, the current waveform for $f_s = 0.5{\cdot}f_d$ or $f_s = 0.6{\cdot}f_d$ shows that it is very much distorted deep below resonance, the first harmonic is no longer dominant, which reflects to the amount of active power transferred to the load. This explains why below-resonance mode of power control is less desirable. The first diagram in Fig. 6 for f_s = 0,5 f_d shows that (12) gets zero values every time the switching period T_s is multiple of the dumping one T_d, in this case T_s = 2 T_d.

C. Comparison of the sine and the pulse excitation cases

A comparison of the phase angle φ dependence on ω_s in both cases, with sinusoidal and pulse excitation, is given in Fig. 7. The switching angular frequency ω_s is normalized, in the first case with the resonant ω_0, and in the second with the dumping angular frequency ω_d. The figure shows that there is a considerable difference, especially further away from the resonance point. However, making PSpice simulations and measuring the phase angle time equivalents t_φ, it was noticed that they have very close values in the above-resonance region as can be clearly seen in Fig. 8. Analyzing this fact lead us to a very interesting conclusion. Namely t_φ is calculated in a different way in both cases. When the circuit excitation is sinusoidal, the current is in the form:

$$i(t)=I_{max}\sin(\omega_s t-\varphi)=I_{max}\sin[\omega_s(t-t_\varphi)] \tag{13}$$

and
$$t_\varphi=\frac{\varphi}{\omega_s}=\varphi\frac{T_s}{2\pi} \tag{14}$$

In the second case with voltage pulses excitation, we have:

$$i(t)=e^{-\alpha t}K\sin(\omega_d t-\varphi)=e^{-\alpha t}K\sin[\omega_d(t-t_\varphi)] \tag{15}$$

and
$$t_\varphi=\frac{\varphi}{\omega_d}=\varphi\frac{T_d}{2\pi} \tag{16}$$

Relations (14) and (16) are similar, but profoundly different: ω_d and T_d are constants determined by the circuit parameters, while ω_s and T_s are variables that are changed by the control method and are used as x-axis in Figs. 7 and 8. This also shows that in the case with voltage pulses, the correct way to measure and calculate the phase angle is by using (16) which is not taken into account in many analysis and papers.

IV. New Method Development

The main objective of the control method is to adjust the switching frequency so that the desired phase angle and power transfer are obtained. To do so, the feedback circuit in Fig. 1

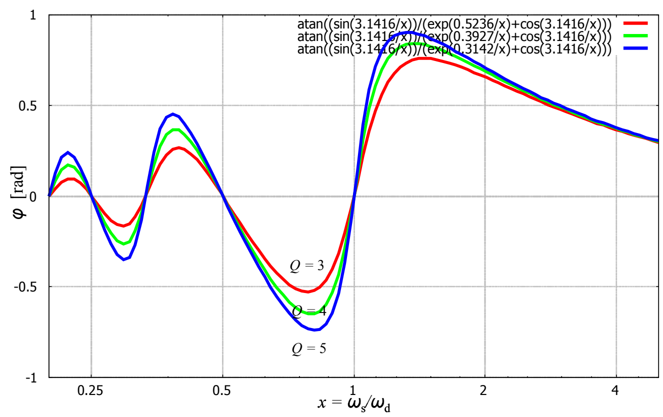


Fig. 5. Dependence of the phase angle φ on the normalized value ω_s/ω_d for $Q = 3$, 4 and 5, when excited by voltage pulses.

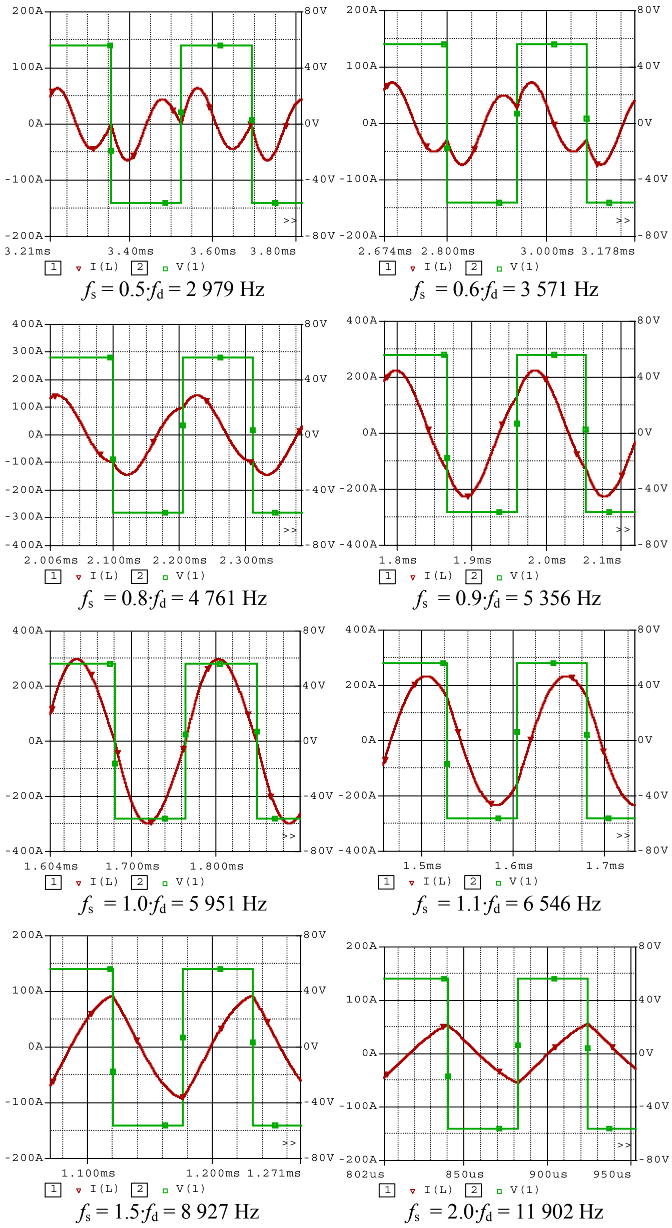


Fig. 6. Steady state voltage and current waveforms below and above resonance ($R = 0.24\ \Omega$, $L = 26.5\ \mu$H, $C = 26.6\ \mu$F and $Q = 4$).

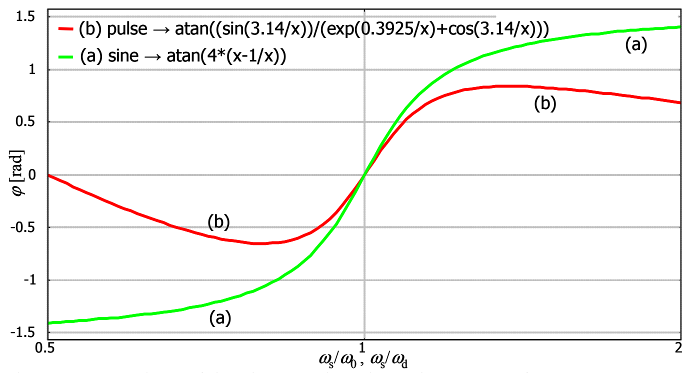


Fig. 7. Comparison of the phase angle φ dependence on ω_s for $Q = 4$: (a) sinusoidal excitation, $x = \omega_s/\omega_0$, (b) pulse excitation, $x = \omega_s/\omega_d$.

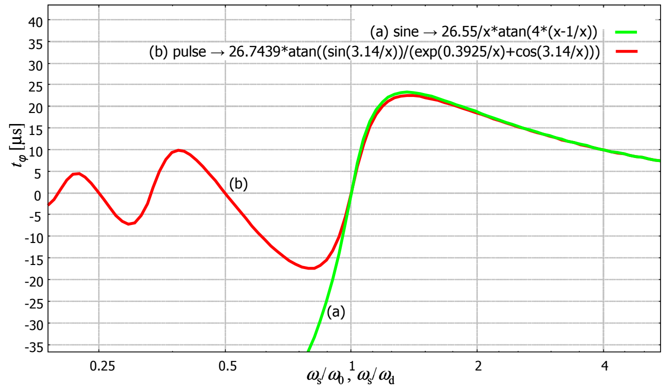


Fig. 8. Comparison of the phase angle time equivalent t_φ dependence on ω_s: (a) sinusoidal excitation, $x = \omega_s/\omega_0$, (b) pulse excitation, $x = \omega_s/\omega_d$.

has possibilities to measure the time between current zero crossings and instants when the switches are turned on or off. This means that the control method as input has the values of the previous cycle switching period $T_{s,i\text{-}1}$ and the current cycle phase angle time equivalent $t_{\varphi,i}$. Having these measured values, the method should determine the new switching frequency $T_{s,i}$ at which the φ_i and $t_{\varphi,i}$ have the desired or the reference values.

To determine $T_{d,i}$ knowing $T_{s,i\text{-}1}$ and $t_{\varphi,i}$, the implicit equation (18) can be used, that now has the form:

$$t_{\varphi,i} = \frac{T_{d,i}}{2\pi}\operatorname{arctg}\left(\frac{\sin(\frac{\pi}{T_{d,i}}T_{s,i-1})}{e^{+\frac{\pi}{2Q}\frac{T_{s,i-1}}{T_{d,i}}}+\cos(\frac{\pi}{T_{d,i}}T_{s,i-1})}\right) \tag{19}$$

Then $T_{s,i}$ is determined using now the implicit equation (12) and $T_{d,i}$ and φ_{ref} as known parameters:

$$\varphi_{ref} = \operatorname{arctg}\left(\frac{\sin(\frac{\pi}{T_{d,i}}T_{s,i})}{e^{+\frac{\pi}{2Q}\frac{T_{s,i}}{T_{d,i}}}+\cos(\frac{\pi}{T_{d,i}}T_{s,i})}\right) \tag{20}$$

Finally, the interval $T_{delay,i}$ is calculated using (1):

$$T_{delay,i} = \frac{T_{s,i}}{2} - t_{\varphi,i} \tag{21}$$

All these calculations should be solved numerically before the positive half-period ends, i.e. during T_{delay} time. Having in mind that $T_s/2$ is less than 100 μs and t_φ in the order of 10 μs, it is not possible to do the calculations with low cost microcontrollers, so the equations need to be simplified.

A. Linearization of phase angle equations

One common approach in electrical engineering and electronics to simplify nonlinear equations is linearization using derivates at the operating point (x = 1). This way a simple linear equation is obtained that matches the nonlinear curve very good around the operating point, and no so well away from that point. In the above-resonance region this linearization can be done on any of the two excitation cases since the curves almost overlap. Having in mind that the method uses time intervals (T_s and t_φ) the linearization is done on the $t_\varphi = f(T_s)$ function, resulting in (23) and graphically presented in Fig. 9.

$$t_\varphi = \frac{Q}{\pi} T_{\mathrm{d}} \left(1 - \frac{T_{\mathrm{s}}}{T_{\mathrm{d}}}\right) \tag{23}$$

Another way is to make linearization is to position the linear dependence in such a way so that there is smallest deviation from the curve on the entire range of t_φ change. To do so, the linear equation has to have a smaller slope, which can be achieved by multiplication by an additional coefficient $a < 1$, as in (24) and illustrated in Fig. 9 with the dashed line for $a = 0.75$ and the doted line for $a = 0.5$.

$$t_\varphi = a \frac{Q}{\pi} T_{\mathrm{d}} \left(1 - \frac{T_{\mathrm{s}}}{T_{\mathrm{d}}}\right) \tag{24}$$

B. Algorithm for the proposed direct phase control method

Based on the above analysis the method for the direct phase control consist of an algorithm with 8 steps:

1. Switch-off T1 and T2, and switch-on T3 and T4, (positive half-period starts, reset the $t_{\varphi,i}$ register);
2. Wait for current zero crossing moment $i(t) > 0$ and measure time interval $t_{\varphi,i}$;
3. Calculate $T_{\mathrm{delay},i}$ using following equations:

$$T_{\mathrm{d},i} = T_{\mathrm{s},i-1} + \frac{\pi}{Q} t_{\varphi,i}, \quad t_{\varphi,\mathrm{ref},i} = \frac{\varphi_{\mathrm{ref}}}{2\pi} T_{\mathrm{d},i},$$
$$T_{\mathrm{s},i} = T_{\mathrm{d},i} - \frac{\pi}{Q} t_{\varphi,\mathrm{ref},i}, \quad T_{\mathrm{delay},i} = \frac{T_{\mathrm{s},i}}{2} - t_{\varphi,i} \tag{25}$$

4. Wait for time interval $T_{\mathrm{delay},i}$;
5. Switch-on T1 and T2, and switch-off T3 and T4 (negative half-period starts, reset the $t_{\varphi,i}$ register);
6. Wait for current zero crossing moment $i(t) < 0$ and measure time interval $t_{\varphi,i}$;
7. Calculate $T_{\mathrm{delay},i}$ using (25);
8. Wait for time interval $T_{\mathrm{delay},i}$ and go to step 1.

The algorithm is graphically shown in Fig. 10.

V. VERIFICATION

To verify the improved method series of investigations have been done including mathematical calculations, PSpice simulations, code simulations in Proteus and hardware implementation. The method was tested in situations when there is abrupt change of L and R values, as well as with abrupt change

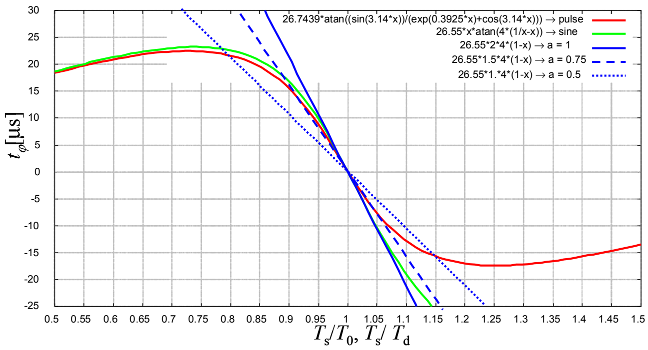


Fig. 9. Linearization of the phase angle time equivalent t_φ dependence on T_s.

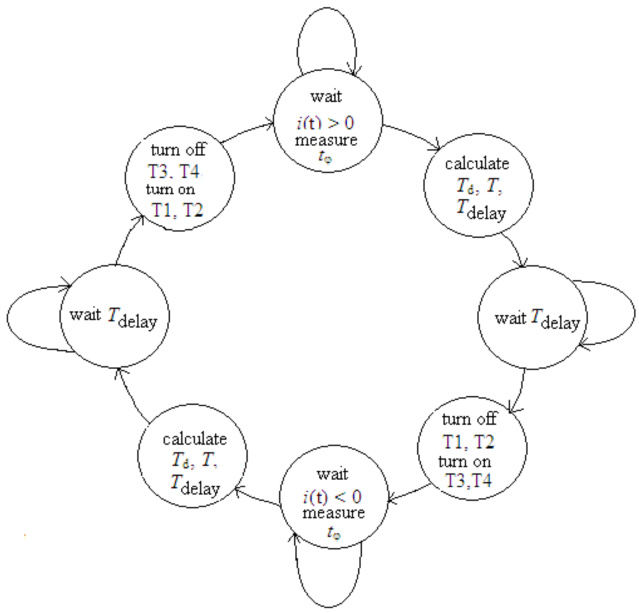


Fig. 10. Block diagram of the control algorithm.

of the reference angle φ_{ref}. Due to limited space in this paper only few of the simulations and experiments are presented.

Fig. 11 shows results for the change of φ_i obtained by a PSpice simulation of the method when there are simultaneous abrupt changes of the inductance value L from 26.5 μH to 31.5 μH and the resistance R from 0.24 Ω to 0.29 Ω, with the reference phase angle set to $\varphi_{\mathrm{ref}} = 5°$. Results obtained for the method in [10, 11] are shown on the same graph to make a comparison. The amplitude of the oscillation of φ_i with the new method is smaller, and also the transient interval is shorter. A comparison and verification when the reference angle φ_{ref} is abruptly changed from 5° to 35° is shown in Fig. 12. Again, the amplitude of oscillation of φ_i before it settles to the new value with the new method is smaller and the transient interval is shorter, too.

Practical verification was done with algorithm implementation on a PIC18F452 microcontroller and an IGBT full-bridge resonant inverter prototype shown in Fig. 13 [14, 15]. The prototype initially works with f_s = 1 268 Hz (T_s = 788 μs), resonant circuit parameters are R = 0.5 Ω, L = 315 μH, C = 55 μF (Q = 4.77) and the control circuitry is set to work with φ_{ref} = 22° (t_φ = 50.80 μs) to have output power of 40VA (Fig.14.a).

The inductor in this experiment is a coil with a moving ferromagnetic coil. When the coil is moved so that its inductance changes to $L = 426$ µH (35% change), the control algorithm successfully changes the switching frequency to $f_s = 1\ 207$ Hz ($T_s = 828$ µs) with $\varphi = 21.6^\circ$ ($t_\varphi = 58$ µs) (Fig.14.b).

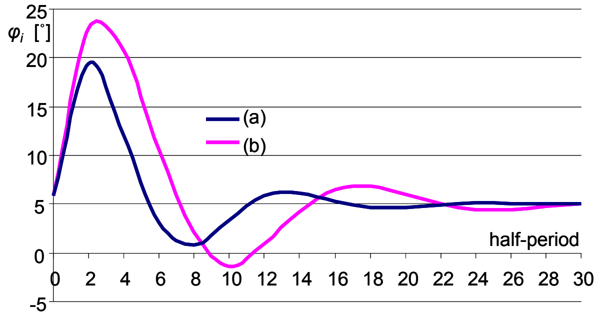


Fig. 11. PSpice results of the change of φ_i due to an abrupt changes of L from 26.5 µH to 31.5 µH and R from 0.24 Ω to 0.29 Ω whit $\varphi_{ref} = 5°$: (a) the new method and (b) the method in [10, 11].

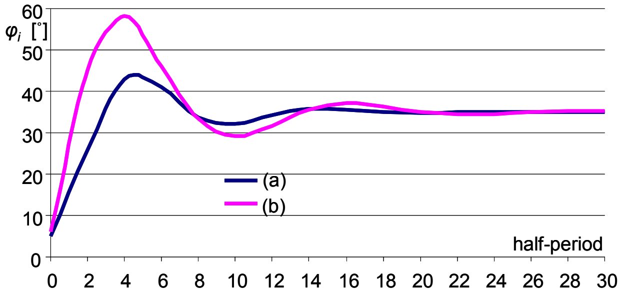


Fig. 12. PSpice results of the change of φ_i due to an abrupt change of φ_{ref} from 5˚ to 35˚: (a) the new method and (b) the method in [10, 11].

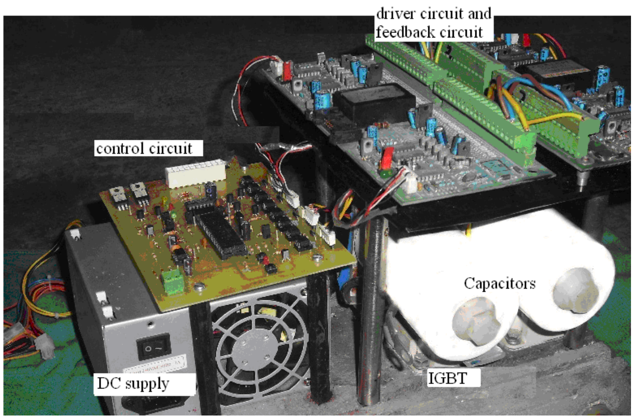
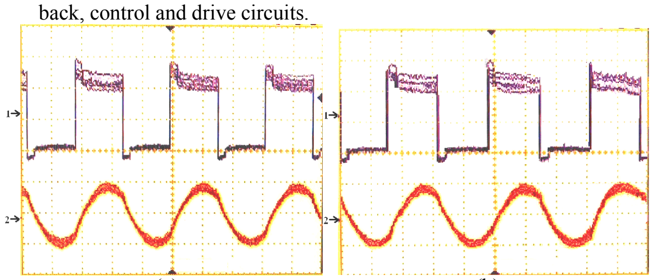


Fig. 13. Prototype of the series-resonant full-bridge inverter with the feedback, control and drive circuits.



(a) (b)

Fig. 14. Oscillograms of the prototype $v_{out}(t)$ and $i_{out}(t)$: (a) $L = 315$ µH with $T_s = 788$ µs, (b) after inductance change to $L = 426$ µH the period adjusts to $T_s = 828$ µs. Ch1: 5V/div, 250 µs/div; Ch2: 2A/div, 250 µs/div.

VI. Conclusion

The analysis of the series resonant circuit excited by a square pulse voltage, obtained from a full-bridge converter, shows a considerably different dependence of the current phase angle on the switching frequency than in the case of sinusoidal excitation. More over, the phase angle in this case has to be calculated in respect to the damping frequency, which depends on circuit parameters, rather then in respect to the switching one, which is constantly varied by the control method. This fact is usually neglected in many analysis and papers.

Phase control provides reliable drive of the series-resonant inverters in the presence of large dynamic changes in the load impedance. Based on above analysis, improvements to the direct phase control method are made. Linearization of the method equations has been made to facilitate its implementation with low cost microcontrollers. An algorithm for digital implementation and method verification with simulations and a prototype are presented at the end of the paper.

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It is a great pleasure to welcome you to EuroCon 2015 in the historical city of Salamanca (Spain), declared a UNESCO world Meritage Site in 1988 and European Capital of Culture in 2002.

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EUROCON 2015 received more than 250 technical submissions during several months. After a rigorous peer-review process, the International Program Committee selected 141 papers, which are published in this conference proceedings. The five finalists of the IEEE Region 8 student paper contest will also present their work during Eurocon 2015 Conference.

The selection of papers was extremely rigorous in order to maintain the high quality of the conference and we would like to thank the Program Committee for their hard work in the reviewing process. This process is very important to the creation of a conference of high standard and the EuroCon conference would not exist without their help.

The large number of submissions is certainly not only to testimony to the vitality and attractiveness of the field but an indicator of the interest in the EuroCon conferences themselves.

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