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Modification of Cockcroft–Walton-Based High-Voltage Multipliers with 220 V and 50 Hz Input for Non-Thermal Food Processing Apparatus

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Abstract: A design of high-voltage multipliers to generate underwater shockwaves is one of the most important factors for successfully providing non-thermal food processing in a cost-effective manner. To be capable of fully utilizing the Cockcroft–Walton-based high-voltage multipliers for underwater shockwave generation, this paper presents a topological modification of three interesting design approaches in bipolar structure for 220 V and 50 Hz AC input to generate more than 3.5 kV DC output within short time periods. In addition to Cockcroft–Walton multipliers (CWMs), the first modified scheme employs a positive full-wave rectifier (FWR) and positive voltage multiplier block (VMB), the second modified scheme employs positive/negative half-wave rectifiers (HWRs), and the last modified scheme employs a switched-capacitor AC-AC converter. To comparatively analyze their performances, the digitally controlled operations of the modified realization schemes as well as their electrical characteristic estimation based on a four-terminal equivalent model are described in detail. The effectiveness of three modified circuit configurations and the correctness of the given theoretical analysis are verified through SPICE (Simulation Program with Integrated Circuit Emphasis) simulation results. The formulas achieved from theoretical estimation are particularly useful when designing the proposed high-voltage multipliers (HVMs) because good agreement between the theoretical and simulation results can be achieved.

Keywords: bipolar structure; Cockcroft–Walton multiplier; full-wave rectifier; high-voltage multiplier; non-thermal food processing; switched-capacitor circuit; underwater shockwave

1. Introduction

It is widely accepted that underwater shockwaves produced by high electric discharge can be effectively used for non-thermal food processing with minimal impact on nutritional properties and at low costs because several benefits, such as short processing time and low energy consumption, can be achieved [1–5]. The underwater shockwave can be generated by utilizing gap discharge [6,7], wire explosion [8,9], wire discharge of dual electrodes [10], or two pairs of restoration electrodes [11]. A typical non-thermal food processing system using underwater shockwaves consists of a high-voltage generator, big capacitor, high-voltage switch, and pressure vessel [10,12]. Generally, a high-voltage multiplier (HVM) that can produce a high DC voltage such as 3.5 kV from a low AC voltage is utilized as the high-voltage generator. The big capacitor acts as a discharging circuit to produce a spark between two electrodes, which are placed in water stored inside the pressure vessel. The high-voltage switch is employed to control the charging/discharging status by connecting/disconnecting the big capacitor

to/from the HVM. The output voltage of the HVM is charged to the big capacitor when turning off the high-voltage switch. On the other hand, in the case of turning on the high-voltage switch, the electric charge stored in the big capacitor is discharged from two electrodes to generate the shockwave in the pressure vessel. The target food is then processed by the mechanism of spall fracture [6] or spalling destruction [13].

The HVM, as well as its discharge technique, is obviously an important component in underwater shockwave generation [10,11]. For example, an HVM that can produce underwater shockwaves at high speed can provide high production capacity for non-thermal food processing [12]. The other basic requirements for good HVMs used in non-thermal food processing are small size, light weight, and high voltage efficiency [14]. One of the most famous switched-capacitor approaches that can usually be employed to realize HVMs is through the use of the traditional Cockcroft–Walton multiplier (CWM), which is an arrangement of capacitors and diode switches in various stages and in different capacitance values [15–18]. The key attractive feature of the half-wave CWM is that low-voltage components can be used in circuit realization because voltages across each capacitor and diode switch do not exceed two times the maximum value of the AC input voltage [19]. This causes the size of the CWM circuit configuration to be smaller than that of a voltage multiplier utilizing high-voltage components. However, there are two major limitations in the designed circuit with the traditional CWM. The first limitation is that increasing the number of multiplier stages increases output voltage drop under load conditions because the voltage drops in all previous connecting capacitors are added up at the output terminal [20]. The second limitation is that the operational speed of the traditional CWM is slow because it is driven by a sinusoidal waveform applied by a commercial power supply [21]. Various design techniques to overcome these limitations have been proposed [22–29]. For example, to provide high voltage gain, transformer-less voltage multipliers made by combining the CWM with a boost-converter have been presented in 2013 [22] and in 2015 [23]. In addition, voltage multipliers in a bipolar topology and hybrid symmetrical topology have been introduced in 2007 [24] and 2014 [25], respectively, to minimize the voltage drop as well as to enhance the transient response. The bipolar voltage multiplier in [24] is realized by using two multipliers with opposite polarities—a positive multiplier and a negative multiplier. The output of the bipolar structure is the sum of the positive multiplier output and the negative multiplier output. These positive/negative multipliers used are driven in parallel by a commercial AC voltage source. The hybrid symmetrical voltage multiplier in [25] is realized by a cascading diode-bridge rectifier and symmetrical voltage multiplier. However, a high-voltage transformer is required to implement these two voltage multipliers. Thus, the usefulness of the bipolar and hybrid symmetrical voltage multipliers is rather limited. Alternatively, inductor-less HVMs in bipolar structure have been suggested in 2014 [26], in 2015 [27,28], and in 2017 [29]. The status of diodes placed in these bipolar HVMs is digitally controlled by rectangular pulses to produce a high DC output voltage at high speed. The bipolar HVM in [26] consists of two switched-capacitor-based DC-DC converters in parallel, while the bipolar HVM in [27] consists of two positive CWMs and two negative CWMs in parallel. The bipolar HVM in [28] consists of two positive/negative CWMs in series. In order to reduce the number of circuit components, an HVM designed by combining a bipolar multiplier and the AC-AC converter has been proposed in 2017 [29]. Nevertheless, the HVMs in [26–29] have been introduced for generating the output voltage between 3.5 to 4.0 kV by supplying 100 V and 50 Hz or 100 V and 60 Hz input. If there exists an HVM which can be applied with other AC inputs, then there are advantages of non-thermal food processing by using underwater shockwaves to be gained.

The goal of this paper is to propose a topological modification of three interesting design techniques for realizing bipolar HVMs with 220 V and 50 Hz input to produce the DC output between 3.5 and 4.0 kV, which is suitable to generate underwater shockwaves for non-thermal food processing apparatus. Three proposed HVMs, called HVM-A, HVM-B, and HVM-C, were modified from the CWM-based circuit configurations in [10,12,28,29], respectively. The design of the circuit topology of each HVM for 220 V and 50 Hz input requires considerable modification of the sub-circuits used for performing

multiplication or amplification because the input signal requirement is changed. The electrical characteristics of each modified scheme can be estimated and compared from the results of theoretical analysis by using a four-terminal equivalent model. Moreover, simulation results verifying their performances are also given.

The rest of this paper is organized as follows. The proposed modification to realize the HVMs and theoretical analysis to illustrate the performances are detailed in Sections 2 and 3, respectively. Simulation results to verify the validity of the proposed modification are shown in Section 4. Lastly, the discussion, conclusions, and possible future work are given in Section 5.

2. Proposed Modification

Figure 1 shows the topological diagrams of three modified HVMs in bipolar structure. Their sequential operations are digitally controlled by non-overlapping two-phase clock pulses, where T_1 and T_2 are the 'on' periods of the first and second clock signals, respectively, and $T = T_1 + T_2$ is the period of clock signals. The parameters V_{in} and $V_{out} = V_{po} - V_{no}$ are the input peak value and the output peak value, respectively, V_{po} is the positive output, and V_{no} is the negative output. In order to convert the 220 V and 50 Hz input into the DC output between 3.5 and 4.0 kV, the step-up gain of three proposed HVMs is set to 14. Each proposed HVM is designed to provide two amplification stages. The gains of the first and second amplification stages are equal to 2 and 7, respectively.

Figure 1. Cont.

Figure 1. Modified schemes of three proposed high-voltage multipliers (HVMs): (a) HVM-A; (b) HVM-B; (c) HVM-C.

The circuit configuration of the first proposed HVM, referred to as HVM-A, is shown in Figure 1a. This modified scheme is based on the design technique suggested in [28]. The main parts are the positive full-wave rectifier (FWR), the positive voltage multiplier block (VMB), and the parallel-connected positive/negative CWMs. The positive VMB, positive CWM, and negative CWM have the gains of 2, 5, and 2, respectively. The DC input provided by the positive FWR is amplified by the positive VMB, and is then amplified again by the parallel-connected positive/negative CWMs. This means that the HVM-A provides two amplification stages. Figure 1b shows the circuit realization of the second proposed HVM, referred to as HVM-B, which is based on the design technique introduced in [10,12]. The main parts of the HVM-B are the positive/negative HWRs (or AC-DC rectifier) and the parallel-connected positive/negative CWMs. The gains of the AC-DC rectifier, positive CWM, and negative CWM are 2, 5, and 2, respectively. The DC input is provided and amplified by the AC-DC rectifier, and is then amplified again by the parallel-connected positive/negative CWMs. Similarly, the HVM-B exhibits two amplifications. The circuit configuration of the third proposed HVM, referred to as HVM-C, is depicted in Figure 1c. This configuration is based on the design technique proposed in [29]. The HVM-C consists of the switched-capacitor AC-AC converter and the positive/negative CWMs. The AC-AC converter, positive CWM, and negative CWM provide the gains of 2, 5, and 2, respectively. The AC input is amplified by the AC-AC converter, and is then amplified again and converted into the DC output by the positive/negative CWMs. Thus, the HVM-C also provides two amplifications.

Table 1 summarizes the outputs V_{po} , V_{no} , and V_{out} of all proposed HVMs, where V_{th} denotes the forward threshold voltage of the diode. It is seen that the voltage outputs of the HVM-A and HVM-B are most affected by the diode switch non-idealities, whereas the voltage outputs of the HVM-C are less affected by the diode switch non-idealities. However, neglecting the small voltage V_{th} will cause the high DC output V_{out} of the three proposed HVMs to closely rise to $14V_{in}$. Hence, the step-up gain of 14 can be achieved.

Table 1. Comparison of the positive output, negative output, and output voltage.

Modified Scheme	V_{po}	$-V_{no}$	V_{out}
HVM-A	$V_{po} = 10V_{in} - 11V_{th}$	$-V_{no} = 4V_{in} - 4V_{th}$	$V_{out} = 14V_{in} - 15V_{th}$
HVM-B	$V_{po} = 10V_{in} - 11V_{th}$	$-V_{no} = 4V_{in} - 4V_{th}$	$V_{out} = 14V_{in} - 15V_{th}$
HVM-C	$V_{po} = 10V_{in} - 6V_{th}$	$-V_{no} = 4V_{in} - 3V_{th}$	$V_{out} = 14V_{in} - 9V_{th}$

3. Theoretical Analysis

In order to provide ease of estimating and comparing the electrical characteristics of the proposed HVMs in practice, the simplified equivalent model, as shown in Figure 2, is theoretically derived. This equivalent model is a general equivalent circuit of capacitor-based converters [30,31]. It is composed of an ideal transformer with turn ratio (1: M), an internal resistance R_{SC} , and an output load R_L . To save space, only the analysis of the proposed HVM-A is described in detail under the following assumed conditions.

- (1) The dielectric loss is not considered.
- (2) The time constant of the proposed HVM-A is greater than the time period of clock pulses.
- (3) The input is a rectangular waveform.

Figure 2. Simplified equivalent model of the proposed HVMs.

The instantaneous equivalent circuits of HVM-A for State- T_{ri} and State- T_{mi} ($i = 1, 2$) are shown in Figure 3a–d, respectively. The diode switch is represented as an ideal diode, an on-resistance (R_d), and a voltage source (V_{th}), while the transistor switch is modeled by the on-resistance (R_{on}). In steady state, the differential values of electric charges in C_{0pk} ($k = 1, 2$), C_{1pq} , C_{2pq} ($q = 1, 2, 3, 4$), and C_{1nr} , C_{2nr} ($r = 1, 2$) can be written as

$$\sum_{i=1}^2 \Delta q_{T_{ri}}^{0pk} = 0, \sum_{i=1}^2 \Delta q_{T_{mi}}^{1pq} = 0, \sum_{i=1}^2 \Delta q_{T_{mi}}^{2pq} = 0, \sum_{i=1}^2 \Delta q_{T_{mi}}^{1nr} = 0, \text{ and } \sum_{i=1}^2 \Delta q_{T_{mi}}^{2nr} = 0, \quad (1)$$

where $\Delta q_{T_{ri}}^{0pk}$, $\Delta q_{T_{mi}}^{1pq}$, $\Delta q_{T_{mi}}^{2pq}$, $\Delta q_{T_{mi}}^{1nr}$, and $\Delta q_{T_{mi}}^{2nr}$ denote the differential values of electric charges of the k -th, q -th, and r -th capacitors in State- T_{ri} and State- T_{mi} , respectively, and the time interval of clock pulses satisfies the following conditions:

$$T_r = \sum_{i=1}^2 T_{ri}, \quad T_m = \sum_{i=1}^2 T_{mi} \text{ and } T_{r1} = T_{r2} = \frac{T_r}{2}, \quad T_{m1} = T_{m2} = \frac{T_m}{2}. \quad (2)$$

Figure 3. Cont.

Figure 3. Instantaneous equivalent circuits of HVM-A: (a) State- T_{r1} ; (b) State- T_{r2} ; (c) State- T_{m1} ; (d) State- T_{m2} .

The time interval T_r depends on the frequency of the AC input V_{in} , and the time interval T_m depends on the switching frequency for driving the transistor switch. Therefore, the relationship of differential values of electric charges in the input and output terminals ($\Delta q_{T_{r1}}, V_{in}^1, \Delta q_{T_{r2}}, V_{in}^1, \Delta q_{T_{m1}}, V_{in}^2, \Delta q_{T_{m2}}, V_{in}^2, \Delta q_{T_{r1}}, V_o, \Delta q_{T_{r2}}, V_o, \Delta q_{T_{m1}}, V_{po}, \Delta q_{T_{m2}}, V_{po}, \Delta q_{T_{m1}}, V_{no},$ and $\Delta q_{T_{m2}}, V_{no}$) for the symmetrical circuit structure in State- T_{ri} and State- T_{mi} can be expressed as

$$\text{State-}T_{r1} \quad \Delta q_{T_{r1}}, V_{in}^1 = -\Delta q_{T_{r1}}^{0p1} + \Delta q_{T_{r1}}^{0p2} \text{ and } \Delta q_{T_{r1}}, V_o = \Delta q_{T_{r1}}^{0p1} \quad (3)$$

$$\text{State-}T_{r2} \quad \Delta q_{T_{r2}}, V_{in}^1 = +\Delta q_{T_{r2}}^{0p1} - \Delta q_{T_{r2}}^{0p2} \text{ and } \Delta q_{T_{r2}}, V_o = \Delta q_{T_{r2}}^{0p2} \quad (4)$$

$$\begin{aligned} \text{State-}T_{m1} \quad \Delta q_{T_{m1}}, V_{in}^2 &= -\Delta q_{T_{m1}}^{1p1} + \Delta q_{T_{m1}}^{2p1} - \Delta q_{T_{m1}}^{2p2} - \Delta q_{T_{m1}}^{2p3} + \Delta q_{T_{m1}}^{1n1} + \Delta q_{T_{m1}}^{2n2}, \\ \Delta q_{T_{m1}}, V_{po} &= \Delta q_{T_{m1}}^{2p4} \text{ and } \Delta q_{T_{m1}}, V_{no} = -\Delta q_{T_{m1}}^{2n2} \end{aligned} \quad (5)$$

$$\begin{aligned} \text{State-}T_{m2} \quad \Delta q_{T_{m2}}, V_{in}^2 &= \Delta q_{T_{m2}}^{1p1} - \Delta q_{T_{m2}}^{1p2} - \Delta q_{T_{m2}}^{1p3} - \Delta q_{T_{m2}}^{2p1} + \Delta q_{T_{m2}}^{1n2} + \Delta q_{T_{m2}}^{2n1}, \\ \Delta q_{T_{m2}}, V_{po} &= \Delta q_{T_{m2}}^{1p4} \text{ and } \Delta q_{T_{m2}}, V_{no} = -\Delta q_{T_{m2}}^{2n2}. \end{aligned} \quad (6)$$

By using Kirchhoff's current law, the input and output currents can respectively be written as

$$\overline{I_{in}} = \frac{1}{T} \left(\sum_{i=1}^2 \Delta q_{T_i, V_{in}} \right) = \frac{\Delta q_{V_{in}}}{T} \text{ and } \overline{I_{out}} = \frac{1}{T} \left(\sum_{i=1}^2 \Delta q_{T_i, V_{out}} \right) = \frac{\Delta q_{V_{out}}}{T}. \quad (7)$$

Substituting (3)–(6) into (7), the relation between the input current and the output current can be given by

$$\overline{I_{in}}^1 = -2\overline{I_{out}}^1, \overline{I_{out}}^1 = \overline{I_{po}}^1 = -\overline{I_{no}}^1 \text{ and } \overline{I_{in}}^2 = -7\overline{I_{out}}^2, \overline{I_{out}}^2 = \overline{I_{po}}^2 = -\overline{I_{no}}^2. \quad (8)$$

From (8), the turn ratios M_1 and M_2 of Figure 2 are equal to 2 and 7, respectively. Thus, the consumed energy W_T in one period of State- T_{ri} and State- T_{mi} can be expressed as

$$\begin{aligned} W_{T_r} &= W_{T_{r1}} + W_{T_{r2}} = \frac{2R_d}{T_r} (\Delta q, V_{out})^2 + \frac{20R_{on}}{T_r} (\Delta q, V_{out})^2, \\ \text{and} \quad W_{T_m} &= W_{T_{m1}} + W_{T_{m2}} = \frac{16R_d}{T_m} (\Delta q, V_{out})^2 + \frac{164R_{on}}{T_m} (\Delta q, V_{out})^2. \end{aligned} \quad (9)$$

From (9), the consumed energy W_T and the internal resistances R_{SC1} and R_{SC2} of the four-terminal equivalent model under the defined conditions can respectively be obtained as

$$W_T = R_{SC} \cdot \frac{(\Delta q_{V_{out}})^2}{T} \quad (10)$$

$$\text{and} \quad R_{SC1} = 2R_d + 20R_{on} \text{ and } R_{SC2} = 16R_d + 164R_{on}. \quad (11)$$

From (11), the turn ratio (1: M) and internal resistance R_{SC} can be calculated as

$$M = 14 \quad \text{and} \quad R_{SC} = 114R_d + 1144R_{on}. \quad (12)$$

Therefore, the equivalent model of the proposed HVM-A can be expressed by the following matrix:

$$\begin{bmatrix} \overline{V_{in}} \\ \overline{I_{in}} \end{bmatrix} = \begin{bmatrix} 1/14 & 0 \\ 0 & 14 \end{bmatrix} \begin{bmatrix} 1 & 114R_d + 1144R_{on} \\ 0 & 1 \end{bmatrix} \begin{bmatrix} \overline{V_{out}} \\ -\overline{I_{out}} \end{bmatrix}. \quad (13)$$

From (13), the output voltage V_{out} , the output power P_{out} , and the power (or voltage) efficiency η for the proposed HVM-A of Figure 1a can be approximated to

$$V_{out} = \eta \times 14 \times V_{in}, \quad P_{out} = \frac{(V_{out})^2}{R_L}, \quad \text{and} \quad \eta = \frac{R_L}{R_L + (114R_d + 1144R_{on})}. \quad (14)$$

From the above expressions, it should be noted that the outputs V_{out} and P_{out} as well as the power efficiency η are affected by the resistance values of R_d , R_{on} , and R_L . The instantaneous equivalent circuits of the proposed HVM-B and HVM-C are depicted in Figures 4 and 5, respectively. The theoretical analysis procedures for these two modified schemes follow the same analysis steps of the proposed HVM-A. The theoretical analysis results for comparing the internal resistance R_{SC} of each modified scheme in the case of the step-up gain of 14 are summarized in Table 2.

Figure 4. Instantaneous equivalent circuits of HVM-B: (a) State- Tr_1 ; (b) State- Tr_2 ; (c) State- Tm_1 ; (d) State- Tm_2 .

Figure 7. Variations of the output voltage with the output power.

Figure 8. Variations of the power efficiency with the output power.

Figure 6 illustrates the simulated and calculated values of the power efficiency of the HVM-A, HVM-B, and HVM-C when varying the value of the output load R_L in the range of 0.01–1000 k Ω . In the case of setting $R_L = 5$ k Ω , the simulated (calculated) values of the power efficiency of the proposed HVM-A, HVM-B, and HVM-C are approximately equal to 79.73% (80.19%), 93.85% (94.31%), and 92.16% (92.62%), respectively. In order to obtain the power efficiency greater than 80%, the values of R_L for the HVM-A, HVM-B, and HVM-C should be greater than 6, 2, and 2 k Ω , respectively.

Figure 7 shows the variations of the output voltage with the output power, while Figure 8 shows the variations of the power efficiency with the output power. The HVM-A, HVM-B, and HVM-C provide the simulated (calculated) output power values of about 2.15 (2.17 kW), 7.08 (7.16 kW), and 6.52 kW (6.59 kW), respectively, when generating the output voltage of about 3.5 kV (or the power efficiency of about 80%). The average errors between the simulated values and the calculated values for the HVM-A, HVM-B, and HVM-C are approximated to 3.51%, 1.15%, and 1.98%, respectively. It is evident that the results from the simulations agree well with the results from the calculations from (14).

For comparison with the previous work, Figure 9a–c represents the simulated output voltages of the proposed HVM-A, HVM-B, and HVM-C, respectively, while Figure 9d represents the simulated output of the traditional CWM with seven stages [16]. In the performed simulations, the output load R_L , shown in Figure 9, was selected to generate the target output voltage, such as 3.5 kV. From Figure 9a, the HVM-A can generate an output voltage of 3.55 kV within 600 ms when setting $R_L = 6$ k Ω .

The HVM-B can produce the output voltage of 3.69 kV within 200 ms in the case of $R_L = 2\text{ k}\Omega$, as shown in Figure 9b. From Figure 9c, the HVM-C can produce the output voltage of 3.58 kV within 1.8 s for setting $R_L = 5\text{ k}\Omega$. The traditional half-wave CWM [16] can generate an output voltage of 3.55 kV within 190 s for $R_L = 1500\text{ k}\Omega$, as illustrated in Figure 9d. It is obvious that the HVMS with different schemes can convert the 220 V and 50 Hz input into the DC output to have more than 3.5 kV within different settling time periods. This implies that the response speed values of the proposed HVMS and the traditional CWM [16] are different.

Figure 9. Simulated output voltages: (a) HVM-A; (b) HVM-B; (c) HVM-C; (d) Cockcroft–Walton multiplier (CWM) [16].

5. Discussion and Conclusions

To generate underwater shockwaves using electrical energy charge for non-thermal food processing, DC voltage in the range of 3.5–4.0 kV is required. The basic requirements for a good high-voltage multiplier (HVM) are high response speed, high voltage efficiency, and small number of components used. For utilizing the conventional Cockcroft–Walton-based HVMS with 100 V and 50 Hz or 100 V and 60 Hz proposed in [10,12,28,29] in the 220 V and 50 Hz power system, this paper presents a technique to modify their circuit realizations to be the new schemes that can apply the 220 V and 50 Hz input to produce the high DC output in the desired range.

Table 3 summarizes the realization methods of the three proposed HVMS (HVM-A, HVM-B, and HVM-C). The main sub-circuits for each modified scheme are given. Each modified scheme provides the step-up gain of 14. All proposed HVMS provide two cascaded stages to amplify the input signal, which are referred to as the first multiple block and second multiple block. The first amplification stage has a step-up gain of 2, while the second amplification stage has a step-up gain of 7. In order to enhance their response speed through bipolar structure realization [28,29], the output of the proposed HVMS can be provided by combining the outputs of the positive CWM and the negative CWM.

Table 3. Summary of main parts used in the proposed HVMs.

Modified Scheme	1st Multiple Block	2nd Multiple Block
HVM-A	Positive FWR Positive VMB	Parallel-Connected Positive CWM Parallel-Connected Negative CWM
HVM-B	Positive/Negative HWRs (AC-DC Rectifier)	Parallel-Connected Positive CWM Parallel-Connected Negative CWM
HVM-C	Positive/Negative VMBs (SC AC-AC Converter)	Positive CWM Negative CWM

Table 4 shows the number of components used in the proposed HVMs and the traditional half-wave Cockcroft–Walton multiplier with seven stages [16], which include diodes, capacitors, and switches. It is seen that the proposed HVM-C uses the minimum number of devices in circuit implementation. On the other hand, the proposed HVM-A uses the maximum number of devices in circuit realization. By considering only the number of diodes used, it is found that the HVM-A and HVM-B consist of the highest number of diodes, whereas the HVM-C consists of the lowest number of diodes. The numbers of diodes shown in Table 4 support the comparison of the output voltages shown in Table 1. Moreover, the results from comparing the size, internal resistance, response speed, and power efficiency among the proposed HVMs and the traditional CWM [16] are summarized in Table 5. It is seen that not only the highest speed operations, but also the highest power (or voltage) efficiency can be provided in the proposed HVM-B. However, the HVM-A also provides a settling time of less than 1 s. From Table 5, among the three proposed HVMs, it is seen that the power efficiency is strongly dependent on the internal resistance R_{SC} , which is in good agreement with the results shown in Table 2. By considering in terms of size of the circuit, the proposed HVM-C can be realized by using the smallest number of components. However, the settling time of the HVM-C is slower than that of HVM-A and HVM-B because the frequency of the input source is fixed at 50 Hz.

Table 4. Comparison of the numbers of components used in the proposed HVMs.

Modified Scheme	Diode	Capacitor	Switch	Total
HVM-A	18	16	8	42
HVM-B	18	14	4	36
HVM-C	8	8	4	20
Traditional CWM [16]	14	14	-	28

Table 5. Summary of comparison results.

Modified Scheme	Size	Internal Resistance	Response Speed	Efficiency
HVM-A	4	4	2	3
HVM-B	3	2	1	1
HVM-C	1	3	3	2
Traditional CWM [16]	2	1	4	4

1: Best–4: Worst.

By converting the 220 V and 50 Hz input, all three proposed HVMs can produce the high DC output in the desired range of 3.5–4.0 kV. However, their electrical characteristics are different. This means that they can respond to the different requirements. The formulas given in this paper can be effectively used to estimate the circuit performances when designing the modified schemes because it appears from all figures that the simulation results are in good agreement with the theoretical values.

Lastly, to confirm the versatility of the proposed modification, the hardware implementation by using the modified high-voltage multiplier to generate underwater shockwaves for non-thermal food processing is the direction for future work.

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