SPICE SIMULATION OF QUASI-RESONANT ZERO-CURRENT-SWITCHING DC–DC CONVERTORS

Indexing terms: Circuit theory and design, Switching and switching circuit, Convertors, Quasi-resonant, Simulation

A simple, topology-independent method is proposed for simulating the responses of open and closed-loop quasi-resonant DC–DC convertors operating in the zero-current-switching mode (ZCS-QRC). The method hinges on the substitution of the resonant network and switch assembly, fundamental in the realisation of such systems, by an equivalent circuit which represents its average behaviour. This permits simulation by a general-purpose electronic circuit simulator such as SPICE. The proposed approach is demonstrated by presenting the simulation results of a buck–boost convertor.

Introduction: Recently, a topology independent model for SPICE simulation of a voltage feedback, continuous mode PWM DC–DC convertor was introduced.[1] We propose an extension of this model to include quasi-resonant convertors operating in the zero-current-switching mode (ZCS-QRC) which have been gaining popularity. As has been shown by others,[2–4] the analytical expressions for the voltage and current waveforms, involve trigonometric functions, and the so-called 'DC' transfer functions (steady-state DC ratio) are topology-dependent. Consequently, the analytical expressions for the frequency domain small-signal transfer functions are complex. This computational complexity is partly to blame for the scarcity of published studies on the transient response and on procedures for the design of the control loop for ZCS-QRC.

Resonant switch model: Closer examination of ZCS-QRC reveals that they all rely on the basic resonant switched-inductor model (RSI) depicted in Fig. 1. It can be shown that the generalised, topology independent, average terminal currents and average capacitor voltage are related by the following equations:

$$I_a = I_d \frac{1}{2\pi}\frac{F_s}{F_0} F_{(x,n)} \tag{1}$$

$$I_b = I_d - I_a \tag{2}$$

$$V_c = V_{ab} \frac{1}{2\pi}\frac{F_s}{F_0} F_{(x,n)} + V_b \tag{3}$$

where

$$F_{(x,n)} = \frac{x}{2} + n\pi - (-1)^n \arcsin(x) + \frac{1}{x}[1 - (-1)^n\sqrt{(1-x^2)}] \tag{4}$$

when $n = 1$ for half-wave (HW) and $n = 2$ for full-wave (FW) operation

$$x = \frac{I_d Z_0}{V_{ab}}$$

$$Z_0 = \sqrt{\left(\frac{L_0}{C_0}\right)}$$

$$F_0 = \frac{1}{2\pi\sqrt{(L_0 C_0)}}$$

The main difference between the relationships given above and the equations reported in previous studies is the fact that our treatment is topology-independent and is therefore applicable to any ZCS-QRC configuration. Our treatment does not assume lossless elements when deriving the expression for x as was done in previous studies.[2–4] It is assumed that the resonant capacitor is AC grounded which is the most common situation. Applying eqns. 1–3, an average, topology-independent behaviour of the RSI of ZCS-QRC (Fig. 1) can be represented by the average model of Fig. 2. As in the case of the PWM model,[1] the dependent voltage and current sources (I_a, I_b, V_c) are nonlinear. Here we have an added computational complexity resulting from the fact that the relationships involve trigonometric functions which cannot be emulated by linear networks. Examination of the nonlinear functions $F_{(x,n)}$ for full- and half-wave operation, reveals that they are smooth and monotonic and can therefore conceivably be approximated by high-order polynomials. The advantage of such an approach is that polynomial dependent sources are compatible with most modern versions of SPICE.[1,5] Applying the least-square-fitting algorithm using the symbolic software MATHEMATICA,[6] we obtain the following approximations for $F_{(x,n)}$:

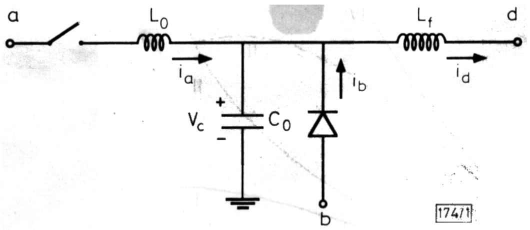

Fig. 1 *Resonant switched-inductor model*

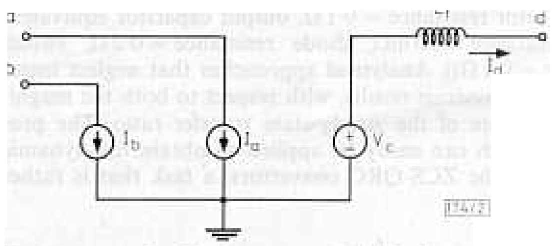

Fig. 2 *Averaged model of resonant switched-inductor*

For HW operation

$$P_{(x,1)} = 1{\cdot}99954x^{-1} + 3{\cdot}15753 + 0{\cdot}865323x + 0{\cdot}439751x^2 - 0{\cdot}575005x^3 + 0{\cdot}323202x^4 \tag{5}$$

For FW operation

$$P_{(x,2)} = 6{\cdot}28351 - 0{\cdot}0169485x^2 - 0{\cdot}0216974x^4 - 0{\cdot}0298489x^6 \tag{6}$$

It was found that the difference between the 'real' expressions and the polynomial approximated function was no larger than 0·025% over the range of interest ($0{\cdot}01 < x < 0{\cdot}99$). The boundary values of x are dictated by the following relationships:

$$x \in (0, 1) \tag{7}$$

$$\frac{F_s}{F_0} < \frac{2\pi}{F_{(x,n)} + \frac{x}{2}} \tag{8}$$

Eqn. 7 is a consequence of the basic requirement for the resonant current to reach zero level during the 'on' time (to permit zero current switching). Eqn. 8 is a manifestation of the requirement that the resonant capacitor should have sufficient time, during the 'off' period, to discharge.

Results and discussions: The model of Fig. 2, defined by eqns. 1–3 along with the polynomial approximations of eqns. 5 and 6, is compatible with most modern versions of computer circuit simulation programs such as SPICE.[5] Simulations test runs were carried out by GRAFSPICE[1] for a buck–boost convertor with the following parameters: $Z_0 = 1\,\Omega$, $L_f =$

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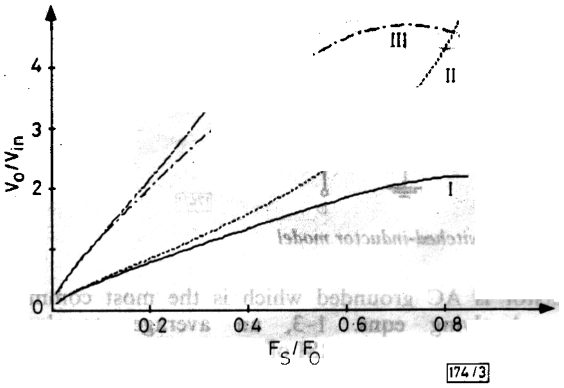

Fig. 3 *DC transfer ratio of ZCS-QRC buck–boost HW*
$Q = 5$ (I, II)
$Q = 20$ (III, IV)
Lossless (II, IV)
Parasitic resistances (I, III)

$100\,\mu$H, C_f(output capacitor) = 1 mF, $Q = R_0$(load resistance)/Z_0. The simulation for the 'DC' transfer ratio, based on the polynomial model (curves II and IV in Fig. 3), were found to be in good agreement with the results of previous studies[2,4] for lossless system. The simulation runs reveal that parasitic resistances can have a significant effect on the 'DC' transfer ratio (curves I and III in Fig. 3, which assume: main inductor resistance = $0{\cdot}1\,\Omega$, output capacitor equivalent series resistance = $50\,\text{m}\Omega$, diode resistance = $0{\cdot}2\,\Omega$, switch resistance = $0{\cdot}1\,\Omega$)). Analytical approaches that neglect losses can lead to erroneous results, with respect to both the magnitude and nature of the steady-state transfer ratio. The proposed approach can easily be applied to obtain the dynamic behaviour of the ZCS-QRC convertors, a task that is rather complex and almost impractical with previously published analytical approaches, if losses are taken into account. A test run of the power stage small-signal response reveals (Fig. 4) that the dynamic behaviour (magnitude and phase, which is not shown for the sake of brevity) of the full-wave and half-wave ZCS-QRC are remarkably different and that losses, caused by parasitic resistances, could have an important effect on the response. The proposed approach was tested successfully for the buck, buck–boost and boost topologies both in half-wave and full-wave operation, and was found to be in good agreement with previously reported results.[3] The preliminary analysis presented here suggests that whereas the full-wave ZCS-QRC response is similar to a conventional PWM convertor, the half-wave response resembles a current mode PWM convertor.[7,8] This similarity could be a manifestation of the shape of the $F_{(x,\,1)}$ function which represents, in fact, negative feedback between the average inductor current (a state variable) and V_c(Fig. 2).

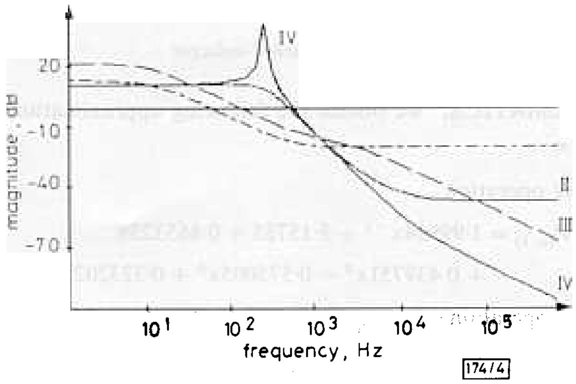

Fig. 4 *Small-signal-control to output response of buck–boost*
ZCS-QRC HW (I, III)
FW (II, IV)
$Q = 20$, $F_sF_0 = 0{\cdot}5$
Lossless (III, IV)
Parasitic resistances (I, II)

Conclusions: The proposed topology independent simulation method and the previously proposed model for PWM convertors[1] can be used in an interactive, trial and error simulation procedure to simplify and accelerate the design phase of switch mode systems.

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