A robust passivity based control strategy for quasi-resonant converters

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Abstract
This paper presents a robust passivity-based control (RPBC) strategy for half-wave zero-voltage switching quasi-resonant (HW-ZVS-QR) buck and boost type converters. The proposed controller is based on energy shaping and damping injection. Theoretical analysis shows that the resulted closed-loop HW-ZVS-QRs are globally asymptotically stabilised even if the physical constraint on the magnitude of the control signal is taken into account. Also, an integral function of the output error is added into the feedback path of the conventional passivity-based control design. As a result, such controller not only offers a global asymptotic stable operation but also provides strong robustness to a wide range of parameter mismatch. It is also demonstrated that the proposed controller stabilises the converter in cases where the widely used linear multiloop controller fails. Finally, some simulation and experimental results comparing the performance of the proposed RPBC with that of the conventional linear multiloop controller are presented.

1 | INTRODUCTION

Resonant converters are defined as the combination of converter topologies and soft switching strategies that result in zero voltage switching (ZVS) or zero current switching (ZCS) [1]. The resonant converters have low switching losses and high-power densities which are widely used in industries [2, 3]. There are many topological variations of the resonant converters. Amongst them, the resonant switch or quasi-resonant converter (QRC) has attracted much attention [4, 5].

The three basic QRCs are buck, boost, and buck-boost configuration [6]. Analysis and modelling of these QRCs have been studied with great efforts. The generalised state-space averaging (GSSA) model has been a basic tool for the analysis and design of linear and non-linear controllers for QRCs [7]. Due to the non-minimum phase nature of the boost QRC, much attention has been directed at the control of these configurations. In most cases reported in the literature, the voltage control is carried out using the small-signal model of the systems [8–11]. The implementation of the traditional linear controller for the QRCs had been reported in [12]. The small-signal models are not adequate to represent the converter's behaviour during large transience. Thus, designing linear controller based on linearised averaged state-space equations of QRCs often fails to have satisfactory performance under large parameter, line and load variations. On the other hand, 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.

The non-linear nature of DC–DC converters has prompted some authors to investigate the application of various non-linear, namely, Lyapunov-function-based control strategies [13], passivity-based control (PBC) [14], sliding mode control [15]. PBC has high degree of flexibility in its design and can meet any constraint on the magnitude of control signal. Such properties make it highly suitable for control applications in DC–DC converters. The application of PBC strategy to QRCs of the boost and buck types was first proposed in [16] and it was shown to be rather useful in controlling these converters. However, the non-linear control strategy in [16] assumes accurate knowledge of all converter parameters including line voltage and load resistance. Whereas there exist 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 [17]. Although the adaptive strategy has a strong theoretical
foundation, it has the disadvantage of leading to a rather complex controller with a large volume of required calculations. To prevent the use of adaptive mechanism, parallel damped (PD) PBC of QRCs was proposed in [18]. 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 PD-PBC 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. The average model also ignores the ripple in the inductor current and the ripple in the output voltage. These assumptions are not valid in real power converters. Our investigation shows that the application of PD-PBC 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 the global stability of these non-linear systems for large-signal perturbations. Besides, the reported methods in literature fall short of a detailed study on the technical aspects of the implementation of non-linear controllers on QRCs, which is equally important for engineering practices. Therefore, designing a suitable robust passivity-based controller (RPBC) for QRCs is still an open problem which needs to be addressed.

As mentioned previously, the PBC provides a promising alternative to linear average controllers for QRCs. The main problem is the conceptional complexity of the control scheme which arises from the formulation of storage functions in design procedures of the PBC. The concern of this paper is to design a simple and systematic approach to the design of a Lyapunov-function-based RPBC strategy for QRCs. Both buck and boost QRCs are addressed. Of particular interest is the half-wave zero-voltage switching quasi-resonant (HW-ZVS-QR) boost converter due to its non-minimum phase nature. Unlike the conventional PBC strategies, it is proposed to incorporate the integral function of the output voltage into the conventional PBC design to reduce the steady-state error of the practical QRCs [19]. The integral approach ensures robustness under all parameter uncertainties that does not destroy the closed-loop stability of the system. Additionally, to show the effectiveness of the proposed strategy, the performance of the proposed controller is compared with that of the multiloop proportional-integral controller (MPIC) [9].

The rest of this paper is organised as follows. Section 2 presents the average model of boost QRC. The controller is designed in Section 3. The simulation and experimental results are presented in Section 4. The conclusions are addressed in Section 5.

2 MODEL OF THE QUASI-RESONANT BOOST CONVERTER

Figure 1 shows the topology of the HW-ZVS-QR boost converter. The operation modes of the converter and the corresponding state-space equations are all defined in [20]. The

GSSA model of this converter operating in continuous conduction mode can be written as [18]

\[
\frac{d\dot{i}_L}{dt} = \frac{E}{L} - \frac{v_o}{L} \dot{v}
\]

\[
\frac{dv_o}{dt} = -\frac{v_o}{RC} + \frac{i_v}{C} \dot{v}
\]

(1)

where

\[
v = uN(\dot{i}_L, v_o), u = f_s f_M f_o = \frac{1}{2\pi\sqrt{L_C}}
\]

\[
N(\dot{i}_L, v_o) = \frac{1}{2\pi} \left[ \frac{\varphi}{2} + \beta + \frac{1}{\varphi} (1 - \cos \beta) \right]
\]

\[
\varphi(\dot{i}_L, v_o) = \frac{v_o}{Z_n}, \beta = \sin^{-1} (-\varphi), Z_n = \sqrt{\frac{L_r}{C_r}}
\]

In (1), \(\dot{i}_L\) and \(v_o\) denote the average values of the inductor current and output voltage, respectively. \(f_s\) is the switching frequency, \(f_M\) is the resonant frequency and \(Z_n\) is the resonant circuit characteristic impedance. Also, \(v\) represents the control input and \(u\) stands for the frequency ratio of the square waveform applied to the switch “S" of the converter such that \(0 \leq u \leq 1\). For the half-wave mode, \(\pi < \beta < 3\pi/2\) and for the full-wave mode, \(3\pi/2 < \beta < 2\pi\). The positive constants \(R, C, C_0, L_n, L_r\) and \(E\) are load resistance, capacitance, resonant capacitance, inductance, resonant inductance and input voltage, respectively. By setting (1) to zero, the steady-state operating point of the converter can be obtained as

\[
i_s = \frac{v_s^2}{RE}, v_e = \frac{E}{v_e}
\]

(2)

where \(i_s, v_s\) and \(v_e\) represents the steady-state values of \(\dot{i}_L, v_o\) and \(v\), respectively.

The operation of HW-ZVS-QR boost converter can be divided into four modes. Each operating mode time interval is calculated as follows [21]

\[
T_i = C_i \frac{v_o}{i_s}
\]

(3)
where $T_s$ is the switching period. In an ideal condition, the input energy is equal to the output energy. Using the conservation of energy theory, the voltage conversion ratio of HW-ZVS-QR boost converter can be express as

$$\frac{E}{v_o} = UN (i_s, v_o)$$

(7)

where $U$ is the steady-state value of frequency ratio. It is observed that the HW-ZVS-QR boost converters are controlled with respect to the change in the switching frequency. However, from (3) and (4), it can be seen that 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.

The above equations are valid for an ideal converter. In real life applications, there are output load disturbance, variation in input voltage, and power losses in various non-ideal parts of the circuit. Therefore, the switching frequency from Equation (7) is ineffective in regulating the output voltage and will result in regulation error. It is also important to note that GSSA Equation (1) contains non-linearity in terms of sinusoidal functions of the states and multiplication of the states and the control input, which makes the system more non-linear than the conventional boost converter [20]. To design a linear controller, the system must be linearized about a desired operating point. However, as shown in [10], linearising the system in one operating point cannot represent a non-linear system over its entire operating trajectory. In the next section, an RPBC based on the non-linear model of the converter is developed which regulates the output voltage by adjusting the switching frequency, accordingly. The control objective is to enforce the average output voltage $v_o$ to track a constant reference signal $V_d^e > E$, despite the system parameters uncertainties.

3 | RPBC OF HW-ZVS-QR BOOST CONVERTER

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

### 3.1 Derivation of control law

To address the RPBC design, we study the perturbed dynamics of the HW-ZVS-QR boost converter. First, we introduce the perturbation variables as

$$i_L = i_s + i_δ, v_o = v_s + v_δ, u = u_s + u_δ$$

(8)

and assume

$$i_s \gg i_δ, v_o \gg v_δ, u_s \gg u_δ$$

(9)

Next, to eliminate the steady-state error, we incorporate the integral function of the output voltage error into the GSSA Equations (1). Substituting Equation (8) into the GSSA Equation (1) and defining $\xi = \int (v_o - V_d)dt$ yields the dynamics of the perturbed system as follows

$$\frac{di_δ}{dt} = -\beta \frac{v_s}{L_i} - \frac{v_o}{L_i} u_δ - \frac{1}{L_i} v_δ u_δ$$

$$\frac{d\xi}{dt} = \frac{v_o}{C} \xi - \frac{1}{RC} \xi - \frac{i_s}{C} u_δ + \frac{1}{C} \xi u_δ$$

(10)

$$\frac{dv_δ}{dt} = v_δ$$

(11)

It can easily be seen that the equilibrium of Equation (10) is the origin. The control objective is to globally asymptotically stabilise the state trajectories of the perturbed system towards the origin, i.e. $v_δ \to 0, i_δ \to 0$, with the control signal $u_δ$ limited to $[-v_o, N(v_s,v_o)-v_o]$ interval.

The RPBC strategy is based on the Lyapunov stability theorem. According to Lyapunov theorem, a system is globally asymptotically stable if there exists a Lyapunov function $V(x)$ that satisfies the following conditions [19]:

1. $V(0) = 0$
2. $V(x) > 0$ for $x \neq 0$
3. $V(x) \to \infty$ as $||x|| \to \infty$
4. $V^(x) < 0$ for all $x \neq 0$

where $x$ is the state vector of the system under consideration and $||x||$ denotes the norm of the state vector. For the perturbed system Equation (10), the state vector is $x = [i_δ, v_δ]$. The RPBC design satisfies the Lyapunov stability theory conditions in two phases: first, energy shaping which involves finding a differentiable continuous positive definite Lyapunov function. Second, damping injection so that the storage function will dissipate whenever $x$ is not identically zero. 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 proposed storage function is the total stored energy in the perturbed system; that is

$$V_1(i_δ, v_δ) = \frac{1}{2} L_δ i_δ^2 + \frac{1}{2} C_δ v_δ^2$$

(11)
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

\[ V_2(x) = \frac{1}{2} (a_i x + b v + c z)^2 \]  

(12)

where \(a\), \(b\) and \(c\) are constant values such that \(c \neq 0\). Now, the aim is to define an output \(y\) and choose the parameters \(a\), \(b\) and \(c\) such that the system Equation (10) becomes passive with respect to the output, i.e. \(V'(x) \leq y u_s\). To do that, the time derivative of Equation (12) along the trajectories of Equation (10) is calculated as follows

\[ \dot{V}_2(x) = \eta(x) + \zeta(x) u_s \]  

(13)

where

\[ \eta(x) = (a_i + b v + c z) \left( \frac{b}{C} i + (-\frac{a v}{L} - \frac{b}{RC} + c) v \right) \]

and

\[ \zeta(x) = (a_i + b v + c z) \left( \frac{b}{C} i - \frac{a v}{L} \right) \]

It can be seen that some terms of \(\eta(x)\) do not have a defined sign in the whole domain. Hence, to provide a passive input-output characteristic for the system, it is essential to drive \(\eta(x)\) to zero by setting \(b = 0\) and \(a = L v / u_r\). By plugging the values of \(a\) and \(b\) into (12), the second storage function is obtained as follows

\[ V_2(x) = \frac{1}{2} K \left( L i + u v \right)^2 \]  

(14)

where \(K = \frac{v^2}{u_s^2}\) is a positive constant. By defining the function \(V(x) = V_1(x) + V_2(x)\) as the Lyapunov function for RPBC design and taking the time-derivative of \(V(x)\) along the trajectories of Equation (10) we have

\[ \dot{V}(x) = -\frac{1}{R^2} + \left[ -K i_v \left( L i + u v \right) + i v - v i - v i \right] u_s \]  

(15)

considering

\[ y = -K i_v \left( L i + u v \right) + i v - v i \]  

the system is passive with respect to the output with a positive definite storage function, i.e. \(V'(x) \leq y u_s\). This completes the first phase of RPBC design. It is evident from Equation (15) that \(V'(x) \leq 0\) if the following condition is satisfied

\[ u_s = -y \phi(y) \]  

(17)

where \(\phi()\) is any function such that \(\phi(y) > 0\). Substituting Equations (17) into (15) yields

\[ \dot{V}(x) = -\frac{1}{R^2} - y^2 \phi(y) \leq 0 \]  

(18)

It is obvious from Equation (18) that the derivative of \(V(x)\) as the Lyapunov function candidate is negative semi-definite. Hence, we use Lasalle’s invariance principle [21] to analyse the asymptotic stability of the closed-loop system. Since \(\phi(y) > 0\), from \(V'(i, r, z) = 0\), we can conclude that \(r = 0\) and \(y = 0\). From Equation (17), if \(y = 0\) then \(u_s = 0\). By studying the dynamics of the voltage perturbation from (10), for \(r = 0\) and \(u_s = 0\) we get \(i_s = 0\). Also, if \(y = 0\), then \(z = 0\). Therefore, the largest invariant set is given by \(i_s = r = 0\). It also proves that the closed-loop system is zero-state observable, i.e. \(y = 0 \Rightarrow (i, r, z) = 0\). Thus, it can be concluded that the system Equation (10) is passive with a radially unbounded positive definite storage function and the required damping is injected by the function Equation (17). This completes the second phase of RPBC design. Since all the Lyapunov stability theory conditions are satisfied, the origin is globally asymptotically stabilised.

Note that there is great freedom in the choice of feedback gain \(\phi().\) Here, to meet the constraint on the magnitude of frequency ratio, i.e. \(0 \leq u \leq 1\), we choose the feedback gain as follows

\[ \phi (y) = \begin{cases} \frac{u}{y} & y \geq \frac{u}{\phi_m} \\ \frac{u}{y} - N (i, r) \phi_m \leq y \leq \frac{u}{\phi_m} \\ \frac{u}{y} - N (i, r) \phi_m \leq y \leq \frac{u}{\phi_m} \end{cases} \]

(19)

where \(\phi_m > 0\) represents the gain in the non-saturated region of the control signal. Adding the steady-state component of the control signal to \(u_s\) renders

\[ u = \frac{E}{v} + \phi_m \left( K L + v \right) + \phi_m \left( k u + v - i \right) \]  

(20)

This completes the controller design. Note that due to the presence of integral function in the control signal Equation (20), the average output voltage and the average inductor current in the steady-state are \(v = V_d\) and \(i = V_d^2 / RE\). After some mathematical manipulation, the control signal of the average model can be represented as

\[ u = \frac{E}{V_d} + \left[ V_d \phi_m \left( K L + 1 \right) \right] (i - V_d^2 / RE) \]

\[ + E K \phi_m \int (v - V_d) \, dt \]

\[ - \frac{V_d^2}{RE} \phi_m (v - V_d) + L K \phi_m (v - V_d) (i - V_d^2 / RE) \]

\[ + \frac{E}{V_d} K \phi_m (v - V_d) \int (v - V_d) \, dt \]  

(21)

where \(v = V_d^2 / RE\) and \(i = i - V_d^2 / RE\) are defined as the errors of the output voltage and inductor current, respectively. The key result is presented below.
Consider the closed-loop system consisting of the non-linear average model (1), and the controller defined by the control law Equation (21). For a given reference voltage $V_{d}$, such that $E < V_{d} < \infty$, the control input $u = N_1(\bar{v}_d, \bar{v}_e)\delta$, where $0 \leq u \leq 1$, globally asymptotically stabilises the average model of the converter towards the equilibrium $[\bar{v}_d, \bar{v}_e] = [N_1, V_{d}^2/RE]$ for any $0 < R < \infty$ and $0 < E < \infty$.

Remark. In practical situations, the load resistance $R$ and the line voltage $E$ are uncertain and their values may not be precisely known. Thus, the controlled switching frequency of the system depends on the nominal values of $R$ and $E$, say $R_0$, $E_0$ and is calculated as follows

$$f_s = \left[\frac{\bar{v}_{\text{nominal}}}{N_1(\bar{v}_d, \bar{v}_e)}\right] f_o$$

(22)

where $\bar{v}_{\text{nominal}}$ is given by setting $R = R_0$ and $E = E_0$ in Equation (21). The above proposition provides a stabilised equilibrium point with zero regulation error in the presence of parameter uncertainties. Under this modification, the control scheme can be illustrated as in Figure 2.

### 3.2 | Closed-loop small-signal analysis

Once the non-linear control law has been established, $\phi_m$ and $K$ will be selected such that the switching converter performs satisfactorily in small-signal operation. To do that, first we linearised the control law (20), resulting in

$$\bar{v}_{\delta,\text{Lin}} = \phi_m \left( (K+1) V_d \delta - \frac{V_d^2}{R E} \bar{v}_g + K E \bar{v}_e \right)$$

(23)

Plugging Equations (23) into (10) and neglecting second and higher-order terms of the small-signal perturbations yield the linearised form of the closed-loop system as follows

$$\dot{x} = Ax$$

(24)

where the matrix $A$ is given by

$$A = \begin{bmatrix}
-\frac{V_d^2 \phi_m}{L} (KL+1) & -E - \frac{V_d^2 \phi_m}{R E} & \frac{K E \phi_m V_d}{L} \\
\frac{E}{C V_d} + \frac{\phi_m V_d^3}{R E} (KL+1) & -1 & -\frac{V_d^4}{R E^3 C} \\
0 & 1 & 0
\end{bmatrix}$$

(25)

The closed-loop poles of the system, which are the eigenvalues of $A$, are given by the following characteristic equation

$$|\lambda I - A| = \lambda^3 + p_1 \lambda^2 + p_2 \lambda + p_3 = 0$$

(26)

where

$$p_1 = \frac{V_d^2 \phi_m}{L} (KL+1) + \frac{1}{RC} + \frac{\phi_m V_d^4}{R^2 E^2 C}$$

$$p_2 = \frac{V_d^2 \phi_m}{L} (KL+1) \left( \frac{1}{RC} + \frac{\phi_m V_d^4}{R^2 E^2 C} \right) - \frac{K E \phi_m V_d^2}{R E (KL+1)}$$

$$p_3 = -\frac{(KL+1) K E \phi_m^2 V_d}{R E C} + \left( \frac{E}{C V_d} + \frac{\phi_m V_d^3}{R E} (KL+1) \right)$$

The positions of the closed-loop poles can be changed by varying the values of control parameters, i.e. $K$ and $\phi_m$. Ideally, negative and real poles are desired so that the system response has no oscillation [20].

### 3.3 | Selection of controller parameters

Both the conventional linear and new developed non-linear controllers require tuning of control parameters in order to respond in a desired manner. However, the advantage of non-linear controllers over the conventional ones are their consistent dynamic behaviour over a wide range of operating conditions without having to retune the parameters. In [22–25] the root locus method is used to gain some insight in the roles of controller gains whereas some other researchers used the time-domain analysis to know how to choose the controller parameters [26–28]. Here, we study the effects of $K$ and $\phi_m$ on the system small-signal performances using the root locus method. This will help the designer to select the appropriate values for the controller parameters to achieve the desired output response. The design criteria for choosing $K$ and $\phi_m$ is to achieve a system with fast response. To illustrate the idea, the
following circuit parameters are used: $R_i = 100 \, \Omega$, $E_i = 15 \, \text{V}$, $L = 2 \, \text{mH}$, $C = 440 \, \mu\text{F}$ and $V_d = 36 \, \text{V}$. The three poles of (25) are plotted in Figure 3(a) for $10^{-4} \leq \phi_m \leq 10^{-1}$ and $40 \leq K \leq 80$. It is evident that the closed-loop system is stable. In Figure 3(a), the text labels for $K$ means that the whole segment of that root locus takes that $K$ value and it starts with $\phi_m = 10^{-4}$ and ends with $\phi_m = 10^{-1}$. It is seen from the root locus plot that with a fixed $K$, the pair of complex-conjugated poles first get farther from and then get closer to the imaginary axis when $\phi_m$ increases. The negative real pole of the system, however, gets farther from the imaginary axis when $\phi_m$ increases. Three negative real poles are desired for the closed-loop system to achieve a critically damped response. As the poles move farther from the imaginary axis, the system response gets faster and its stability gets better. It is seen that by tuning of the parameter $\phi_m$, the closed-loop system can achieve a dominant negative real pole that moves farther from the imaginary axis as $K$ increases. However, for large values of $K$ ($K = 80$), the closed-loop system will have a pair of complex-conjugated poles with negative real parts which leads to an underdamped system, regardless of the parameter $\phi_m$. Hence, the controller parameter $K$ is set as 70. Figure 3(b) shows the root locus plot for $10^{-4} \leq \phi_m \leq 10^{-1}$ and $K = 70$ in detail. It is seen from root locus plot in Figure 3(b) that for small values of $\phi_m$, the system has a pair of complex-conjugated poles in black and blue and a negative real pole in red. The root locus in black and blue reach the point $K_1$ at $\phi_m1 = 0.0010584$. Afterward, the three poles become real. By increasing the parameter $\phi_m$, the root locus in blue and red reach the point $K_2$ at $\phi_m2 = 0.0010593$. Afterward, the poles in blue and red become a pair of complex-conjugated poles while the pole in black is still real. Apparently, the three poles are real for $\phi_m \leq \phi_m \leq \phi_m2$. Here, the control parameters are selected as $K = 70$ and $\phi_m = 0.0010593$. With the listed nominal parameters, the poles of (25) are $-320, -248$ and $-238$. According to the given investigation on the effect of the controller gains on the position of the poles, 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 reduce the overshoot and settling time.

**Remark.** Although the RPBC of the HW-ZVS-QR boost converter has been presented, similar results can be established for the simpler HW-ZVS-QR buck converter. The average model of the converter is described by [18]

$$
\frac{d\dot{v}}{dt} = -\frac{1}{E} r_o + \frac{E}{L} uM (\dot{v})
$$

(27)

where

$$
M (\dot{v}) = \frac{1}{2\pi} \left\{ \frac{\alpha + \psi}{2} + \frac{1}{\psi} (1 - \cos \alpha) \right\}
$$

$$
\psi (\dot{v}) = \frac{Z_o \dot{v}}{E}, \alpha = \sin^{-1} (-\psi)
$$

Following the methodology used in establishing the result for HW-ZVS-QR boost converter, the following RPBC for the buck type converter can be obtained as

$$
v = \frac{V_d}{E} + E \phi_m (KL + 1) \left( \frac{V_d}{R} - \dot{v} \right)
$$

$$
+ E K \phi_m \int (V_d - r_o) \, dt
$$

(28)

The key result is presented below.

**Proposition.** Consider the closed-loop system consisting of the non-linear average model (27), and the controller defined by the control law (28). For a given reference voltage $V_d$, such that $V_d < E < \infty$, the control input $v = N_i(s, V_d)u$, where $0 \leq u \leq 1$, globally asymptotically stabilises the average model of the converter towards the equilibrium $[N_i, v_o] = [N_{i0}, V_d/R]$ for any $0 < R < \infty$ and $0 < E < \infty$.

## 4 SIMULATION AND EXPERIMENTAL RESULTS

In this section, some simulations and experimental results are provided to demonstrate the performance of the proposed
TABLE 1 Specifications of HW-ZVS-QR boost converter

| E   | L    | C    | C_r | R    | V_d |
|-----|------|------|-----|------|-----|
| 15 V| 2 mH | 440 μF| 200 μH| 50 nF| 100 Ω| 36 V |

FIGURE 4 Output voltage response (a) start-up followed by a step change in the load resistance from $R = 100 \, \Omega$ to $R = 50 \, \Omega$ (vice-versa), (b) for step changes in input voltage from $E = 15$ to $E = 20 \, V$ (vice-versa)

RPBC for the boost QRC. The specifications of the HW-ZVS-QR boost converter are given in Table 1. The simulations are conducted on the actual system 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 $R_F = 0.05 \, \Omega$ and $V_F = 0.85 \, V$, associated with the freewheeling diode, are also included in the simulation. The experiment was conducted using the STM32F407 board. The performance of the proposed controller is compared with that of the traditional MPIC. The MPIC for the HW-ZVS-QR boost converter is given as [9]

$$u = \frac{E}{V_d N \left( V_d^2 / RE, V_d \right)} + K_p \left( t - \frac{V_d^2}{RE} \right)$$

$$+ K_i \int (v - V_d) \, dt \quad (29)$$

As for the parameter selection of MPIC, in linear controllers, the overshoot of the start-up response of the multiloop controller can be improved by decreasing the integral gain $K_i$, but the settling time of load change response will increase. On the other hand, we can decrease the settling time of load change response by increasing $K_i$ but the overshoot will get worse which may even lead to destabilisation of the system. Therefore, to make a fair comparison and to ensure that the control ability of both controllers is at the same level, the parameters of the linear controller are tuned based on a good trade-off among overshoot during start-up and the speed of the load change response. Also, to ensure ZVS conditions, the duty ratio of the control pulse is determined using the expression:

$$D = 1 - T_{OFF} f_s \quad (30)$$

Notice that in this case, the switching frequency is used to regulate the output voltage, while the duty ratio ensures the ZVS condition for the QRC. For the HW-ZVS-QR with specifications given in Table 1, the calculated off-time of the switch is $T_{OFF} = 14.29 \, s$ and the frequency ratio is calculated as $u = 0.38$. 

FIGURE 5 Experimental test setup

FIGURE 6 Configuration of the experimental platform
4.1 Simulation results

In this section, some simulations were carried out using MATLAB Version R2019a. To obtain three negative poles for the closed-loop system, the parameters of RPBC were taken as $K_r = 70$ and $\phi_m = 0.0010593$. With these parameters, the poles are $-320$, $-248$ and $-238$. To obtain the best performance for MPIC, the parameters were selected as $K_p = 0.02$ and $K_i = 1.8$, respectively. Figure 4(a) plots the start-up response of the output voltage followed by a change in resistance for both controllers. During the simulation, we assume that the load decreases from $R = 100 \Omega$ to $R = 50 \Omega$ at $t = 0.2$ s and then rises back to $R = 100 \Omega$ at $t = 0.35$ s. The simulation result shows that, as compared to MPIC, the load change response of the RPBC converter is critically damped and has shorter settling times. Figure 4(b) displays the simulation result of the output voltage response when the input voltage of the converter changes from $E = 15$ V to $E = 20$ V at $t = 0.2$ s, and then back to $E = 15$ V at $t = 0.35$ s. In this case, as compared to MPIC, the RPBC enjoys a faster transient response without an oscillation at the expense of experiencing a larger overshoot. As expected, for both the step changes of output load and input voltage, the RPBC improves the dynamic performance of the system in most cases.

4.2 Experimental results

Next, to validate the practical performance of the proposed method, an experimental platform for the voltage regulation of boost QRC was built in the laboratory. The experimental test setup is shown in Figure 5. 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, the output voltage sensor is provided using a voltage divider circuit, $R_1$ and $R_2$, with a gain of 0.0435. The inductance current is measured using the ACS712 20 A module with a sensitivity of 100 mV/A. Two 10-b A/D converters with a sampling period of 7.5 $\mu$s are used to convert the analogue signal of the output voltage divider and current sensor module into digital signals. The gate pulse signal output of the STM32F407 is considered as the input of the low-side driver TC427. The TC427 is used for driving the Power MOSFET IRFP260N in high-frequency applications. Two diodes F10U60S and UF5400 are also used for the implementation.
The detailed configuration of the implementation strategy is shown in Figure 6.

The controller parameters used were: $K = 220$, $\phi_m = 0.0005$, $K_p = 0.05$ and $K_i = 2.2$. These values were selected so that both controllers achieve the best performances when they are operating in the nominal condition. Figure 7a and b shows the experimental waveforms of the resonant capacitor voltage $V_{cR}$, and gate pulse $V_{GS}$, when the output load resistance is $R = 100 \, \Omega$ and when it decreases to $R = 50 \, \Omega$, respectively. It is observed that in both cases, the duty-cycle is tuned to a value given by Equation (30) so that the resonant capacitor voltage becomes zero when the switch turns on. The measured off-time of the switch is $T_{OFF} = 16 \, \mu s$ and the frequency ratio is $u = 0.31$ which are very close to the calculated values. The small difference may be because of parasitics.

Next, the performances of RPBC and linear multiloop controller in handling variations in reference voltage are studied. To do that, the reference voltage step-changes between $V_d = 36 \, V$, $V_d = 25 \, V$ and $V_d = 50 \, V$, respectively. Figure 8(a) gives the output voltage response and the frequency ratio of the controlled system with linear multiloop controller. It can be seen that a large step increment in reference voltage from $V_d = 25 \, V$ to $V_d = 50 \, V$ leads to zero-switching frequency and destabilisation of the system. This is because as the voltage error becomes negative, the integral term used in controller’s design given by (29) causes the frequency ratio to decrease and reach zero after some time period. This can also happen during large step decrement in input voltage or load resistance. To avoid this, a non-linear saturation/limiter block along with an anti-windup controller [29] should be used to prevent the MPIC from driving the switching frequency to zero. Figure 8(b) shows the output voltage response and the frequency ratio of the converter controlled with RPBC. It can be seen, as compared to Figure 8(a), as the reference voltage step changes from $V_d = 25 \, V$ to $V_d = 50 \, V$, the frequency ratio converges to $u = 0.16$ and drives the output voltage to its desired reference. It can be observed that unlike the linear controller, the RPBC is capable of handling a large step change in reference voltage and the zero-switching frequency is expected to be a less frequent occurrence. This validates the performance of RPBC for wide operating ranges as compared to the conventional linear multiloop controller.

Finally, the dynamical property of the proposed RPBC in starting-up and handling of disturbances is compared to that of the conventional linear multiloop controller. Figure 9(a–c) shows the experimental waveforms of linear multiloop controlled boost QRC. Figure 9(a) gives the waveforms of the converter during start-up using a reference voltage of $V_d = 36 \, V$. The output voltage response has a settling time of 144 ms with an overshoot of 4.6 V and steady-state frequency ratio of $u = 0.31$. Figure 9(b,c) shows the ability of MPIC in handling the variations in load resistance and input voltage.

Figure 10(a–c) shows the experimental waveforms of the boost QRC controlled with RPBC. Figure 10(a) displays the waveforms of the converter during start-up using a reference voltage of $V_d = 36 \, V$. It can be seen, as compared to Figure 9(a), the output voltage has no oscillation during start-up and has comparatively less settling time of 80 ms with an overshoot of 4.6 V and steady-state frequency ratio of $u = 0.31$. Figure 10(b) gives the output voltage and the corresponding frequency ratio when the load resistance changes. It can be seen that as the load changes from $R = 100 \, \Omega$ to $R = 50 \, \Omega$, the output voltage experiences an undershoot of 4.5 V and takes about 88 ms to track its reference. The steady-state value of frequency ratio becomes $u = 0.16$ while its calculated value using (7) is 0.27. As the load resistance changes back to $R = 100 \, \Omega$, the output voltage experiences an overshoot of 5.6 V and takes about 68 ms to track its reference. It can be seen, as compared to the conventional MPIC in Figure 9(b), after the onset of load changes, the output voltage was rapidly restored to the desired voltage without oscillation. Figure 10(c) shows the output voltage and the corresponding frequency ratio when the input voltage was changed. It can be seen that as the input voltage changes from $E = 15 \, V$ to $E = 20 \, V$, the output voltage experiences an overshoot of 5 V and takes about 102 ms to track its reference. The steady-state value of frequency ratio becomes $u = 0.49$ while its calculated value using Equation (7) is 0.39. As the input voltage changes back to $E = 15 \, V$, the output voltage experiences an undershoot of 4 V and takes about 128 ms to track its reference. It can be seen, as compared to Figure 9(c), the output voltage and frequency ratio for a step change in load resistance from $R = 100 \, \Omega$ to $R = 50 \, \Omega$ (vice versa); (c) output voltage response and frequency ratio for a step change in input voltage from $E = 15 \, V$ to $E = 20 \, V$ (vice versa).
voltage enjoys a faster transient response with no oscillation at the expense of higher overshoot and undershoot ripples. The performance comparison between RPBC and MPIC in terms of maximum output voltage raise/drop relative to the reference voltage and settling times are tabulated in Table 2. It is clear that the experimental results are in good agreement with the simulation results shown previously.

5 | CONCLUSIONS

This paper has addressed the voltage regulation problem of HW-ZVS-QR buck and boost converters using an RPBC strategy. The proposed controller assures zero voltage regulation error and guarantees the global stability of QRCs even if the physical constraint on the magnitude of control signal is taken into account. The transient responses of QRCs under step changes in the load, in the input voltage, and in the reference output voltage are analysed. With simulations and experimental studies, it has been shown that, compared to the conventional linear controller, the proposed controller not only offers global stability but also leads to a faster dynamic response and strong system robustness. Moreover, it is shown that the proposed controller can reduce the occurrence of zero-switching frequency as compared to the MPIC. Also, the systematic approach to the design of RPBC, makes the extension of the method to high-order QRCs relatively easy as compared to other non-linear control methods. On the other hand, the main advantage of the traditional multiloop controller is that the design procedure of linear controller is simpler and the implementation circuits are relatively easy.

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