Frequency-Modulation Controlled Load-Independent Class-E Inverter

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ABSTRACT This paper proposes the frequency-modulation (FM) controlled load-independent class-E inverter, which has robustness against simultaneous variations of the load and the components in the output resonant filter. The main idea is to apply the FM control of the output-voltage regulation to the load-independent class-E inverter. By applying that, the proposed inverter can keep the phase shift between the driving signal and the output voltage constant without using any time information. As a result, the proposed inverter can maintain the constant output voltage and high power-conversion efficiency operation at high frequencies despite the load variations and the component tolerances in the output resonant filter. We give analytical expressions of the proposed inverter and quantitative evaluations. Additionally, an experimental prototype of the proposed inverter was implemented. The theoretical and experimental results showed the validity and effectiveness of the proposed inverter. The implemented inverter achieved 95% efficiency with the 5.7 W output and 1-MHz operating frequency at the rated operation.

INDEX TERMS Class-E inverter, load-independent operation, FM control, load variation, component tolerance.

I. INTRODUCTION

The class-E inverter [1]–[17], [21]–[33] is one of the high-frequency resonant inverters, which can achieve high power-conversion efficiency due to the class-E zero-voltage and zero-derivative switching (ZVS/ZDS). The class-E inverter has a switching device whose source terminal is grounded. Therefore, it is advantageous that no floating-gate driver is necessary compared with bridge-type inverters, especially at high frequencies. Therefore, the class-E inverter is expected to apply to various high-frequency applications, such as the dc-ac inverter part of the resonant converters, power-factor correction converters [1], [2], and transmitter part of wireless power transfer (WPT) systems [3]–[11]. In such modern applications, the class-E inverter is required to maintain the constant output voltage against the load variation at the megahertz frequency range. When the load variation occurs in the class-E inverter, the switch-voltage waveform varies drastically, which does not satisfy the ZVS condition.

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Namely, the output voltage varies, and the power-conversion efficiency deteriorates seriously. Besides, it is also required to address the issue of the output resonant-filter component variations. The changes of the resonant capacitance and inductance occur because of component tolerances or aging deteriorations. Therefore, the variation range is limited and the high-speed response is unnecessary, which are different from the load variation.

The load-independent class-E inverter [11]–[14] achieves constant output voltage, ZVS, and fixed phase shift between the driving signal and the output voltage at any load resistance without any control. This inverter attracts much attention recently [9]–[15] because of the appearance of the suitable applications, such as the WPT systems and the high-frequency resonant converters. On the other hand, the load-independent class-E inverter has a weakness against the component tolerance in the output resonant filter as well as the original class-E inverter. The class-E inverter usually has a high-Q output-resonant filter. Therefore, the output voltage and the power-conversion efficiency are highly sensitive to the component values in the output resonant filter.
By the way, the phase shift between the driving signal and the output voltage is determined uniquely in the nominal-operation class-E inverter. The self-tuned class-E inverter [15]–[17] uses this characteristic for achieving high power-conversion efficiency against the component tolerance in the output-resonant filter. The self-tuned class-E inverter adjusts the operating frequency to keep the nominal phase shift, making the inverter satisfy the class-E ZVS/ZDS conditions and the rated output voltage automatically. However, it is not very easy to implement a feedback network at the megahertz operation, in particular. This is because it is necessary to measure and adjust the propagation delays of feedback-network components [17].

There are the same issues in the LLC converter, which is a typical resonant converter. The LLC converter with automatic resonant frequency tracking (ARFT) [18]–[20] adjusts the operating frequency to keep the phase shift constant against the resonant-filter component tolerances. In [18], the frequency-modulation (FM) control is performed by detecting the zero-crossing point of the diode current, namely time information of the voltage like the self-tuned class-E inverters. In [19], [20], it is proposed that the frequency is adjusted for keeping the amplitude ratio of the two kinds of voltage or current constant. The phase shift between the switching timing and the resonant current can be kept constant against the resonant frequency variations by keeping the ratio. Therefore, the ARFT LLC converter in [19], [20] can achieve the constant phase shift and the low switching losses against the resonant-component tolerance. Besides, no time information is used in the control, the propagation delays in the feedback network are unnecessary to consider.

This paper proposes the FM-controlled load-independent class-E inverter. The FM control regulates the output voltage constant with the prerequisite of no input variation. With the similar operation mechanism to the ARFT LLC converter in [19], [20], the proposed inverter has the self-tuning function without using any time information. Besides, the proposed inverter works on the load-independent operation, even with output resonant-filter component tolerances because the phase shift can be kept constant. Namely, the proposed inverter has robustness against simultaneous variations of load and the output resonant filter components, which achieves constant output voltage with high power-conversion efficiency. This paper gives the analytical expressions of the waveforms and the power-conversion efficiency of the proposed inverter. The validity and effectiveness of the proposed inverter were confirmed from the experimental and theoretical results. It can be stated that we succeeded in loading both the load-independent operation and the self-tuning function onto the class-E inverter simultaneously, similar to the ARFT LLC converter.

II. BACKGROUND
A. ORIGINAL CLASS-E INVERTER

Fig. 1 shows a circuit topology of the class-E inverter. The class-E inverter is composed of input voltage source $V_I$, RF-choke inductance $L_I$, switching device $S$, shunt capacitance $C_S$, and output resonant filter $L_0 - C_0 - R_L$.

In [19], [20], it is proposed that the frequency is adjusted for keeping the amplitude ratio of the two kinds of voltage or current constant. The phase shift between the switching timing and the resonant current can be kept constant against the resonant frequency variations by keeping the ratio. Therefore, the ARFT LLC converter in [19], [20] can achieve the constant phase shift and the low switching losses against the resonant-component tolerance. Besides, no time information is used in the control, the propagation delays in the feedback network are unnecessary to consider.

This paper proposes the FM-controlled load-independent class-E inverter. The FM control regulates the output voltage constant with the prerequisite of no input variation. With the similar operation mechanism to the ARFT LLC converter in [19], [20], the proposed inverter has the self-tuning function without using any time information. Besides, the proposed inverter works on the load-independent operation, even with output resonant-filter component tolerances because the phase shift can be kept constant. Namely, the proposed inverter has robustness against simultaneous variations of load and the output resonant filter components, which achieves constant output voltage with high power-conversion efficiency. This paper gives the analytical expressions of the waveforms and the power-conversion efficiency of the proposed inverter. The validity and effectiveness of the proposed inverter were confirmed from the experimental and theoretical results. It can be stated that we succeeded in loading both the load-independent operation and the self-tuning function onto the class-E inverter simultaneously, similar to the ARFT LLC converter.

Fig. 2(a) shows example nominal waveforms of the class-E inverter for 50% duty ratio, where $\theta = \omega t = 2\pi ft$ represents the angular time, and $f$ is the operating frequency. The switching device is driven by the driving signal $v_{gs}$. When the switch is in the OFF state, the pulse-type voltage appears in the switch voltage $v_S$ as shown in Fig. 2(a). In the nominal operation, the class-E inverter satisfies the class-E ZVS/ZDS conditions, namely

$$v_S(\pi) = 0, \quad \text{and } \left. \frac{dv_S(\theta)}{d\theta} \right|_{\theta=\pi} = 0,$$

at turn-ON instant, as shown in Fig. 2(a). By satisfying the class-E ZVS/ZDS conditions in (1), it is possible to achieve high power-conversion efficiency at high frequencies.

The output voltage becomes sinusoidal through the $L_0 - C_0 - R_L$ series resonant filter at high loaded quality factor, which is defined as

$$Q = \frac{\omega L_0}{R_L}.$$

Fig. 2(b) and (c) show the example waveforms of the class-E inverter for fixed $R_L$ and fixed $L_0$, respectively, where the subscript ‘r’ means the rated component value. The component values for satisfying the class-E ZVS/ZDS conditions and the rated output voltage automatically. However, it is not very easy to implement a feedback network at the megahertz operation, in particular. This is because it is necessary to measure and adjust the propagation delays of feedback-network components [17].

**FIGURE 1.** Circuit topology of the class-E inverter.
conditions are determined uniquely under the design specifications [23]–[29], which are operating frequency, duty ratio, $Q$ value of the output resonant filter, and load resistance. Besides, the class-E inverter performance is sensitive to the resonant-filter components because of the high-$Q$ resonant filter. Therefore, the inverter works outside the nominal operations [30]–[33] against load variations or component tolerances in the resonant filter, as shown in Fig. 2(b) and (c). In these cases, the class-E inverter cannot achieve the rated output voltage or the class-E ZVS/ZDS. Therefore, the power-conversion efficiency decreases due to the switching losses. It is necessary to take measures against the load variations and the component tolerances in the resonant filter in the class-E inverter.

### B. LOAD-INDEPENDENT CLASS-E INVERTER

The load-independent class-E inverter [11]–[14] keeps the constant output voltage and the ZVS against load variations without any control. The load-independent operation is a particular operating mode of the class-E inverter, which can be obtained by setting the proper component values. The circuit topology of the load-independent class-E inverter is completely the same as the original class-E inverter. Still, the RF choke in the original class-E inverter is replaced with a finite dc-feed inductance. By considering the finite dc-feed inductance, a new resonant structure of $L_{I} - C_{S}$ appears, which is a key point for realizing the load-independent operation [11]–[14]. For obtaining the load-independent operation, the resonant frequency of $L_{I} - C_{S}$ should satisfy

$$\omega_{r}^{2} = \frac{\omega_{I}}{\omega_{r}} = \frac{1}{\omega_{r} \sqrt{L_{I} C_{S}}} \approx 1.2915. \quad (3)$$

Fig. 3(a) shows example nominal waveforms of the load-independent class-E inverter. The load-independent class-E inverter achieves both the constant output voltage and the ZVS against the load variations, as shown in Fig. 3(b). Additionally, it is a sufficient condition for the load-independent operation that the phase shift between the turn-ON instant of the driving signal and the output voltage satisfies

$$\phi = \pi, \quad (4)$$

at any load resistance. When the conditions in (1), (3), and (4) are satisfied, the amplitude of the output voltage becomes

$$V_{m} \approx 1.5893V_{I}. \quad (5)$$

However, the load-independent class-E inverter still has a weakness against the component tolerances in the output resonant filter. Fig. 4 shows superimposed waveforms of the load-independent class-E inverter for fixed output inductance and output capacitance. We can see from Fig. 4(a) that the output-voltage amplitudes are mainly changed from the rated condition when $L_{0}$ or $C_{0}$ varies. Additionally, the switch-voltage waveforms are also changed from the rated condition. A high switch jump appears when $L_{0}$ and $C_{0}$ are lower than the rated values, which causes the efficiency deterioration significantly. Moreover, the inverter cannot keep the rated output voltage and phase shift with $\phi = \pi$ as shown in Fig. 4, which means the condition for achieving the load-independent operation is collapsed. Therefore, the load-independent class-E inverter loses the robustness to the load variations if there are the output resonant-filter component tolerances.

Fig. 5 shows superimposed waveforms of the load-independent class-E inverter for fixed input inductance and shunt capacitance. We can see from Fig. 5 that the output-voltage amplitudes are almost constant against $L_{I}$ and $C_{S}$ variations. Additionally, the phase shift $\phi = \pi$ is kept constant. We can also see that the changes of the switch-voltage waveform are limited, and the switching conditions are almost the same in the ±10% component-value variations.
The load-independent class-E inverter has low sensitivity to the variations of $L_I$ and $C_S$. This is because $\omega_1^* \approx 1.2915$ in (3) means that the resonant frequency is far from the operating frequency, and $Q$-value of the $L_I$ and $C_S$ is low compared with that of the output resonant filter. Namely, the load-independent inverter with $L_I$ and $C_S$ tolerances can maintain the load-independent operation.

From the above discussions, we can confirm that the load-independent class-E inverter has a high sensitivity to $L_0$ and $C_0$ but low sensitivity to the $L_I$ and $C_S$. Additionally, it is confirmed that the waveforms in Fig. 4(a) are completely the same as those in Fig. 4(b). This is because the product of $L_0$ and $C_0$ governs the inverter dynamics, as described in the waveform equations at Section IV. Therefore, this paper focuses on the variations of load and the output resonant inductance.

III. PROPOSED INVERTER

This paper proposes the FM-controlled load-independent class-E inverter. Fig. 6(a) shows a circuit topology of the proposed inverter. In the proposed inverter, the FM control network for the output voltage regulation [34] is added to the load-independent class-E inverter, described in Section II-B.

The FM control network is composed of peak detection circuit, voltage follower, and digital signal processor (DSP). The output of the DSP becomes an input signal of the gate driver. The load-independent class-E inverter has low sensitivity to $L_0$ and $C_0$ but low sensitivity to the $L_I$ and $C_S$. Namely, the proposed inverter can satisfy the load-independent operation conditions by varying the operating frequency despite the component tolerance in the output-resonant filter when the input voltage is constant. Therefore, the inverter achieves the ZVS at any load resistances due to the load-independent operation. As a result, the proposed inverter provides the constant output voltage with high power-conversion efficiency against the load variations and the component tolerance in the output-resonant filter.

FIGURE 6. Proposed inverter. (a) Circuit topology. (b) Example waveforms in the control network.

Because the DSP has a limitation for its input-voltage range, the values of $R_1$ and $R_2$ should be determined so that $V_{C_1}$ is in the range of the limitation. Besides, the smoothing capacitance should be sufficiently large to reduce the voltage ripple. The voltage follower is inserted to prevent the current from flowing into the DSP.

The operating frequency is adjusted for minimizing

$$g(f) = \left[ \frac{V_{C_1}(f)}{R_2 + R_2 V_m R_1} \right]^2. \quad (7)$$

The hill-climbing method is adopted for determining the frequency in this paper. The output of the DSP is the square-waveform voltage with the frequency $f$, which is an input of the gate driver. The load-independent class-E inverter is driven by the gate driver.

Because of the characteristics of the load-independent inverter, the phase shift satisfies $\phi = \pi$ in case of $V_m/V_I = V_{mr}/V_1$. Namely, the proposed inverter can satisfy the load-independent operation conditions by varying the operating frequency despite the component tolerance in the output-resonant filter when the input voltage is constant. Therefore, the inverter achieves the ZVS at any load resistances due to the load-independent operation. As a result, the proposed inverter provides the constant output voltage with high power-conversion efficiency against the load variations and the component tolerance in the output-resonant filter.

From the above discussion, we can say that the self-tuning function can be obtained by adding the feedback network for keeping the constant phase shift $\phi$ to the load-independent inverter. In the proposed method, it is unnecessary to consider the propagation delay of the FM-control network for realizing the fixed phase shift, which is an advantage we would like to claim.

This phase-shift keeping mechanism of the proposed inverter is similar to that in the ARFT LLC converter in [18]–[20]. The proposed inverter keeps the ratio of two kinds of voltages in the system, namely the input voltage to output one, constant. Therefore, it is an essential prerequisite that smoothing capacitance is

$$V_{C_1} = \frac{R_2}{R_1 + R_2} V_m. \quad (6)$$
there is no input variation in the proposed inverter, which is the same prerequisite as the load-independent inverter.

IV. ANALYTICAL WAVEFORM EQUATIONS OF THE CLASS-E INVERTER WITH FINITE DC-FEED INDUCTANCE

The feature of the load-independent inverter is to have finite dc-feed inductance. Besides, the load-independent class-E inverter achieves only the ZVS, namely outside of the nominal conditions. Therefore, this section gives the steady-state analytical waveform equations of the class-E inverter with finite dc-feed inductance outside of the nominal operation.

A. SWITCHING PATTERNS

Fig. 7 shows the typical waveforms of the switch voltage and current. We classify the switching patterns into two patterns. Fig. 7(a) shows the switch-voltage waveform, which does not reach zero at the moment of the switch turn-ON. This switching pattern is called non-ZVS. Here, we define the parameter \( \theta_1 \), indicating the timing when the switch voltage becomes zero. In non-ZVS operation, we have \( \theta_1 = \pi \).

Fig. 7(b) shows the switch-voltage waveform which reaches zero prior to turn-on switching. In this case, the MOSFET anti-parallel diode turns ON at \( \theta = \theta_1 \), and the switching device turns ON during the diode is in the ON state, which can be regarded as ZVS.

The switching pattern highly depends on the output-current waveform because the switch voltage is expressed as the integration of the difference between input current and output current. Namely, the switching patterns have sensitivity to the load variations and the component tolerance in the output resonant filter.

B. ASSUMPTION FOR ANALYSIS

In the analysis, the voltages and currents are normalized by the input voltage and load resistance, for example,

\[
v_o^* = \frac{v_o}{V_i}, \quad \text{and} \quad i_o^* = \frac{R_{Lr} i_o}{V_i},
\]

where ‘*’ means a label of the ‘normalized’. Additionally, we use dimensionless parameters and variables to give a general discussion. The passive components are set to dimensionless, for example,

\[
\lambda_0 = \frac{\omega_r L_0}{R_{Lr}}, \quad \gamma_0 = \omega_r C_0 R_{Lr}, \quad \text{and} \quad \rho_L = \frac{R_L}{R_{Lr}}.
\]

For simplifying the analysis, the following assumptions are given.

(a) \( Q \) value of output-resonant filter is sufficiently high so that the output current is a pure sinusoid with the operating frequency \( f \), namely

\[
i_o^*(\theta) = I_m^* \sin(\theta + \phi),
\]

where \( I_m^* \) is the normalized output-current amplitude.

(b) The MOSFET and its anti-parallel diode are modeled as an ideal switch.

(c) The MOSFET turns OFF and ON at \( \theta = 0 \) and \( \theta = \pi \), respectively.

(d) The parasitic capacitance of the MOSFET is included to the shunt capacitance.

(e) All the passive components, including the MOSFET parasitic capacitance, works as linear components. Besides, the equivalent series resistances (ESRs) of the inductances and the capacitances are ignored for the waveform-equation derivations.

Following the above assumptions, the equivalent model of the inverter is illustrated, as shown in Fig. 8.
C. WAVEFORM EQUATIONS
Following the procedure of the analysis in [29], input current, switch current, and switch voltage are expressed as (11)–(13), shown at the bottom of the previous page, where

\[
\begin{align*}
A &= \frac{1}{1 - \cos \left(\frac{\omega_m}{\omega_1} \phi \right)} \left\{ \frac{\omega_m^2 (2 \pi - \theta_1) + \omega_m \sin \left(\frac{\omega_m}{\omega_1} \theta_1 \right)}{\lambda I} \right. \\
- q I_s^* \left[ \cos \phi \sin \left(\frac{\omega_m}{\omega_1} \theta_1 \right) - \omega_m s \left[ \sin (\theta_1 + \phi) - \sin \phi \right] \right] \right. \\
B &= \frac{1}{\lambda I} - q I_s^* \cos \phi, \\
\omega^* &= \frac{\omega_m}{\omega_r}, \\
\text{and} \\
q &= \left(\frac{\omega_m}{\omega_r} - \frac{\omega_m}{\omega_1} \right)^{-1}.
\end{align*}
\]

From the assumption (a), the output-resonant filter extracts only the fundamental frequency component. Now, we divide the output-resonant inductance into the resonant inductance \(L_a\) and the reactive inductance \(L_b\), as shown in Fig. 8. \(L_b\) has a role of phase shift adjustment with satisfying

\[
\omega_a = \frac{\omega_a}{\omega_r} = \frac{1}{\sqrt{\lambda a \gamma_0}} = 1. \quad (18)
\]

By applying KVL, the fundamental frequency component of the switch voltage is expressed as

\[
\begin{align*}
V_{S_{\text{fund}}}^*(\theta) &= V_m^*(\theta) + V_{L_b}^*(\theta) \\
&= V_m^* \sin (\theta + \phi) + V_{L_b}^* \cos (\theta + \phi). \quad (19)
\end{align*}
\]

where \(V_m^*\) and \(V_{L_b}^*\) are the normalized amplitude of the output voltage and the voltage across \(L_b\), respectively. By applying Fourier transform to (19), we have the detailed expressions of the amplitudes as given in (20) and (21), as shown at the bottom of the page.

D. QUANTITATIVE WAVEFORM DERIVATIONS
It is necessary to determine the unknown variations, namely \(I_m^*, \phi, \text{ and } \theta_1\), for obtaining the quantitative waveforms. Fig. 9 shows a flow chart for the unknown-variation determinations.

As an initial step, we assume that there is no MOSFET anti-parallel diode, namely \(\theta_1 = \pi\). In this case, \(I_m^*\) and \(\phi\) are calculated by solving the simultaneous algebraic equations of (20) and (21). When the derived switch voltage satisfies \(v_S(\pi) \geq 0\), the switch voltage is in the non-ZVS state, and the condition of \(\theta_1 = \pi\) is valid. Therefore, it is regarded to obtain the quantitative waveforms.

In the case of \(v_S(\pi) < 0\), the inverter achieves the ZVS and the MOSFET anti-parallel diode works. Therefore, \(\theta_1 (< \pi)\) should be derived as unknown parameters.
The condition of $\theta_1$ is expressed as
\[ v^2_S(\theta_1) = 0. \tag{22} \]

By resolving the simultaneous algebraic equations (20)-(22), we obtain $\phi$, $I^*_{m}$, and $\theta_1$ and the quantitative waveforms in the ZVS state.

### E. PROPOSED INVERTER WAVEFORMS

In the proposed inverter, the operating frequency $f$ depends on the component values. This is because the FM control is adopted for regulating the output voltage. Namely, we need to add one more condition for obtaining the quantitative waveforms of the proposed inverter, which is
\[ \rho L I^*_{m} - V^*_{mr} = 0. \tag{23} \]

The waveform derivation process is the same as that in Fig. 9. By solving (20), (21), and (23) with or without (22), we can determine the unknown variations and obtain the quantitative waveforms.

### V. OUTPUT POWER AND POWER-CONVERSION EFFICIENCY

By using the waveform equations derived in Section IV, the power-conversion efficiency can be expressed theoretically. Fig. 10 shows a power-loss model. In this paper, the power-losses in MOSFET with its anti-parallel diode and conduction losses in the inductance ESRs are considered. It is assumed that the power-loss factors, namely switch on-resistances, the forward voltage of the diode, and ESRs, do not affect the waveforms derived in Section IV.

The normalized output power is expressed as
\[ P_o^* = \frac{\rho L}{2\pi} \int_0^{2\pi} i^*_o(\theta)^2 d\theta = \frac{\rho L I^2_{m}}{2}. \tag{24} \]

In the MOSFET, the conduction loss at ON-resistance occurs as
\[ P^*_S = \frac{\rho S}{2\pi} \int_0^{2\pi} i^*_S(\theta)^2 d\theta. \tag{25} \]

In the case of non-ZVS, the switching loss should be considered as
\[ P^*_SW = \frac{\gamma S}{4\pi} v^2_S(\pi)^2. \tag{26} \]

The power loss in the MOSFET anti-parallel diode is expressed as
\[ P^*_{SD} = \frac{1}{2\pi} \int_0^{2\pi} V^*_th i^*_S(\theta) d\theta + \frac{\rho D}{2\pi} \int_0^{2\pi} i^*_S(\theta)^2 d\theta, \tag{27} \]

where $V^*_th$ is the normalized forward voltage of the anti-parallel diode. The power-losses in the ESRs of the dc-feed and output-resonant inductances are given as
\[ P^*_{L_i} = \frac{\rho L_i}{2\pi} \int_0^{2\pi} i^*_o(\theta)^2 d\theta, \tag{28} \]

and
\[ P^*_{L_o} = \frac{\rho L_o}{2\pi} \int_0^{2\pi} i^*_o(\theta)^2 d\theta = \frac{\rho L_o I^2_{m}}{2}. \tag{29} \]

According to the power loss factors in (24)-(29), the power-conversion efficiency of the inverter can be obtained analytically as
\[ \eta = \frac{P_o^*}{P_o^* + P^*_S + P^*_SW + P^*_SD + P^*_{L_i} + P^*_{L_o}}. \tag{30} \]

The resulting equations of the power-loss factors are shown in Appendix.

### VI. ANALYTICAL INVESTIGATIONS OF THE PROPOSED INVERTER PERFORMANCE

This section investigates the proposed inverter performance theoretically. The proposed inverter is compared with the FM-controlled original class-E inverter and the load-independent class-E inverter against load and output resonant-filter component variations. As described in Section II-B, we investigate the tolerance of $L_0$ as the output resonant-filter component tolerances.
Fig. 11 shows the normalized amplitude of the output voltage on the load resistance and the output resonant-inductance space. It is seen from Fig. 11(a) and (c) that the FM-controlled class-E inverter and the proposed one can keep the rated output voltage in the entire parameter space. This is because the FM control works to keep the output voltage constant. In the load-independent class-E inverter, the output voltage is constant against load variations only at $L_0 = L_{0r}$ as shown in Fig. 11(b). However, it decreases exceedingly against the variation in the output resonant-inductance at the heavy load, in particular. This result shows that the sensitivity to the resonant frequency is high at a high $Q$ resonant filter. The output voltage variation at large load resistance is slight because the output $Q$ becomes low with the increase in the load resistance as in (2).

Fig. 12 shows the phase shift $\phi$ on the load resistance and the output resonant-inductance space. It is seen from Fig. 12(a) that the phase shift varies against the load variations. However, it is constant against the variation in $L_0$ for the fixed $R_L$. This result shows that the proposed FM-controlled class-E inverter has a self-tuning function like the inverters in [15]–[17]. We can see from Fig. 12(b) that the load-independent inverter keeps the phase shift constant with $\phi = \pi$ against load variations at $L_0 = L_{0r}$, which is a feature of the load-independent operation. However, the phase shift varies for $L_0 \neq L_{0r}$ because the inverter works outside of the
TABLE 1. Component values of the proposed inverter at the rated condition.

| Component | Theoretical | Measured |
|-----------|-------------|----------|
| $f_r$     | 1 MHz       | 1 MHz    |
| $V_I$     | 10.0 V      | 10.0 V   |
| $L_{0r}$  | 31.83 µH    | 31.18 µH |
| $C_S$     | 4.36 nF     | 4.32 nF  |
| $C_D$     | 0.819 nF    | 0.830 nF |
| $R_{Lr}$  | 20.0 Ω      | 20.1 Ω   |
| $C_1$     | 0.1 µF      | 0.1 µF   |

![FIGURE 14. Photo of the implemented inverter.](image)

load-independent operation. It is confirmed from Fig. 12(c) that the proposed inverter keeps the phase shift almost constant in the entire parameter space. Namely, the constant phase shift against both the load and output-resonant component variations can be realized by regulating the ratio of the input voltage and the output voltage with the operating frequency variations.

Fig. 13 shows the voltage across the switch at turn-ON instant on the load resistance and the output resonant-inductance space. The non-ZVS generates switching loss as given in (26). It is seen from Fig. 13(a) that the FM-controlled original class-E inverter can satisfy the class-E ZVS/ZDS conditions against the variation in $L_0$ at $R_L = R_{Lr}$. This operation is the same as the self-tuning operation. Namely, the phase shift does not vary at the fixed $R_L$, as shown in Fig. 12(a). It is also seen from Fig. 13(a) that the switching loss occurs and increases with the increase in the load resistance. The FM-controlled original class-E inverter has a weakness against load variations. It can be seen from Fig. 13(b) that the load-independent class-E inverter achieves the ZVS at $L_0 \geq L_{0r}$. It is well known that the ZVS can be achieved when the output resonant-filter is inductive from the ZVS condition with the help of the MOSFET anti-parallel diode. Conversely, non-ZVS appears for $L_0 < L_{0r}$. It is confirmed from Fig. 13(c) that the proposed inverter can suppress the switching voltage in the entire region of the $R_L - L_0$ space.

We can state from Fig. 11-13 that the proposed inverter has both load-independent and self-tuning functions. The proposed inverter achieves the output regulation and high power-conversion efficiency against load variations and output resonant-filter component tolerances. In other words, we succeeded in incorporating similar characteristics to the ARFT LLC converter in [18]–[20] to the class-E inverter.

![FIGURE 15. Superimposed analytical and experimental waveforms for fixed $L_0$. (a) and (b) Analytical waveforms. (c) and (d) Experimental waveforms. (a) and (c) On the load-independent class-E inverter. (b) and (d) On the proposed inverter.](image)

VII. EXPERIMENTAL VERIFICATIONS

This section shows the experimental verifications of the proposed inverter. In this paper, the design specifications were given as follows: rated operating frequency $f_r = 1$ MHz, input voltage $V_I = 10$ V, rated load resistance $R_{Lr} = 20$ Ω, loaded quality factor $Q = 10$, and duty ratio $D = 0.5$. From the design specifications, we designed the load-independent class-E inverter following the analytical equations in [13]. Table 1 gives the theoretical and measured component values of the designed inverter at the rated condition, where the component values were measured by Keysight E4990A impedance analyzer.

In the implemented inverter, the Vishay IRF530 MOSFET was used for the switching device. From the datasheet, we obtained that ON-resistance of the MOSFET and ON-resistance and forward voltage of the MOSFET anti-parallel diode are $R_S = 0.16$ Ω, $R_D = 0.15$ Ω, and $V_{th} = 0.7$ V, respectively. Besides, we have the MOSFET drain-to-source parasitic capacitance $C_{ds} = 0.19$ nF from the datasheet, which is much less than the shunt capacitance $C_S = 4.36$ nF. Therefore, the assumption (e) was valid in this experiment. The Renesas EL7104 power-MOSFET driver was adopted for driving the MOSFET.

The design value of the shunt capacitance decreases as the operating frequency increases. Therefore, the parasitic-capacitance nonlinearity influence on the inverter performance increases at higher frequencies. This problem was pointed out and considered in [14]. The design technique in [14] can be reflected in the proposed inverter design.
In the FM control network, STPSH100 diode from STMicroelectronics, LM7171 operational amplifier, and TMS320F28379D Dual-Core Delfino Microcontroller from Texas Instruments were used. The TMS320F28379D Microcontroller can detect up to 3 V. Therefore, we determined \( R_1 = 90 \, k\Omega \) and \( R_2 = 10 \, k\Omega \). Additionally, \( R_3 = 10 \, M\Omega \) and \( C_1 = 0.1 \, \mu F \) are determined for obtaining low-ripple voltage for the input to the DSP. Fig. 14 shows a photo of the implemented inverter.

We compared the proposed inverter performance with the identical load-independent class-E inverter without FM control in the experiments.

A. WAVEFORM INVESTIGATIONS FOR COMPONENT TOLERANCES

Fig. 15 shows the analytical and experimental waveforms for ±10% \( L_0 \) tolerances. We can see from Fig. 15(a) and (c) that the load-independent class-E inverter had a significant jump in the switch voltage at \( L_0/L_{QR} = 0.9 \). Besides, the amplitude and phase of the output voltage varied from the rated condition at \( L_0/L_{QR} \neq 1 \).

In contrast, the proposed inverter satisfied the ZVS condition despite the \( L_0 \) tolerance, as shown in Fig. 15(b) and (d). It is seen from this result that the FM control effectively reduced the switching loss against the \( L_0 \) variations. Additionally, it is confirmed that the amplitude and the phase shift of the output voltage were constant regardless of \( L_0 \). This is a piece of evidence that the proposed inverter could keep load-independent operation even if the \( L_0 \) tolerances occur.

The above results confirm that the proposed inverter has robustness against the output resonant-filter component tolerances. Besides, we can confirm that the experimental waveforms agreed well with the analytical waveforms, showing the validity of the analytical expressions.

B. WAVEFORM INVESTIGATIONS FOR LOAD VARIATIONS

Fig. 16 shows experimental waveforms against load variations for fixed \( L_0 \) in the load-independent class-E inverter and the proposed inverter. We can see from Fig. 16(b) that the switch voltage achieved the ZVS against load variations at \( L_0/L_{QR} = 1 \) in the load-independent class-E inverter. However, the load-independent class-E inverter could not achieve the load-independent operations when \( L_0 \) varied, as shown in Fig. 16(a) and (c). It is seen from these figures that the output voltage and the phase shift were changed against load variations. We see from Fig. 16(a) that the switching voltage depended on the load resistance, and the high switch-voltage jump occurred when both the resonant inductance and the load resistance were lower than the rated values, as shown in Fig. 16(a).

In contrast, we can see from Fig. 16(d)-(f) that the proposed inverter achieved the load-independent operations at any \( L_0 \). All the output voltage amplitudes in Fig. 16(d)-(f) are identical regardless of \( R_1 \) and \( L_0 \). Compared with Fig. 16(a) and (d), we can confirm that the jump of the switch voltage at turn-on instant was effectively suppressed in the proposed inverter. By comparisons between Fig. 16(a)-(c) and Fig. 16(d)-(f), it can be stated that the proposed inverter acquires the ability to achieve the load-independent operation against the output resonant-filter component tolerances by adding the FM-control network.

The harmonic distortions were obviously included in the output-voltage waveforms for \( R_1/L_{QR} = 5 \) in Fig. 16. This is caused by the low \( Q \) output-resonant filter, and the assumption (a) is not valid in the strict sense. However, the measured waveform characteristics fully agreed with the theoretical investigations in Figs. 11-13. This means that the assumption (a) is acceptable for the proposed inverter investigations. Additionally, the agreements complement each other in the accuracy of the analysis and the validity of the experiment.

C. PROPOSED INVERTER EVALUATIONS

1) OUTPUT VOLTAGE AND POWER-CONVERSION EFFICIENCY

Fig. 17 shows the power-conversion efficiency and the amplitude of the output voltage as functions of the load resistance. In Fig. 17, the measured power-conversion efficiency was obtained from

\[
\eta = \frac{P_o}{P_i} = \frac{V_o^2}{2R_LV_II_i}, \quad (31)
\]

where \( V_o \) is the RMS of the output voltage, which was measured by the Keysight 3458A Digital Multimeter. The input voltage and current were measured by the Iwatsu VOAC7523H Digital Multimeter.

Fig. 17 shows the inverter characteristics for fixed \( L_0 \). We can see from Fig. 17(b) that both the load-independent class-E inverter and the proposed inverter achieved constant output...
FIGURE 17. The power-conversion efficiency $\eta$ and output-voltage amplitude $V_m$ as a function of the normalized load resistance. (a) For $L_0/L_{QR} = 0.9$. (b) For $L_0/L_{QR} = 1$. (c) For $L_0/L_{QR} = 1.1$.

FIGURE 18. Operating frequency as a function of the output resonant inductance.

Voltage and sufficiently high power conversion efficiency at any load resistances. This is because both the inverter worked on the load-independent operations. The ZVS achievement leads to high power-conversion efficiency.

We can see from Fig. 17(a) and (c) that the load-independent class-E inverter decreased the output voltage significantly at low load resistance. This is because the inverter worked outside the load-independent operation when there were the $L_0$ tolerances. The power-conversion efficiency also decreased at low output resistance for $L_0/L_{QR} = 0.9$, as shown in Fig. 17(a). If we consider the power-conversion efficiency performance and the waveforms in Fig. 16, we can see that the reason for the efficiency deterioration was non-ZVS. In contrast, the proposed inverter kept the output voltage constant and high power-conversion efficiency against load variations even with output resonant-inductance tolerances. This is because the proposed inverter works on the load-independent operation regardless of the output resonant-inductance tolerances.

It is confirmed from Fig. 17 that all the measured characteristics agreed with the theoretical predictions quantitatively, which showed the validity of the analytical expressions of the power-loss analysis given in Section V.

We can confirm from Fig. 17 that the high power-conversion efficiency and the rated output voltage were kept against load variations and component tolerances by the FM control. In particular, the proposed inverter overcame the issue related to the component variation in the output-resonant filter of the load-independent class-E inverter. This performance improvement could be realized by the exquisite combination of the load-independent inverter and the FM control. From Fig. 17, it can be confirmed that the measured characteristics are in good agreement with the theoretical prediction even when the load resistance is large. This result also confirms that the error in harmonic distortion of the output voltage due to assumption (a) has a negligible effect on the characteristics. The agreements in the power-conversion efficiency showed the validity of the analytical expressions of the power-loss analysis given in Section V.

At the rated condition, the implemented inverter achieved 95% power-conversion efficiency with the 1-MHz and 5.7 W output. Besides, the implemented inverter achieved the constant output voltage with high power-conversion efficiency against the load variations and the component tolerance in the resonant inductance.

2) OPERATING FREQUENCY

Fig. 18 shows the operating angular frequency of the proposed inverter as a function of the normalized output-resonant inductance. In the experiments, the frequency was measured by the Keysight 3458A Digital Multimeter. We can see from Fig. 18 that the operating frequency decreased as the output-resonant inductance increased. Roughly speaking, the operating frequency tracks the resonant frequency for keeping $\omega_0$ constant. The variation range of the frequency was from $-6.0\%$ to $+6.8\%$ according to $\pm10\%$ variations of the output resonant inductance. We can confirm from Fig. 18 that the
experimental results were in good agreement with the analytical predictions, assuring the credibility of the experiment and analysis mutually.

D. Dynamic Response for the Step Changes in the Load Resistance

Fig. 19 shows experimental waveforms with sudden load variations. It is seen from Fig. 19(b) that the load-independent class-E inverter with \( L_0 = L_{0r} \) kept the output voltage constant against sudden load variations without any control. However, we can see from Fig. 19(a) and (c) that the load-independent inverter with \( L_0 \) tolerance could not keep the rated output voltage. The output voltage depended on the load resistance. In Fig. 19(a), the negative switch voltage appeared due to the ringing associated with non-ZVS.

It can be seen from Fig. 19(d)-(f) that all the output-voltage amplitudes in the steady-state were identical regardless of the resonant inductance and load resistance. Additionally, there was no difference in the transient response among Fig. 19(d)-(f). Also, it can be confirmed that the transient responses in Fig. 19(d)-(f) were almost the same as that in Fig. 19(b). These results showed that the proposed inverter could keep the output voltage constant against sudden load variations by the transient behaviors based on the load-independent operations. Although the time constant of the FM-control network generated by \( C_3 \) and \( R_3 \) is much lower than the operating frequency, it does not degrade the frequency tracking performance against sudden load variations. Namely, the FM control mainly works on the variations in the filter components. Still, it does not work on the output-voltage regulation in the transient response against the sudden load variations. As stated in Section I, the high-speed response is unnecessary against the output resonant-filter component variations. Therefore, it can be concluded that the proposed inverter has sufficient ability against both the load variations and the output resonant-filter component tolerances.

VIII. Conclusion

This paper has proposed the FM controlled load-independent class-E inverter. The main idea is to apply the FM control to the load-independent class-E inverter. Because the proposed control can keep the phase shift constant even though the output resonant-filter components vary. As a result, the proposed inverter can keep high power-conversion efficiency and the rated output voltage against both load variations and output resonant-filter component tolerances. Because the class-E inverter has an advantage for high-frequency operation, it is expected that the proposed inverter will be applied to wide-area applications. The validity and effectiveness of the proposed inverter were confirmed from the good agreements
between the experimental and theoretical results. The implemented inverter achieved 95% with 5.7 W output and 1-MHz operating frequency at the rated operation.

In the proposed inverter, we used the ratio of the input to output voltages for achieving the constant phase shift. The other combinations of voltages and currents are also the candidates for obtaining the similar performance, as investigated in [20]. To find out the best combination needs to be addressed in the future.

**APPENDIX**

**RESULTING EQUATIONS OF POWER LOSS**

The power-loss factors we expressed analytically are

\[
P^*_{SW} = \frac{V_{th}^2}{2\pi} \left[ \frac{1}{\rho \cdot \omega^2} \left( A \sin \left( \frac{\omega^2}{\omega^2 - \theta^2} \right) \right) + B \left( \frac{\cos \left( \frac{\omega^2}{\omega^2 - \theta^2} \right) + 2qI_m^2 \cos \phi}{\omega^2 - \theta^2} \right) \right] + \frac{1}{2\pi} \left[ \frac{1}{\rho \cdot \omega^2} \left( A \sin \left( \frac{\omega^2}{\omega^2 - \theta^2} \right) \right) + B \left( \frac{\cos \left( \frac{\omega^2}{\omega^2 - \theta^2} \right) + 2qI_m^2 \cos \phi}{\omega^2 - \theta^2} \right) \right]
\]

\[
P^*_{SD} = \frac{1}{2\pi} \left\{ \frac{1}{\rho \cdot \omega^2} \left( A \sin \left( \frac{\omega^2}{\omega^2 - \theta^2} \right) \right) + B \left( \frac{\cos \left( \frac{\omega^2}{\omega^2 - \theta^2} \right) + 2qI_m^2 \cos \phi}{\omega^2 - \theta^2} \right) \right\}
\]

and

\[
P^*_{Li} = \frac{\omega^2 A}{\omega^2} \sin \left( \frac{\omega^2}{\omega^2 - \theta^2} \right) + \frac{\omega^2 B}{\omega^2} \left[ 1 - \cos \left( \frac{\omega^2}{\omega^2 - \theta^2} \right) \right] + \frac{2\omega qI_m^2}{\omega^2} \cos \phi + \frac{3\pi^2 \omega^2}{2\lambda^2} + \frac{\kappa \cdot \pi}{\omega^2 - \theta^2} \cos \phi + \frac{3\pi^2 \omega^2}{2\lambda^2} + \frac{\kappa \cdot \pi}{\omega^2 - \theta^2} \cos \phi - 1.
\]

where \( \kappa \) is defined as

\[
\kappa = A \cos \left( \frac{\omega^2}{\omega^2 - \theta^2} \right) + B \cos \left( \frac{\omega^2}{\omega^2 - \theta^2} \right) + \frac{\omega^2 qI_m^2}{\omega^2} \sin \left( \theta + \phi \right) - 1.
\]

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