Design of Class-$E_M$ Power Amplifier Taking into Account Auxiliary Circuit

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Abstract—This paper presents a novel design procedure for Class-$E_M$ amplifier. To improve the whole power conversion efficiency, the ZVS class-$E$ doubler is applied to the auxiliary circuit. The auxiliary circuit satisfies zero voltage switching, which makes the power conversion efficiency be improved. Moreover, we applied the numerical design procedure for the class-$E$ amplifier to the design of the class-$E_M$ amplifier. As a result, it is possible to design both the main circuit and the auxiliary one as one circuit. In addition, the derived element values have high accuracy for achieving all the switching conditions simultaneously. By comparison with calculated results and experimental ones, we can show the validity and effectiveness of the proposed design procedure. The laboratory experiment shows the amplifier with proposed design procedure achieves the 94.0 % power conversion efficiency under the 13.4 W output power and 3.5 MHz operation.

I. INTRODUCTION

The class-$E_M$ amplifier[1] is an improved version of class-$E$ amplifier[2]–[4], which achieves smooth switching at not only the turn-on transition but also, the turn-off transition. By achieving zero and zero slope of current switching at the turn-off transition, the auxiliary circuit whose output has a harmonic frequency is added to the class-$E$ amplifier. Because of smooth switching, the class-$E_M$ amplifier achieves higher power conversion efficiency than the class-$E$ amplifier when the transistor has long turn-off-switching time, in particular. Allowing slow switching means that the power of driving circuit can be decreased. Therefore, the class-$E_M$ amplifier improves the power conversion efficiency with low cost.

The design example and the experimental results of the class-$E_M$ amplifier were presented in [1]. However, the design of the main circuit is dominant in the explanations of [1]. It seems that the design and the power conversion efficiency of the auxiliary circuit were not considered in detail. The auxiliary circuit was considered only to inject a biharmonic current into the main circuit with the proper phase and amplitude. We recognize the design of both the main circuit and the auxiliary one is important to achieve high power conversion efficiency. In particular, it is necessary to consider the switching conditions of the auxiliary circuits to reduce the switching losses. In [1], there is a discussion about the possibility that the main circuit and the auxiliary one can be designed separately. We consider, however, that it is important to design both circuits as one amplifier.

This paper presents a novel design procedure for Class-$E_M$ amplifier. The features of the proposed design are as follows: 1. the class-$E$ frequency doubler is applied to the auxiliary circuit[5]–[7], 2. the zero voltage switching (ZVS) is achieved in the auxiliary circuit, 3. the design procedure presented in [4] is adopted for the design, 4. the design procedure takes the effects of parasitic resistance and the switch-on resistance into account. It is possible to design both the main circuit and the auxiliary one as one circuit by using the proposed procedure. Moreover, the derived element values have high accuracy for achieving all the switching conditions simultaneously. By comparison with calculated results and experimental ones, we can show the validity and effectiveness of the proposed design procedure. The laboratory experiment shows the amplifier with proposed design procedure achieves the 94.0 % power conversion efficiency under the 13.4 W output power and 3.5 MHz operation.

II. CLASS-$E_M$ AMPLIFIERS

A. Circuit Topology

Fig. 1(a) shows the circuit topology of the class-$E_M$ amplifier[1]. The class-$E_M$ amplifier has a main circuit whose output has a fundamental frequency and an auxiliary circuit whose one has a harmonic frequency. Both of the circuits have similar topologies to the class-$E$ amplifier. Both the main and auxiliary circuits consist of a dc-supply voltage source $V_{DC}$, a dc-feed inductor $L_C$, a MOSFET $S$ as a switching device, a shunt capacitance $C_S$, and a series resonant circuit $L−C$. The output current of the main circuit flows through the load resistance $R$. On the other hand, that of the auxiliary circuit flows through the switch of the main circuit.

B. Principle Operation

The example waveforms of the class-$E_M$ amplifier are shown in Fig. 2 when the switch-off duty ratio of the main circuit $D_1$ is 0.5. The resonant filter on the main and auxiliary circuits are tuned on a fundamental frequency and
a biharmonic one, respectively. Since the resonant filters have a high quality factor, both the currents $i_o$ and $i_{inj}$ are sinusoidal with the tuned frequency.

The switch $S_1$ is driven by a driving pattern of $D_r1$ in Fig. 2. When the switch $S_1$ is in an off-state, the sum of current through dc-feed inductance and resonant filters of main and auxiliary circuits flow through the shunt capacitance $C_{S1}$, which produces the switch voltage $v_{S1}$. In this interval, the current through the switch $i_{S1}$ is almost zero. On the other hand, when the switch $S_1$ is in an on-state, the voltage across the switch is almost zero. In this interval, the current $i_{S1}$ flows through the switch $S_1$. At the instance of turn-on switching, the switch voltage $v_{S1}$ achieves zero and zero derivation switching simultaneously. From this switching, the connection of the switch voltage and current at turn-on transition are very smooth and there is no discontinuity. Similarly, the current $i_{S1}$ achieves zero and zero derivation switching simultaneously at the instance of turn-off switching. Therefore, the connection of the switch current at the turn-off transition is also very smooth. These conditions are expressed as

$$v_{S1}(2\pi D_1) = 0, \quad \left. \frac{d}{d\theta} v_{S1}(\theta) \right|_{\theta = 2\pi D_1} = 0,$$  

$$i_{S1}(2\pi) = 0, \quad \left. \frac{d}{d\theta} i_{S1}(\theta) \right|_{\theta = 2\pi} = 0.$$  

### C. Benefits of The Class-E M Amplifier

The switching conditions of (1) is same as those of class-E amplifier. In the operation of the class-E amplifier, however, appears a step change on the current at the turn-off instance as shown in Fig. 3. From this figure, $t_{off}$ means the drain current fall time. In the interval of drain current falling, the voltage and current appears simultaneously. Therefore, the power losses occur in this interval. If $t_{off}$ is large, the power losses at turn-off instance cannot be ignored. To minimize the $t_{off}$, it is effective to use high-speed MOSFET or to increase the power of driving signal $D_r$. The former, however, takes high-cost to realize the amplifier. The latter suffers from the power conversion efficiency.

The class-E M amplifier is one of the solutions of this problem. By using the auxiliary circuit, the conditions of (2) can be achieved and the step change of the switch current disappears. Therefore, the class-E M amplifier improves the power conversion efficiency with low cost. The class-F power amplifier[8] is based on the same idea. The difference between class-F and the class-E M is the operation of the transistor, namely, using a linear region or switching mode.

### D. Indication of Problem

The design example and the experimental results of the class-E M amplifier were presented in [1]. The experiment-
III. PROPOSED OPERATION OF CLASS-E_M AMPLIFIER

This paper presents a novel design procedure of the class-E_M amplifier. The power conversion efficiency of the auxiliary circuit is considered and the design curves are shown. The features of the proposed design are as follows.

1. The class-E frequency multiplier is applied to the auxiliary circuit. The class-E frequency doubler [5]–[7] generates a higher frequency output than the switching frequency. If the frequency doubler is used as the auxiliary circuit, the injected current into the main circuit is obtained while the switching frequency of the auxiliary circuit is same as that of the main circuit.

2. The zero voltage switching (ZVS) is achieved in the auxiliary circuit. Due to the ZVS, the switching losses are suppressed compared with the auxiliary circuit without any switching conditions.

3. We need to derive the element values so that the class-E_M amplifier achieve live switching conditions simultaneously. Moreover, the class-E_M amplifier has 8 dimensional circuit equations. For achievement of the design under these situations, we use the design procedure presented in [4]. By applying this procedure to the design of the class-E_M amplifier, the design values that satisfies all conditions are obtained with high accuracy. Both the main circuit and the auxiliary one are designed simultaneously as one amplifier.

4. The design procedure takes the effects of parasitic resistance and the switch-on resistance into account. Figure 4 shows an example waveforms of the class-E_M amplifier with proposed design procedure. The operation of the main circuit is same as the previous paper described in Sec. II. The auxiliary circuit is driven by the driving signal \( D_{s2} \) whose frequency is identical to that of the main circuit. To realize frequency multiplier, the switching voltage should include a harmonic frequency component. Therefore, the switch-on duty ratio of the auxiliary circuit is larger than that of main circuit. When the switch \( S_2 \) is in the off-state, the switch voltage appears as shown in Fig. 4. The switch voltage becomes zero at the turn-on transition of the switch, namely the ZVS is achieved. Since the ZVS class-E frequency multiplier is applied to the auxiliary circuit, there are two effects that are decreasing of the switchings and suppression of the switching losses at turn-on instant. Therefore, the power conversion efficiency of the auxiliary circuit becomes higher than that of [1].

IV. DESIGN PROCEDURE

A. Parameters and Assumptions

First, the parameters are defined as follows.

1. \( f = \omega/2\pi \): The operating frequency.
2. \( f_1 = \omega_1/(2\pi) = 1/2\pi\sqrt{L_{ser}/C_{ser}} \): The resonant frequency in the main circuit.
3. \( f_2 = \omega_2/(2\pi) = 1/2\pi\sqrt{L_{inj}/C_{inj}} \): The resonant frequency in the auxiliary circuit.
4. \( A_1 = f_1/f = \omega_1/\omega \): The ratio of the resonant frequency in the main circuit to the operating frequency.
5. \( A_2 = f_2/f = \omega_2/\omega \): The ratio of the resonant frequency in the auxiliary circuit to the operating frequency.
6. \( B_1 = C_{ser}/C_{S1} \): The ratio of the resonant capacitance to the shunt capacitance in the main circuit.
7. \( B_2 = C_{inj}/C_{S2} \): The ratio of the resonant capacitance to the shunt capacitance in the auxiliary circuit.
8. \( H_1 = L_{ser}/L_{C1} \): The ratio of the resonant inductance to the dc-feed inductance in the main circuit.
9. \( H_2 = L_{inj}/L_{C2} \): The ratio of the resonant inductance to the dc-feed inductance in the auxiliary circuit.
10. \( Q_1 = \omega L_{ser}/R \): The loaded quality factor in the main circuit.
11. \( Q_2 = \omega L_{inj}/R \): The tentative loaded quality factor in the auxiliary circuit.
12. \( D_1 \): The switch-off duty ratio in the main circuit.
13. \( D_2 \): The switch-off duty ratio in the auxiliary circuit.

Next, the following assumptions are given for the proposed design of the class-E amplifier.

a. The switching devices \( S_1 \) and \( S_2 \) have infinite OFF resistance and ON resistance \( r_S \). The equivalent model of MOSFET is shown in Fig. 1(b).

b. Shunt capacitances of each switching device, namely \( C_{S1} \) and \( C_{S2} \) include switch device capacitances.

c. In this paper, \( r_{L_{C1}} \) and \( r_{L_{ser}} \) are defined as parasitic resistances of \( L_{C1} \) and \( L_{ser} \), respectively. The equivalent series resistances of all capacitance are neglected.

d. All passive elements including switch on resistances operate as linear elements.

e. Both of the switches turn off at \( \theta = 0 \) as shown in Fig. 4. The phase-shift of the driving signals is fixed.

B. Circuit Equations

We consider operations for \( 0 \leq \theta \leq 2\pi \) to design the circuit, where \( \theta = \omega t \) presents the angular time. Using the defined parameters, the circuit equations are expressed as follows:

\[
\begin{align*}
\frac{d\theta}{dt} & = \frac{H_1}{Q_1} \left( V_{DC\text{main}} - v_{S1} - r_{L_{C1}} i_{C1} \right) \\
\frac{di_{S1}}{dt} & = \frac{1}{Q_1} \left( v_{S1} - v_{ser} - (R + r_{L_{ser}}) i_o \right) \\
\frac{dv_{ser}}{dt} & = A_2^2 B_1 Q_1 R \left( i_{C1} - \frac{v_{S1}}{R_{S1}} - i_o + i_{inj} \right) \\
\frac{dv_{inj}}{dt} & = A_2^2 Q_1 R \left( i_{C1} - \frac{v_{S1}}{R_{S1}} - i_o \right) \\
\frac{dv_{S2}}{dt} & = A_2^2 B_2 Q_2 R \left( i_{C2} - \frac{v_{S2}}{R_{S2}} - i_{inj} \right) \\
\frac{dv_{inj}}{dt} & = A_2^2 Q_2 R \left( i_{C2} - \frac{v_{S2}}{R_{S2}} - i_{inj} \right)
\end{align*}
\]

In (3), \( R_{S1} \) and \( R_{S2} \) are the equivalent resistance of MOSFETs \( S_1 \) and \( S_2 \), respectively. The switch resistances \( R_{S1} \) and \( R_{S2} \) are given as follows.

\[
R_{S1} = \begin{cases} 
      r_{S1} & \text{for } 2\pi D_1 < \theta < 2\pi \\
      \infty & \text{for } 0 < \theta \leq 2\pi D_1
\end{cases}
\]

(4)

\[
R_{S2} = \begin{cases} 
      r_{S2} & \text{for } 2\pi D_2 < \theta < 2\pi \\
      \infty & \text{for } 0 < \theta \leq 2\pi D_2
\end{cases}
\]

(5)

When we define \( x(\theta) = [x_1, x_2, \ldots, x_8]^T = [i_{C1}, i_o, v_{S1}, v_{ser}, i_{C2}, i_{inj}, v_{S2}, v_{inj}]^T \in \mathbb{R}^8 \), Eq. (3) can be re-written as

\[
\frac{dx}{d\theta} = f(\theta, x, \lambda)
\]

(6)

where \( \lambda = [A_1, A_2, B_1, B_2, H_1, H_2, Q_1, Q_2, V_{DC\text{main}}, V_{DC\text{inj}}, R, r_{L_{C1}}, r_{L_{ser}}, r_{S1}, r_{S2}, \omega, D_1, D_2]^T \in \mathbb{R}^{18} \).

C. Conditions for the Design

We assume (3) has a solution \( x(\theta) = \varphi(\theta, x_0, \lambda) = [\varphi_1, \varphi_2, \ldots, \varphi_8]^T \) defined on \( -\infty < \theta < \infty \) with every initial condition \( x_0 \) and every \( \lambda : x_0 = \varphi(0, x_0, \lambda) \). The steady state of the amplifier is expressed by

\[
\varphi(\theta + 2\pi, x_0, \lambda) = \varphi(\theta, x_0, \lambda) \quad \forall \theta.
\]

(7)

Therefore,

\[
\varphi(2\pi, x_0, \lambda) - \varphi(0, x_0, \lambda) = 0 \in \mathbb{R}^8
\]

(8)

is given as the boundary conditions between \( \theta = 0 \) and \( \theta = 2\pi \). Moreover, we should consider the switching conditions of both the switches \( S_1 \) and \( S_2 \). About the switching conditions on \( S_1 \), (1) and (2) should be considered as a fundamental class-E operation. Additionally, the ZVS condition is given as a switching condition on \( S_2 \). Therefore, the following conditions are obtained.

\[
\begin{align*}
\varphi_2(2\pi D_1) & = 0, \\
\frac{d\varphi_2}{d\theta}(\theta) & = 0, \\
\varphi_1(2\pi) - \varphi_2(2\pi) + \varphi_0(2\pi) & = 0, \\
\frac{d(\varphi_1(\theta) - \varphi_2(\theta) + \varphi_0(\theta))}{d\theta} & = 0, \\
\varphi_2(2\pi D_2) & = 0.
\end{align*}
\]

From above considerations, we recognize that the design of the class-E amplifier boils down to the derivations of the resolutions of the algebraic equations (8) and (9). In these equations, there are 13 algebraic equations and 8 unknown initial values. Therefore, five parameters can be set as the design parameters from \( \lambda \in \mathbb{R}^{18} \). In this paper, the parameters \( A_1, A_2, B_1, B_2 \) and \( Q_2 \) are set as unknown parameters and the other parameters are given as the design specifications. The algebraic equations are solved by Newton’s method, which are described in [4] in detail. The derived results mean the design values, that is, \( A_1, A_2, B_1, B_2 \) and \( Q_2 \).

V. DESIGN CURVES

In this section, the design curves of the class-E amplifier with proposed design procedure are calculated and shown. First, the design specifications are given as follows: \( f = 3.5 \) MHz, \( V_{DC\text{main}} = 11.5 \) V, \( R = 13.5 \) Ω, \( D_1 = 0.5 \), \( D_2 = 0.25 \), \( Q_1 = 10 \), \( H_1 = 0.0358 \), \( H_2 = 0.0478 \). Additionally, \( r_S = 0.07 \) Ω is given since IRFZ24N MOSFETs are used as the switching devices. From the design specifications of \( Q_1, R, \) and \( H_1 \), the values of the inductances in the main circuit are determined as \( L_{C1} = 180 \) μH and \( L_{Cser} = 6.44 \) μH. Before the calculation, we make the inductors \( L_{C1} \) and \( L_{Cser} \) and measured their parasitic resistances as \( r_{L_{C1}} = 0.0127 \) Ω and \( r_{L_{ser}} = 0.421 \) Ω. We use Micrometals T130-2 as \( L_{Cser} \) and EI core as \( L_{C1} \). Parasitic resistances are measured by the impedance meter of 978-1-4244-1766-7/08/$25.00 ©2008IEEE 682
HP4284A. We use these values for the design calculations. The values are calculated by following design procedure in Sec. IV.

Figure 5 shows the design parameters $Q_2$, $A_1$, $B_1$, $A_2$, and $B_2$, the output power $P_o$ and the power conversion efficiency $\eta$ of the class-E amplifier with proposed design procedure as a function of $V_{DCinj}$. In these figures, the output power $P_o$ is given by

$$P_o = I_o^2 R.$$  (10)

Here, $I_o$ is the average root-mean-square of output current $i_o$

$$I_o = \sqrt{\frac{1}{2\pi} \int_0^{2\pi} (i_o(\theta))^2 d\theta}. $$  (11)

The power-conversion efficiency $\eta$ is defined as

$$\eta = \frac{P_o}{P_{main} + P_{inj}}. $$  (12)

Here, $P_{main}$ and $P_{inj}$ are input power in the main circuit and auxiliary circuit, given by

$$P_{main} = V_{DCmain}I_{C1}, \quad P_{inj} = V_{DCinj}I_{C2}. $$  (13)

$I_{C1}$ and $I_{C2}$ are the averages of the dc-supply current $i_{C1}$ and $i_{C2}$, respectively,

$$I_{Ck} = \frac{1}{2\pi} \int_0^{2\pi} i_{Ck}(\theta)d\theta, \quad \text{for } k = 1 \text{ and } 2 $$  (14)

For the calculations of the integrations in (11) and (14), we apply a trapezoidal method in this paper.

From Fig. 5(a), we can find that the $Q_2$ increases with the increase in $V_{DCinj}$. High loaded quality factor means that the power through the resonant circuit is limited. When the dc-supply voltage $V_{DCinj}$ becomes large, the $Q_2$ need become large. In particular, the $Q_2$ become large steeply after about $V_{DCinj} = 7.3$ V, namely $Q_2 = 10$. This means that the effects of the increase of $Q_2$ are small when $Q_2$ is large enough. From the calculation results, the maximum $V_{DCinj}$ is about 8.0 V.

From Figs. 5(b) and (c), $A_1$ and $A_2$ have the similar characteristics, that is, they become large as $V_{DCinj}$ increases. However, the characteristic of $B_1$ is different from that of $B_2$. Both $B_1$ and $B_2$ are the parameters for the shunt capacitances. However, the works of them are not identical. The shunt capacitance of the main circuit is adjusted to achieve the class-E switching conditions. On the other hand, that of the auxiliary circuit is for tuning the phase-shift of the injected current. Therefore, the design curves of them have different features. Moreover, Fig. 5(b) means that the design values of the main circuit is affected by $V_{DCinj}$, that is the design parameter of the auxiliary circuit. This result indicates that the class-E amplifier should be designed as one circuit, which is the important result in this paper.

From Fig. 5(d), the maximum output power $P_o$ is 13.5 W is obtained at $V_{DCinj} = 6.26$ V. For the range of $V_{DCinj} < 6.26$ V, the output power decreases drastically. In our calculation, the minimum of $V_{DCinj}$ is 4.62 V. If $V_{DCinj}$ is less than 4.62 V, the reverse current flows through the auxiliary circuit and output power is almost zero. Moreover, it can be said that the power conversion efficiency is independent of $V_{DCinj}$ because the power conversion efficiency indicates only the slight increase.

From these results, we should select $V_{DCinj}$ for proper $Q_2$ and the output power $P_o$. It is difficult to construct the real circuits for high $Q$ because of a large inductance and a sensitivity to the component tolerances.

VI. EXPERIMENTAL RESULTS

We show the design example and the results of the circuit experiments. The design specifications are the same as in Sec. V. From the design curves, $V_{DCinj} = 6.26$ V is given for maximum output power. From above design parameters, the element values are determined as shown in Tab. I. We use Micrometals T106-6 as $L_{Cinj}$ and EI core as $L_{C2}$. The dc-supply voltage and the dc-supply current are obtained from the digital multimeter of Iwatsu VOAC7532. The output voltage is obtained from the oscilloscope of Tektronix TDS3014B.

Figure 6 depicts the calculated waveforms and the experimental ones for the conditions in Tab. I. From Fig. 6, it is confirmed that the class-E amplifier with proposed design procedure satisfies all the conditions, namely the zero/zero-slope of voltage/current switching of the main circuit and the ZVS of the auxiliary amplifier. In this experiment, the amplifier with proposed design procedure achieves the 94.0 % power conversion efficiency under the 13.4 W output power and 3.5 MHz operation. From Fig. 6
TABLE I
Design values of the class-EM amplifier with proposed design procedure for the experiment

<table>
<thead>
<tr>
<th></th>
<th>Calculated</th>
<th>Measured</th>
<th>Difference</th>
</tr>
</thead>
<tbody>
<tr>
<td>$L_{G1}$</td>
<td>180 $\mu$H</td>
<td>180 $\mu$H</td>
<td>0.0 %</td>
</tr>
<tr>
<td>$L_{P2}$</td>
<td>4.44 $\mu$H</td>
<td>4.44 $\mu$H</td>
<td>0.0 %</td>
</tr>
<tr>
<td>$C_{S1}$</td>
<td>1140 pF</td>
<td>1140 pF</td>
<td>0.0 %</td>
</tr>
<tr>
<td>$C_{S2}$</td>
<td>321 pF</td>
<td>319 pF</td>
<td>0.6 %</td>
</tr>
<tr>
<td>$L_{T2}$</td>
<td>49.6 $\mu$H</td>
<td>49.8 $\mu$H</td>
<td>0.4 %</td>
</tr>
<tr>
<td>$r_{T2}$</td>
<td>2.37 $\mu$H</td>
<td>2.38 $\mu$H</td>
<td>0.4 %</td>
</tr>
<tr>
<td>$C_{S2}$</td>
<td>654 pF</td>
<td>654 pF</td>
<td>0.0 %</td>
</tr>
<tr>
<td>$C_{inj}$</td>
<td>253 pF</td>
<td>254 pF</td>
<td>0.4 %</td>
</tr>
<tr>
<td>$r_{inj}$</td>
<td>13.5 $\Omega$</td>
<td>13.4 $\Omega$</td>
<td>-0.7 %</td>
</tr>
<tr>
<td>$D_1$</td>
<td>0.5</td>
<td>0.5</td>
<td>0.0 %</td>
</tr>
<tr>
<td>$D_2$</td>
<td>0.25</td>
<td>0.25</td>
<td>0.0 %</td>
</tr>
<tr>
<td>$V_{DCmain}$</td>
<td>11.5 V</td>
<td>11.5 V</td>
<td>0.0 %</td>
</tr>
<tr>
<td>$V_{DCinj}$</td>
<td>6.26 V</td>
<td>6.27 V</td>
<td>0.2 %</td>
</tr>
<tr>
<td>$P_{0}$</td>
<td>13.5 W</td>
<td>13.4 W</td>
<td>-0.7 %</td>
</tr>
<tr>
<td>$P_{inj}$</td>
<td>11.1 W</td>
<td>11.1 W</td>
<td>0.0 %</td>
</tr>
<tr>
<td>$P_{off}$</td>
<td>3.11 W</td>
<td>3.15 W</td>
<td>1.3 %</td>
</tr>
<tr>
<td>$\eta$</td>
<td>94.7 %</td>
<td>94.0 %</td>
<td>-0.7 %</td>
</tr>
</tbody>
</table>

Fig. 6. Waveforms of the amplifier with proposed design procedure for $Q_1 = 10$. (a) Calculated waveforms. (b) Experimental waveforms. Vertical: $D_{13}, D_{25}$:5V/div, $v_{S1}, v_{S2}$: 20V/div, $i_{S1}$: 2A/div, and $v_{o}$: 20V/div. Horizontal: 100ns/div.

and Tab. I, the experimental results are agree with the calculated ones quantitatively, which denotes the validity of the design curves in Section V.

VII. CONCLUSION

This paper has presented a novel design procedure for Class-EM amplifier. To improve the whole power conversion efficiency, the ZVS class-E doubler is applied to the auxiliary circuit. Moreover, the proposed procedure is based on that in [4]. As a result, it is possible to design both the main circuit and the auxiliary one as one circuit by using the proposed procedure. In addition, the derived element values have high accuracy for achieving all the switching conditions simultaneously. The laboratory experiment shows the class-EM amplifier with proposed design procedure achieves the 94.0 % power conversion efficiency under the 13.4 W output power and 3.5 MHz operation.

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