Analysis and robust passivity-based control of zero-voltage switching quasi-resonant Cuk converter

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Abstract
Here, a general state-space average model of the half-wave zero-voltage switching quasi-resonant Cuk converter (HW-ZVS-QRCC) is derived and validated. The development of an extended Lyapunov-function-based robust passivity-based control strategy for HW-ZVS-QRCC is carried out using the derived model. The first feature of the proposed controller is that the resulted closed-loop HW-ZVS-QRCC is globally asymptotically stabilized even if the physical constraint on the magnitude of the control signal is taken into account. Also, to deal with the uncertainties in the conventional passivity-based control (PBC) design, an integral function of the output voltage error is incorporated in PBC structure such that zero-voltage regulation error is assured. Lastly, it shows which variables are most suitable for feedback purposes over a wide range of controller gains. Simulation and experimental results are also provided to validate the theoretical conclusions.

1 INTRODUCTION

Switched-mode pulse-width modulated (PWM) DC–DC converters are efficient power processors that are widely used in industries such as wind energy [1], fuel-cell vehicles [2], and photovoltaic power systems [3]. Having switching power losses and low power density at high frequencies have motivated power electronics researcher to work on soft switching converters [4]. Soft switching or resonant converters are composed of the conventional PWM circuits and soft switching strategies that result in zero-voltage switching (ZVS) or zero current switching (ZCS) [5]. There are many topological variations of the resonant converters. Amongst them, the resonant switch or quasi-resonant converter (QRC) has attracted much attention [6, 7].

The problem of regulating the QRCs has also attracted the attention of many control system researchers. Analysis and modelling of QRCs have been studied with great efforts and the non-linear average models of these converters have been derived using the general state-space average (GSSA) technique [8]. The GSSA model has been a basic tool for the analysis and design of linear and non-linear controllers for QRCs. In most cases reported in the literature, the voltage control is carried out using the small-signal model of the systems [9–11]. However, the small-signal models are not adequate to represent the converter’s behaviour during large transience. Thus, designing linear controller based on linearized averaged state-space equations of QRCs often fails to perform satisfactorily under large parameter, line and load variations. Moreover, the use of linear controller in such converters only ensures local stability of the system and does not support converter applications in a wide range of operating conditions.

On the other hand, it has been shown that the non-linear control of DC–DC converters results in better stability, faster voltage regulation and improved dynamic performances [12–15]. Among the non-linear control strategies, the passivity-based control (PBC) has received much attention due to its major advantages such as meeting any constraint on the magnitude of control signal and having high flexibility in its design [16]. The PBC strategy for QRCs of the boost and buck types was first proposed in [17] and it was shown to be rather useful in controlling the electrical systems. However, the non-linear control strategy in [17] assumes accurate knowledge of all converter parameters including line voltage and load resistance. Whereas, there exists non-idealities in real QRCs and precise values of the converter parameters are not always known. Therefore, it is not robust against the uncertainties of the system. To deal with the uncertainties in input voltage and load resistance, an adaptive strategy to compensate for them is given in [18]. Although, the...
adaptive strategy has a strong theoretical foundation, it has the disadvantage of leading to a rather complex controller with large volume of required calculations. To prevent the use of adaptive mechanism, parallel damped passivity-based control (PDPBC) of QRCs was proposed in [19]. It was shown that by proper design of PBC, the control law can be independent of the load resistance. However, no experimental result was provided to testify the control strategy and simulations were performed on the full-wave average model of QRCs to assess the effectiveness of the PDPBC strategy. By using the average models in simulations, the author ignores the non-ideal circuit factors such as parasitic elements of practical components and voltage drops associated with diodes. These assumptions are obviously not valid in real power converters. Our investigation shows that the application of PDPBC on actual systems of QRCs leads to a significant level of steady-state error and is not robust against the load variations. The second disadvantage of the aforementioned controllers for QRCs is that they cannot ensure global stability of these non-linear systems for large-signal perturbations.

Moreover, most of the PBC laws have been applied to second-order converters. However, filter addition to converters [20], enhancement of the voltage transfer gain [21], reduction of the ripple in voltage as well as current, improvement of system efficiency [22] and more have increased the circuit components and its equilibrium values are presented and validated using frequency- and time-domain techniques. Section 3 presents the derivation of the average model converter and its equilibrium values are presented and validated through simulations and experimental results are presented in Section 4. The conclusions are drawn in Section 5.

2 | MODELLING OF ZVS-QRCC

In this section, the GSSA model of half-wave zero-voltage switching quasi-resonant Cuk converter (HW-ZVS-QRCC) and its equilibrium values are presented and validated using frequency- and time-domain techniques. The ZVS-QRCC in half-wave mode finishes at $t_f$ when the diode $D_2$ is forward biased. This mode finishes at $t_f$ when the diode $D_1$ is forward biased. This mode finishes at $t_f$ when the diode $D_1$ is forward biased. This mode finishes at $t_f$ when the diode $D_1$ is forward biased.

2.1 | Generalized state-space average model

Figure 1 shows the topology of HW-ZVS-QRCC. According to GSSA technique, the input current $i_L$, the output current $i_{L_o}$, the diode voltage $v_d$, and the output capacitor voltage $v_C$ are considered as the state variables, while the resonant inductor current $i_{L_r}$ resonant capacitor voltage $v_r$ and the input voltage $E$ are considered as the input variables. Also, the voltage across the output load resistance $v_o$ is the output voltage. The operation of ZVS-QRCC can be divided into four modes. The formulation of each mode is as follows.

Mode 1) $C_r$ – Resonant capacitor charging mode [$t_0$, $t_1$]: The switch $SW$ is turned off at $t_0$. Resonant capacitor voltage $v_r$ rises linearly due to the input current. This mode finishes at $t_1$ when $v_r$ reaches $V_c = E + V_{o_r}$, where $V_c$ and $V_{o_r}$ are equilibrium values of $v_r$ and $v_o$, respectively. The reduced differential equations of HW-ZVS-QRCC for this mode of operation is:

$$
\begin{bmatrix}
\frac{di_{L_j}}{dt} \\
\frac{di_{L_o}}{dt} \\
\frac{dv_{C}}{dt} \\
\frac{dv_{C}}{dt}
\end{bmatrix} =
\begin{bmatrix}
0 & 0 & 0 & 0 \\
0 & 0 & 1 & 1 \\
0 & \frac{1}{C_r} & 0 & 0 \\
0 & 0 & 1 & \frac{1}{R_{C_r}}
\end{bmatrix}
\begin{bmatrix}
i_{L_j} \\
i_{L_o} \\
v_{C} \\
v_{C}
\end{bmatrix} +
\begin{bmatrix}
\frac{E - v_r}{R_{L}} \\
\frac{L}{R_{L}} - v_r \\
v_{o} \\
v_{o}
\end{bmatrix} .
$$

(1)

Since the storage components are large, during the steady-state operation, the state variables can be regarded as constants for each switching cycle. With this assumption, the resonant states are calculated as follows:

$$
v_r (t) = \frac{L}{C_r} (t - t_0) , i_{I_j} = i_{L_j} + i_{L_o},$$

$$i_{I_o} (t) = i_{L_o},$$

where $i_{L_j}$ and $i_{L_o}$ are equilibrium values of $i_{I_j}$ and $i_{I_o}$, respectively. The duration of this stage is given by

$$
T_1 = \frac{V_f}{I_f} C_r , \quad V_f = E + V_{o_r} .
$$

(2)

Mode 2) Resonant mode [$t_1$, $t_2$]: $L_r$ and $C_r$ start to resonate at $t_1$ when the diode $D_1$ is forward biased. This mode finishes at $t_2$ when the resonate voltage $v_r$ reaches zero. The reduced order
state equation for this mode is given by
\[
\begin{bmatrix}
\frac{diL_j}{dt} \\
\frac{diL_o}{dt} \\
\frac{dv_{Ct}}{dt} \\
\frac{dv_{Co}}{dt}
\end{bmatrix}
= \begin{bmatrix}
0 & 0 & 0 & 0 \\
0 & 0 & -\frac{1}{L_o} & 0 \\
0 & 0 & 0 & 0 \\
0 & 1 & 0 & -\frac{1}{RC_o}
\end{bmatrix}
\begin{bmatrix}
i_{L_j} \\
i_{L_o} \\
v_{Ct} \\
v_{Co}
\end{bmatrix}
+ \begin{bmatrix}
E - v_{cr} \\
\frac{L_q}{L_q} \\
\frac{L_q}{L_q} \\
\frac{L_q}{L_q}
\end{bmatrix},
\tag{4}
\]
with
\[v_{cr} (t) = I_F Z_a \sin \omega_a (t - t_1) + V_F, \]
\[i_{L_j} (t) = I_F \cos \omega_a (t - t_2), \]
and duration time of \(T_2\) which is given by
\[T_2 = \frac{\alpha (I_F, V_F)}{\omega_a}. \tag{6}\]

where
\[
\alpha = \sin^{-1} (\beta) + \pi, \omega_a = 2\pi f_s, f_a = \frac{1}{2\pi \sqrt{L_c C_r}},
\]
\[
\beta (i_j, v_f) = \frac{v_f}{Z_a}, v_f = v_{st}, i_j = i_{Lj} + i_{Lm} Z_a = \sqrt{\frac{L_q}{C_r}}.
\]

Mode 3) Resonant inductor charging mode [\(t_2, t_3\): This mode starts at \(t_2\) when the switch SW turns on under ZVS. This mode finishes when the value of resonant inductor current reaches the constant equilibrium value of the average output inductor current \(I_{Lq}\). The reduced order state equation for the mode is given by
\[
\begin{bmatrix}
\frac{di_{Lj}}{dt} \\
\frac{di_{Lq}}{dt} \\
\frac{dv_{Ct}}{dt} \\
\frac{dv_{Co}}{dt}
\end{bmatrix}
= \begin{bmatrix}
0 & 0 & 0 & 0 \\
0 & 0 & -\frac{1}{L_o} & 0 \\
0 & 0 & 0 & 0 \\
0 & 1 & 0 & -\frac{1}{RC_o}
\end{bmatrix}
\begin{bmatrix}
i_{Lj} \\
i_{Lq} \\
v_{Ct} \\
v_{Co}
\end{bmatrix}
- \begin{bmatrix}
\frac{E}{L_q} \\
\frac{i_{Lj}}{L_q} \\
\frac{i_{Lq}}{L_q} \\
\frac{i_{Lq}}{L_q}
\end{bmatrix}, \tag{7}
\]
with
\[v_{cr} (t) = 0, \]
\[i_{Lj} (t) = \frac{V_F}{L_q} (t - t_2) - I_{Lj} + I_F \cos \alpha. \tag{8}\]

The duration of this mode is given by
\[T_3 = \frac{I_{Lq} I_F (1 - \cos \alpha)}{V_F}. \tag{9}\]

Mode 4) Controlled mode [\(t_3, t_4\): This mode begins at \(t_3\) when the SW is still closed and the diode \(D_1\) becomes reversed biased. This mode finishes when the switch turns off again at \(t_4\). The reduced order state equation for this mode is given by
\[
\begin{bmatrix}
\frac{di_{Lj}}{dt} \\
\frac{di_{Lq}}{dt} \\
\frac{dv_{Ct}}{dt} \\
\frac{dv_{Co}}{dt}
\end{bmatrix}
= \begin{bmatrix}
0 & 0 & 0 & 0 \\
0 & 0 & -\frac{1}{L_o} & 0 \\
0 & 0 & 0 & 0 \\
0 & 1 & 0 & -\frac{1}{RC_o}
\end{bmatrix}
\begin{bmatrix}
i_{Lj} \\
i_{Lq} \\
v_{Ct} \\
v_{Co}
\end{bmatrix}
+ \begin{bmatrix}
\frac{E}{L_q} \\
\frac{i_{Lj}}{L_q} \\
\frac{i_{Lq}}{L_q} \\
\frac{i_{Lq}}{L_q}
\end{bmatrix}, \tag{10}
\]
with
\[v_{cr} (t) = 0, \]
\[i_{Lj} (t) = I_{Lj}. \tag{11}\]

The duration of this stage is given by
\[T_4 = T_s - T_1 - T_2 - T_3, \tag{12}\]
where \(T_s\) is the switching period. The differential Equations (1), (4), (7) and (10) can be described by
\[
\dot{x} (t) = A x (t) + B_i (t), i = 1, 2, 3, 4, \tag{13}
\]
where \(x = [i_{Lj}, i_{Lq}, v_{Ct}, v_{Co}]^T\). The duration of the four operation modes \(T_i\), where \(i = 1, 2, 3, 4\) are given by Equations (3), (6), (9) and (12). By defining \(d_i = T_i / T_s\), the GSSA model of the QRC can be characterize as follows [8].

Applying it to HW-ZVS-QRCC, the GSSA of the converter is obtained as follows.
\[
\begin{align*}
\frac{di_{Lj}}{dt} &= \frac{E}{L_q} - \frac{1}{L_q} u_f N (i_f, v_f) v_{di}, \\
\frac{di_{Lq}}{dt} &= -\frac{1}{L_o} v_{co} - \frac{1}{L_o} [u_f N (i_f, v_f) - 1] v_{di}, \\
\frac{dv_{Ct}}{dt} &= \frac{1}{C_q} u_f N (i_f, v_f) i_{Lj} + \frac{1}{C_q} [u_f N (i_f, v_f) - 1] i_{Lq}, \\
\frac{dv_{Co}}{dt} &= \frac{1}{C_q} i_{Lq} - \frac{1}{RC_o} v_{co}.
\end{align*} \tag{15}
\]
where
\[N (i_f, v_f) = \frac{1}{2\pi} \left[ \frac{\beta}{2} + \frac{1}{\beta} (1 - \cos \alpha) \right], \]
and \(n_f = f_s / f_a\) is the ratio of switching frequency \(f_s\) and resonant frequency \(f_a\). From GSSA Equation (15), we can obtain the GSSA equivalent circuit of HW-ZVS-QRCC as shown in
By letting the right-hand side of (15) to zero, the relationships between the constant equilibrium values of the average states are obtained as follows:

\[ I_{Ld} = \frac{V_o^2}{R E}, \quad I_{La} = \frac{V_o}{R}, \quad V_d = E + V_{ac}. \]

The result for the voltage conversion ratio of HW-ZVS-QRCC can also be obtained as

\[ \frac{V_o}{E} = \frac{1 - \frac{E}{J_s} N (I_f, V_F)}{\frac{E}{J_s} N (I_f, V_F)}, \]

where \( E \) is the equilibrium value of \( E \). It is observed from (18) that the voltage conversion ratio can be regulated by varying the switching frequency. However, it is obvious from (3) and (6) that in HW-ZVS-QRCC, the off-time of the switch \( T_{OFF} = T_1 + T_2 \) is predetermined by the system structure and the control law varies the on-time of the switch \( T_{ON} = T_s - T_{OFF} \) to regulate the output voltage.

It is important to note that the above equations are valid for an ideal converter. In real-life applications, the output load changes and there are power losses in various non-ideal parts of the circuit. Therefore, the switching frequency from (18) is ineffective in regulating the output voltage and will result in a regulation error. In other words, a closed-loop control scheme is needed to continuously monitor the output voltage and adjust the switching frequency accordingly.

2.2 Model validation using frequency-domain method

Let superscript \( \sim \) represent the small-signal variations around the operating point. Therefore, the variables of HW-ZVS-QRCC can be presented as

\[ i_{d} = I_{d}, i_{a} = I_{a}, v_{d} = V_{d}, v_{a} = V_{a}, f_{d} = f_{d}, F_{d} = F_{d}. \]

We also define \( E \) as the perturbation of \( E \) from its operating point and \( u = u \) as the control signal. It follows from (18) that the constant equilibrium value of control signal \( U \) is related to \( V_o \) by the relation \( U = E / V_o \). Plugging the system variable into (15) such that \( x = X + \Delta X \) and \( u = U + \Delta U \) yields the following perturbed dynamics.

\[
\frac{d\tilde{i}_{d}}{dt} = \frac{1}{L_d} \tilde{E} - \frac{1}{L_d} (V_o \tilde{u} + U \tilde{v}_{d} + \tilde{u}_{d}), \\
\frac{d\tilde{i}_{a}}{dt} = -\frac{1}{L_a} \left[ \tilde{v}_{a} + V_o \tilde{u} + (U - 1) \tilde{v}_{a} \right], \\
\frac{d\tilde{v}_{d}}{dt} = \frac{1}{C_v} \left[ I_{Ld} \tilde{u} + U \tilde{i}_{d} + I_{La} \tilde{u} + (U - 1) \tilde{i}_{a} + \tilde{v}_{d} \right], \\
\frac{d\tilde{v}_{a}}{dt} = \frac{1}{C_a} \tilde{v}_{a} - \frac{1}{R C_a} \tilde{v}_{a},
\]

where

\[ \tilde{u} = \frac{1}{2 \pi f_a} H J_f - \frac{F_f}{2 \pi f_a} J \tilde{i}_{d} - \frac{U}{2 \pi f_a} I_f \tilde{L}_a, \]

\[ f \left( I_f, V_f \right) = \frac{\beta}{2} - \frac{1}{\beta} \left( 1 - \cos \alpha \right), H = 2 \pi N (I_f, V_f). \]

It can easily be seen that the equilibrium point of (19) is the origin. To verify the accuracy of the GSSA model of the converter, the small-signal linearized model of the HW-ZVS-QRCC was obtained by linearizing (19) about the origin as follows.

\[
\frac{d\tilde{i}_{d}}{dt} = \frac{1}{L_d} \tilde{E} - \frac{1}{2 \pi L_d} \frac{F_f}{f_a} \left[ V_o H \tilde{f}_{d} - \tilde{v}_{d} \right], \\
\frac{d\tilde{i}_{a}}{dt} = -\frac{1}{L_a} \left[ \tilde{v}_{a} + V_o \tilde{u} + (H + f) \tilde{v}_{a} \right], \\
\frac{d\tilde{v}_{d}}{dt} = \frac{1}{C_v} \left[ I_{Ld} \tilde{u} + U \tilde{i}_{d} + I_{La} \tilde{u} + (H - f) \tilde{i}_{a} \right] + (H - f) \tilde{i}_{d}, \\
\frac{d\tilde{v}_{a}}{dt} = \frac{1}{C_a} \tilde{v}_{a} - \frac{1}{R C_a} \tilde{v}_{a},
\]

For the nominal parameters of the converter given in Table 1, the switching frequency to output capacitor voltage transfer function \( G_{s}(s) = v_o / f_a(s) \) and input voltage to output capacitor voltage transfer function \( G_{s}(s) = v_o / E(s) \) are

\[ G_{s} = \frac{-1256s^2 + 9.65 \times 10^7 s - 8.4 \times 10^8}{s^4 + 4203s^3 + 2.56 \times 10^8 s^2 + 7.15 \times 10^7 s + 3.3 \times 10^6}. \]
TABLE 1 Specifications of HW-ZVS-QRCC

| E_0 | L_i | L_o | C_i | C_o | C_r | L_r | R_0 | V_m (V_d) | F_s |
|-----|-----|-----|-----|-----|-----|-----|-----|----------|-----|
| 15 V| 2 mH| 2 mH| 220 μF| 440 μF| 50 nF| 200 μH| 100 Ω| 36 V | 13.58 kHz |

FIGURE 3 Magnitude Bode plot of the transfer function for ZVS-QRCC operating in half-wave mode for: (a) switching frequency to output capacitor voltage, (b) input voltage to output capacitor voltage

\[ G_E = \frac{-2.37 \times 10^9 s + 7.95 \times 10^{11}}{s^4 + 4203s^3 + 2.56 \times 10^6 s^2 + 7.15 \times 10^5 s + 3.3 \times 10^{11}}. \]  

(22)

To validate the small-signal linearized GSSA model of the converter, in Figure 3a,b, the Bode plots of the transfer functions in Equations (21) and (22) are compared to the Bode diagram obtained using the ideal converter circuit implemented in Simulink toolbox in MATLAB 2019b. The good agreement between the Bode plots validate the accuracy of the small-signal linearized GSSA model of the converter in Equation (20) for intermediate frequencies.

2.3 Model validation using time-domain technique

Next, to verify the accuracy of the GSSA model (given by (15)) in characterizing the transient behaviour of HW-ZVS-QRCC, the simulation results of GSSA equivalent circuit of HW-ZVS-QRCC of Figure 2 and the actual converter circuit are plotted in Figure 4. The same set of converter parameter values, given by Table 1, are used. The results show a good agreement between the average model and the actual electrical circuit. Since the large-signal model can predict both the steady-state and transient behaviour of the system, it can suitably be used to design a controller for HW-ZVS-QRCC.

3 CONTROLLER DESIGN

In this section, the RPBC for HW-ZVS-QRCC is presented and the stability of the closed-loop system is analysed.

3.1 Derivation of control law

The control objective is to enforce the average output capacitor voltage \( v_{\text{co}} \) to track a constant reference signal \( V_d \), that is, \( V_{\text{in}} = V_d \), despite the system parameter uncertainties in practical systems and non-ideality of the electrical components. To do the analysis in a global manner, the non-linear model of the converter presented in (19) is used to design an RPBC that globally stabilizes the converter. To eliminate the steady-state error in practical systems, we incorporate the integral function of the output capacitor voltage error \( z = \int (v_{\text{co}} - V_d) \, dt \) into the state-space equations of the converter. The presented integral approach compensates for the simplification of the average model and ensures asymptotic regulation under all parameter perturbations that do not destroy the stability of closed-loop system. Also, the input voltage is considered a constant by setting \( \Esad = 0 \). Now, the goal is to design a state feedback control law \( \hat{u} \), limited to the \([-U, N(l,p) - U] \) interval, that globally asymptotically stabilizes the state trajectories of the perturbed system (19) towards the origin.
The RPBC strategy is developed based on the Lyapunov stability theorem [16]. The derivation of a positive definite Lyapunov function for RPBC design is based on a combination of two positive semi-definite storage functions. The first storage function is proposed as the total stored energy in the perturbed system (19); that is,

\[
V_1(x) = \frac{1}{2} I_t^2 L_t + \frac{1}{2} I_o^2 L_o + \frac{1}{2} C_i \tilde{r}_{id}^2 + \frac{1}{2} C_o \tilde{r}_{od}^2.
\]  

(23)

To incorporate the integral function in RPBC design, a second storage function is proposed. The general form of the second storage function is as follows.

\[
V_2(x,z) = \frac{1}{2} (\tilde{a}_t L_t + \tilde{b}_t L_o + \tilde{c}_v d + d \tilde{x})^2,
\]

(24)

where \(d, b, c, a\) are constant values such that \(d \neq 0\). Noticeably, including the term \(v_{oa}\) in the second storage function is ineffective in finding a passive output for system (19). This is because the output capacitor voltage dynamics does not include the control input \(u\). Now, the aim is to define an output Lyapunov function for the RPBC design. The time derivative of \(V_2(x,z)\) along the trajectories of (19) is given by

\[
\dot{V}_2(x,z) = \eta(x,z) + \xi(x,z) \tilde{z},
\]

(25)

where

\[
\eta(x,z) = (\tilde{a}_t L_t + \tilde{b}_t L_o + \tilde{c}_v d + d \tilde{x}) \left( \frac{\alpha U}{L_t} + \frac{\epsilon (U - 1) \gamma}{C_t} \tilde{i}_{L_t} + \frac{b (U - 1)}{L_o} \tilde{i}_{L_o} \right) - \left[ \frac{\alpha U}{L_t} + \frac{b (U - 1)}{L_o} \right] \tilde{v}_{ct},
\]

and

\[
\xi(x,z) = (\tilde{a}_t L_t + \tilde{b}_t L_o + \tilde{c}_v d + d \tilde{x}) \left[ \frac{c \gamma}{L_t} \tilde{i}_{L_t} - \left( \frac{a}{L_t} + \frac{b}{L_o} \right) \tilde{r}_{oa} \right].
\]

It can be seen that some terms of \(\eta(x,z)\) do not have a defined sign in the whole domain. To provide a passive input–output characteristic for the system, it is essential to drive \(\eta(x,z)\) to zero by setting \(c = 0, b = dL_o\) and \(a = dL_o(U-1)/U\). The second storage function can be obtained by plugging the values of \(a, b\), and \(c\) into (24) as follows.

\[
V_2(x,z) = \frac{1}{2} \tilde{K} \left( I_t ((1 - U) \tilde{i}_{L_t} + I_o \tilde{i}_{L_o} + U \tilde{z}) \right)^2,
\]

(26)

where \(\tilde{K} = d^2/U^2\) is a positive constant. Consider the function \(V(x,z) = V_1(x) + V_2(x,z)\) as a continuous positive definite Lyapunov function for the RPBC design. The time derivative of \(V(x,z)\) along the trajectories of (19) is given by

\[
\dot{V}(x,z) = \left\{ -K \left( I_t ((1 - U) \tilde{i}_{L_t} + I_o \tilde{i}_{L_o} + U \tilde{z}) \right) + I_{c} \tilde{r}_{id} - \tilde{V}_{oa} \right\} \tilde{z} - \frac{1}{R} \tilde{v}_{oa}^2
\]

(27)

To apply a passivity-based control, consider

\[
y = -K_{v_{oa}} \left( I_t ((1 - U) \tilde{i}_{L_t} + I_o \tilde{i}_{L_o} + U \tilde{z}) + I_{c} \tilde{r}_{id} - \tilde{V}_{oa} \right)
\]

(28)

as the output which provides a passive input–output relationship for system (22), that is, \(V'(x,z) \leq yu\). It is evident from (27) that \(V'(x,z) \leq 0\) if the following condition is satisfied.

\[
\tilde{z} = -y \phi(y),
\]

(29)

where \(\phi(y)\) is any function such that \(\phi(y) > 0\). Substituting (29) into (27) yields

\[
V'(x,z) = -\frac{1}{R} \tilde{v}_{oa}^2 - K y^2 \phi(y) \leq 0.
\]

(30)

It is obvious from (30) that the derivative of the proposed Lyapunov function is negative semi-definite. Hence, we use Lasalle's invariance principle [16] to analyse the asymptotic stability of the closed-loop system. Since \(\phi(y) > 0\), we can conclude from (30) that the system can maintain the \(\dot{V}'(x,z) = 0\) condition where \(\tilde{r}_{oa} = 0\) and \(y = 0\). From (29), if \(y = 0\) then \(u^* = 0\). By studying the dynamics of the perturbed system from (19), for \(\tilde{r}_{oa} = 0\) and \(\tilde{u} = 0\) we get \(\tilde{i}_{L_t} = 0, \tilde{i}_{L_o} = 0\) and \(\tilde{r}_{id} = 0\). Also, if \(y = 0\), then \(z = 0\). Therefore, the largest invariant set is given by \((\tilde{x}, z) = 0\). It also proves that the closed-loop system is zero-state observable, that is, \(y = 0 \Rightarrow \tilde{u} = 0 \Rightarrow (\tilde{x}, z) = 0\). Thus, it can be concluded that system (19) is passive and zero-state observable with a radially unbounded positive definite storage function. Therefore, the origin is globally stabilized by the control law (29). Note that there is great freedom in the choice of feedback gain \(\phi(y)\). Here, to meet the constraint on the magnitude of frequency ratio, that is, \(0 \leq \nu_f \leq 1\), we choose the feedback gain as follows.

\[
\phi(y) = \begin{cases} 
\frac{U}{y} & \text{if } y \geq U/\nu_f \\
\frac{U - N(i_f, v_f)}{y} & \text{if } U - N(i_f, v_f) \leq y \leq U/\nu_f \\
\phi_{max} & \text{if } y < U - N(i_f, v_f) 
\end{cases}
\]

(31)

where \(\phi_{max} > 0\) represents the gain in non-saturated region of the control input. It is clear that the function (31) takes the constraint on the magnitude of the control signal into account, that is, \(u^*\) is limited to the \([-U, N(i_f, v_f) - U]\) interval. Adding the steady-state component of control signal to \(u^*\) renders

\[
u = K_{\phi_{max}} \left( I_t ((1 - U) \tilde{i}_{L_t} + I_o \tilde{i}_{L_o} + U \tilde{z}) + \phi_{max} [I_{c} \tilde{r}_{id} - \tilde{V}_{oa}] + \frac{I_{c}}{L_{id}} \right).
\]

(32)

Having obtained the control signal, the frequency ratio can be calculated from \(\nu = u/N(i_f, v_f)\). This completes the controller design. The key result is presented below.
Proposition 1: Consider the closed-loop system consisting of the non-linear average model (15), and the controller defined by the control law (32). For the given control input \( u = \eta T \), where \( 0 \leq \eta \leq 1 \), the controller globally asymptotically stabilizes the average model of the converter towards the equilibrium \([U_{dc}, I_d, V_{dc}, V_{ac}] = [V_d^2 / RE, V_d/R, E + V_d, V_d] \) for any \( 0 < R < \infty \) and \( 0 < E < \infty \). We now study the control strategies under parameter perturbations. The control law (32) comprises the steady-state components depends on the load perturbations. The control law (32) comprises the steady-state component of the converter towards the equilibrium parameters.

\[ u = K\Phi_m (L_v V_d e_1 + L_v E_0 e_2 + E_0 e_3) + \Phi_m (E_0 + V_d) (e_1 + e_2) \]
\[ - \Phi_m (V_d / R_0 E_0) e_1 + K\Phi_m (V_d / E_0 + V_d) e_1 e_3 \]
\[ + K\Phi_m L_v V_d (E_0 / (E_0 + V_d)) e_2 e_3 + K\Phi_m (E_0 / (E_0 + V_d)) e_3 \]
\[ + \frac{E_0}{E_0 + V_d}. \]

where
\[ e_1 = i_{d2} - \frac{V_d^2}{R_0 E_0}, e_2 = i_{d2} - \frac{V_d}{R_0}, \]
\[ e_3 = v_d - (E_0 + V_d), e_4 = v_0 - V_d. \]

The average error dynamics are obtained as follows.

\[ \dot{e}_1 = \frac{E}{I_d} - \frac{1}{L_d} u (e_3 + E_0 + V_d), \]
\[ \dot{e}_2 = -\frac{1}{L_0} (e_4 + V_d) - \frac{1}{L_0} (u - 1) (e_3 + E_0 + V_d), \]
\[ \dot{e}_3 = \frac{1}{C_0} (e_1 + \frac{V_d}{R_0 E_0}) + \frac{1}{C_0} (u - 1) (e_2 + \frac{V_d}{R_0}), \]
\[ \dot{e}_4 = \frac{1}{C_0} (e_2 + \frac{V_d}{R_0}) - \frac{1}{RC_0} (e_4 + V_d), \]
\[ \dot{\bar{e}} = e_4. \]

The equilibrium point of (34) is obtained by setting \( \dot{e}_1 = \dot{e}_2 = \dot{e}_3 = \dot{e}_4 = \dot{\bar{e}} = 0 \) and is given by (35). It can be seen from (35) that the inclusion of integrator in (32) forces the regulation error to be zero at equilibrium, that is, \( e_{\infty} = 0 \). Now, if \( R = R_0 \) and \( E = E_0 \), then \( (e_1, e_2, e_3, e_4, e_{\infty}, e_{\infty, \infty}) = 0 \).

Linearization of (34) about the equilibrium point gives the following system (see (36)).

\[ \frac{d\theta}{dt} = M\theta \]

\[ M = \begin{bmatrix} -\frac{1}{L_0} U_{in} V_d & -\frac{1}{L_0} U_{in} V_d & -\frac{1}{L_0} U_{in} V_d & 0 & -\frac{1}{L_0} U_{in} V_d \\ -\frac{1}{L_0} U_{in} V_d & -\frac{1}{L_0} U_{in} V_d & -\frac{1}{L_0} U_{in} V_d & 0 & -\frac{1}{L_0} U_{in} V_d \\ -\frac{1}{C_0} (U_{in} V_d + U) & -\frac{1}{C_0} (U_{in} V_d + U - 1) & -\frac{1}{C_0} U_{in} V_d & 0 & -\frac{1}{C_0} U_{in} V_d \\ 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 \end{bmatrix} \]

\[ U_{1a} = K\Phi_m L_v V_d \left( 1 + \frac{e_{\infty}}{E_0 + V_d} \right) + \phi_m (E_0 + V_d), \]
\[ U_{1b} = K\Phi_m L_v E_0 \left( 1 + \frac{e_{\infty}}{E_0 + V_d} \right) + \phi_m (E_0 + V_d), \]
\[ U_{1c} = K\Phi_m L_v V_d \left( 1 + \frac{e_{\infty}}{E_0 + V_d} \right) + \phi_m \left( \frac{E_0}{E_0 + V_d} \right), \]
\[ U_{1d} = K\Phi_m E_0 \left( 1 + \frac{e_{\infty}}{E_0 + V_d} \right) \]

where \( \theta = [\theta_1, \theta_2, \theta_3, \theta_4, \theta_5]^T \) and \( \theta_1 = e_1 - e_{\infty}, \theta_2 = e_2 - e_{\infty}, \theta_3 = e_3 - e_{\infty}, \theta_4 = e_4 - e_{\infty}, \theta_5 = e_{\infty} - e_{\infty, \infty} \). The closed-loop form of (34) will be asymptotically stable if all the eigenvalues of \( M \) lie in the left-half complex plane, that is, \( \theta \rightarrow 0 \). In other words, for any perturbations of \( R \) from \( R_0 \) and \( E \) from \( E_0 \), if \( M \) matrix is Hurwitz, the control strategy (33) leads to an asymptotically stable closed-loop system with zero output error.

To show the theoretical reason that motivates the integral term, we now study the PBC law without integrator. A passivity-based control law without integrator is based only on \( V_1(x) \) and is obtained by setting \( \bar{K} = 0 \) in control law (33) as follows.

\[ u = \frac{V_d^2 / RE - V_d^2 / R_0 E_0}{e_{\infty}} + \frac{V_d^2 / R_0 E_0}{e_{\infty}} + \frac{V_d^2 / R_0 E_0}{e_{\infty}} + \frac{V_d^2 / R_0 E_0}{e_{\infty}} = 0. \]
Substituting (37) into (34) and omitting the integrator from the state-space equation yields the closed-loop system with equilibrium point given by

\[
\epsilon_{1\infty} = \frac{\epsilon_4 + V_d}{E} \left( \frac{V_d}{R} + \frac{\epsilon_{4\infty}}{R} \right) - \frac{V_d^2}{R_0 E_0},
\]

\[
\epsilon_{2\infty} = \frac{V_d}{R} - \frac{V_d}{R_0} + \frac{1}{R} \epsilon_{4\infty},
\]

\[
\epsilon_{3\infty} = E - E_0 + \epsilon_{4\infty}, \quad U = \frac{E}{E + V_d + \epsilon_{4\infty}},
\]

(38)

where \(\epsilon_{4\infty}\) is obtained by solving the following equation:

\[
\phi_m (E_0 + V_d) \left[ \epsilon_{4\infty} + \frac{V_d}{E} \left( \frac{V_d}{R} + \frac{\epsilon_{4\infty}}{R} \right) \right.
\]

\[
- \frac{V_d^2}{R_0 E_0} + \frac{V_d}{R} - \frac{V_d}{R_0} + \frac{1}{R} \epsilon_{4\infty} \left. \right] - \phi_m \left( \frac{V_d^2}{R_0 E_0} + \frac{V_d}{R_0} \right) (E - E_0 + \epsilon_{4\infty})
\]

\[
+ \frac{E_0}{E_0 + V_d} - \frac{E}{E + V_d + \epsilon_{4\infty}} = 0.
\]

(39)

To show that the control strategy (37) leads to a stable system with output error, consider a HW-ZVS-QRCC with the nominal parameters given in Table 1. As will be explained in the following section, the control parameters in (32) and (37) are selected as \(K = 200\) and \(\phi_m = 4 \times 10^{-4}\). For example, the perturbed values of load resistance and input voltage are selected as \(R = 0.5 R_0\) and \(E = 0.8 E_0\). In case of PBC without integrator, by solving (39) using the given perturbed parameters, the steady-state value of output capacitor error is obtained as \(\epsilon_{4\infty} = -9.3\) V, that is, \(V_{oc} = 26.7\) V. Now, in case of RBPC, for the integrator \(\hat{z} = \epsilon_4\) to have a constant output at steady state, its input \(\epsilon_4\) must be zero, that is, \(\epsilon_{4\infty} = 0\). Figure 5 shows the simulation result for regulating the output capacitor voltage to \(V_{oc} = 36\) V with RPBC and PBC without the integral action. In both cases, the control laws are calculated using the nominal parameters given in Table 1, while the perturbed variables are \(R = 0.5 R_0\) and \(E = 0.8 E_0\). The simulation shows zero output error in the steady-state response with integral action, while the PBC without integrator fails to achieve \(V_{oc}\) in the presence of uncertain load and line voltage.

### 3.3 Tuning guidelines

Once the non-linear control law has been established, \(\phi_m\) and \(K\) will be selected to ensure that the switching converter performs satisfactorily in small-signal operation. The closed-loop poles of the system, which are the eigenvalues of \(M\), are given by \(m(j\omega) = |J_{5\times5}\cdot M| = 0\), where \(j\) is a complex variable. There are two control design parameters \(\phi_m\) and \(K\). Corresponding to each pair of \(\phi_m\) and \(K\), the five poles of the closed-loop system can be obtained. A heuristic approach to the selection of the controller gains is to fix one controller gain at a constant value while fine-tuning the other controller gain. To illustrate this approach, consider a HW-ZVS-QRCC with the nominal parameters given in Table 1. The three root loci for a fixed \(K\) and a range of \(\phi_m\) are shown in Figure 6: (a) the root loci for \(K = 200\) and \(10^{-5} \leq \phi_m \leq 2.6 \times 10^{-3}\) is displayed in Figure 6a; (b) the root loci for \(K = 100\) and \(10^{-5} \leq \phi_m \leq 2.6 \times 10^{-3}\) is shown in Figure 6b; (c) the root loci for \(K = 40\) and \(10^{-5} \leq \phi_m \leq 2.6 \times 10^{-3}\) is displayed in Figure 6c. The arrows show how the poles are moving for the increasing value of \(\phi_m\). It can be seen from Figure 6a that when \(K\) is in high value range, regardless of the parameter \(\phi_m\), the closed-loop system will have four complex-conjugated dominant poles with negative real parts which leads to an underdamped system. Figure 7a shows the corresponding output capacitor voltage response of the converter for \(K = 200\) and \(\phi_m = 2.6 \times 10^{-4}\). It can be seen from Figure 6b, as the value of \(K\) decreases, the root locus can reach point \(P_1\) for \(\phi_m = 2.55 \times 10^{-4}\) where the closed-loop system has three real and negative poles and two dominant complex-conjugated poles. The poles are \(-260, -150, -150\) and \(-90, -1500\). Figure 7b shows the corresponding output capacitor voltage response of the converter. It can be seen from Figure 6c that for low value of \(K\), the root locus can reach the point \(P_2\) for \(\phi_m = 0.4 \times 10^{-3}\) where the closed-loop system has three real and negative dominant poles and two complex-conjugated poles. The poles are \(-713, -77, -77, -120 + 1467i\) and \(-120 - 1467i\). In comparison to Figure 6b, the pair of complex-conjugated poles move farther away from the imaginary axis while real poles move toward the imaginary axis and become dominant. Figure 7c shows the corresponding output capacitor voltage response of the converter. It can be concluded from the results that lowering the value of \(K\) can reduce the oscillatory behaviour of the response and depress the overshoot. However, as we continue lowering the value of \(K\), the dominant real poles get closer to the imaginary axis which leads to a slower system response. Therefore, the choice of \(K\) is the compromise of system overshoot and speed.

Now, to study which variables are most suitable for feedback purposes, we can investigate the magnitude of the corresponding controller gain in (32). It can be seen that there is...
no proportional feedback of the output voltage error and the proportional gain of $r_{\nu_d}$ is relatively small in comparison to other controller gains. Therefore, for ease of design and implementation, a low-order linear controller which uses only the inductor current errors and the integral of the output voltage error can be used.

### 4 Simulation and Experimental Results

In this section, some simulations and experimental results are provided to demonstrate the performance of the proposed RPBC for the HW-ZVS-QRCC. The same set of converter parameter values as used in Section 2, given by Table 1, are used to obtain the results. The simulations are conducted on the actual system and not the average model, using the Simulink of MATLAB. Small parasitic resistances of $R_L = 0.22 \, \Omega$, $R_s = 0.22 \, \Omega$, and $R_{DS} = 0.2 \, \Omega$ were used in series with the inductors, capacitors, and power switch, respectively. Forward resistance and voltage drop of $R_{FD} = 0.05 \, \Omega$ and $V_{FD} = 0.85 \, V$, associated with the freewheeling diode, are also included in the simulation. Furthermore, to ensure ZVS conditions, the duty ratio of the control pulse is determined using the expression

$$D = 1 - T_{OFF} f_s.$$

Notice that in this case, the switching frequency is used to regulate the output voltage, while the duty ratio ensures the ZVS conditions for the QRC.

In order to show the merits of the proposed controller, the results are compared with that of the linear multiloop controller [10]. The multiloop control law for HW-ZVS-QRCC can be written as

$$u_f = \frac{E}{V_d N (I_F, V_F)} + K_p I_L + K_v V_c$$

(41)

where $K_p > 0$ and $K_v > 0$ are the current and voltage proportional gains, respectively. In this comparison study, the value of load resistance was changed from $R = 100 \, \Omega$ to $R = 50 \, \Omega$ at $t = 0.6 \, s$ and it restored to $R = 100 \, \Omega$ at $t = 0.9 \, s$. At nominal operating condition $R = 100 \, \Omega$, the switching frequency $f_s$ is 13 kHz. As the load changes to $R = 50 \, \Omega$, to maintain the output voltage at $-V_d$, the switching frequency varies to $f_s = 8.75 \, kHz$. Figure 8 shows the output voltage response for
Figure 8: Output voltage response of the regulated HW-ZVS-QRCC using the multiloop controller with a constant $K_p$ and various $K_i$s and using the proposed controller.

The output voltage response of the multiloop controlled converter is plotted for $K_p = 0.01$ and various $K_i$s, that is, $K_i = 0.1, 0.15, 0.6$, while the control parameters of RPBC were selected as $\phi_m = 0.4 \times 10^{-3}$ and $K = 40$ as calculated in the previous section. It can be seen that increasing the integral gain $K_i$ of the multiloop controller results in a short settling time and small variation in the output voltage response after the onset of the load disturbance. However, this also leads to an increased overshoot and oscillations at the start-up stage of the converter. Thus, it is evident that there exists a trade-off between the transient performances at the start-up stage and after the onset of the load disturbances. It can be seen from Figure 8 that both excellent transient response at the start-up stage and after the onset of load disturbances can be achieved using the proposed controller. Therefore, the proposed controller is more suitable for regulating the HW-ZVS-QRCC.

To validate the practical performance of the proposed method, an experimental platform for the voltage regulation of HW-ZVS-QRCC was built in the laboratory. The whole RPBC algorithm was implemented by the STM32F407 with a clock frequency of 168 MHz using the STM32CubeMX 5.4.0 and Keil uVision V5.24.2.0. For implementation, $v_{cm}$ and $v_o$ were sensed using the op-amp IC LM342N with a negative feedback gain of 0.05. The inductance currents were measured using the ACS712 5A module with a sensitivity of 185 mV A$^{-1}$. To convert the analog signal into digital signals, 10-b A/D converters with sampling periods of 7.5 $\mu$s were used to convert the analog signals into digital signals. The TC427 was used to drive the Power MOSFET IRFP260N and MUR1560 was used as the freewheeling diode. The values of $\phi_m$ and $K$ were $0.4 \times 10^{-3}$ and 40, respectively as calculated in the previous section. Figure 9 shows the experimental waveforms of the resonant capacitor voltage $V_{Cr}$ and the gate pulse $V_{GS}$ when the output load resistance is $R = 100 \, \Omega$. The duty cycle is tuned to a value given by (40) so that the resonant capacitor voltage becomes zero as the switch turns on. The measured off-time of the switch is $T_{OFF} = 17.5 \, \mu$s and the frequency ratio is $\omega = 0.155$. Figure 10 shows the output voltage responses of HW-ZVS-QRCC for both RPBC and multiloop controllers. Figure 10a shows the output voltage start-up response and the output response of the multiloop controlled system when load resistance was changed from $R = 100 \, \Omega$ to $R = 50 \, \Omega$ and then back to $R = 100 \, \Omega$. Here, to achieve an output voltage response with the same start-up overshoot as the proposed controller, the control parameters of the multiloop controller were selected as $K_p = 0.01$ and $K_i = 0.15$. As can be seen from Figure 10a, the overshoot and settling time of the load change are $\sim 25\%$ of $V_d$ and 0.4 s,
respectively. Figure 10b shows the start-up response and the output response for a change in load resistance from \( R = 100 \, \Omega \) to \( R = 50 \, \Omega \) (vice versa) obtained using the proposed controller. The settling time and overshoot of the start-up output voltage were \( \sim 0.2 \, \text{s} \) and \( \sim 10\% \) of \( V_{d0} \) respectively. As compared to the output given in Figure 10a, the settling time of the load change response was reduced to \( \sim 0.2 \, \text{s} \) and the overshoot of the load change response were reduced to \( \sim 16\% \) of \( V_{d} \) when the proposed controller was employed. The experimental results are in good agreement with the simulation results. Moreover, since the output voltage response of the proposed controller has smaller overshoots, it has more stability and robustness towards the load variations. We conducted the experiment for different load disturbances and found that the RPBC is able to regain for \( \sim 80\% \) increase in load while the multiloop controller can handle \( \sim 60\% \) variation.

Next, the ability of the proposed controller to handle input voltage disturbances was verified. Figure 11a,b shows the simulation and experimental results when the input voltage of the converter changes from \( E = 15 \, \text{V} \) to \( E = 20 \, \text{V} \) and then back to \( E = 15 \, \text{V} \). It is evident that for the changes in \( E \), the output voltage quickly restores to \( -V_{d} \) with a settling time of \( \sim 0.12 \, \text{s} \) and an overshoot ripple of \( \sim 10\% \). Noticeably, for the step changes of input voltage, the system response displays a non-minimum phase behaviour [23]. From Figure 11a, it can be seen, as the input voltage increases at time point of \( t = 0.4 \, \text{s} \), the output voltage experiences an undershoot behaviour and then increases to its maximum value before converging to \( -V_{d} \). Similarly, as the input voltage decreases at time point of \( t = 0.7 \, \text{s} \), the output voltage experiences an overshoot and then decreases to its minimum value before converging to \( -V_{d} \). This is the non-minimum phase behaviour and can be explained by inspecting the input-to-output voltage transfer function \( G_{E}(s) \) given by (21). The zero of \( G_{E}(s) \) is 335.44 and is in the right-half plane. Since \( G_{E}(s) \) has an odd number of real right-half plane zeros, the output voltage displays a non-minimum phase behaviour for step changes of the input voltage. Figure 11b validates the non-minimum phase dynamics of the closed-loop converter through experimental results. Also, it is evident from Figure 11b that the voltage response does not show a non-minimum phase behaviour if the input voltage changes slowly. All these results demonstrate that the proposed PBC maintains robustness and provides zero steady-state error.

5 | CONCLUSIONS

The modelling of a HW-ZVS-QRCC and its control using an RPBC strategy was addressed. The accuracy of the derived model of the QRC was verified via time-domain and frequency-domain techniques. The control design has been carried out using the derived model and is based on the storage function derivation and damping injection. Experimental results have demonstrated that, the proposed controller assures zero-voltage regulation error for the load and the input voltage variations as compared to the conventional PBC. Also, the generic structure of proposed controller makes the extension of the method to other types of QRCs relatively easy as compared to other nonlinear control techniques.

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