Abstract—Replacing the batteries of wearable medical devices is an inconvenience to the user. Wireless power transfer (WPT) could be a solution for this problem. However, the structure should be compact to satisfy the wearable device requirements and the power efficiency should be as high as possible for prolonging the user experience. Besides, as algorithms embedded in these devices become more complex, the device power consumption also increases. Fulfilling these specifications in a compact structure with better efficiency is the aim of this research work. This paper proposes a power processing module for WPT applications. This module is embedded within the WPT path to convert receiving end AC power to DC power, and to simultaneously provide multiple DC-DC power conversion for wearable medical devices. The AC-DC circuit is specifically designed using a compact topology to convert wireless AC power to stable DC power. The DC power is further regulated by different mode controls either to perform battery charging or to provide independent voltage supplies to DC loads. The integrated circuit occupies an area of 1.9 mm × 2.3 mm when implemented on a 0.25 µm high voltage CMOS process.

Index Terms—Rectifier, AC-DC power converter, wearable medical devices, battery charger, PWM rectifier, transcutaneous power transmission, wireless power transfer.

I. INTRODUCTION

WEARABLE medical devices have become promising alternatives to monitor, and treat Parkinson’s disease, epilepsy and a plethora of other diseases and disorders [1]-[4]. The devices in medical application are required to be small volume, low noise, and high efficiency. However, there would be several challenges to take these three considerations into account. For example, power hungry device could increase the overall circuit volume to maintain good voltage regulation. If one want to reduce the volume, operating frequency should be raised. This strategy would cause high frequency noise problem and drop of power efficiency due to the switching loss. In wireless applications, reliable power to support the wearable medical devices is further critical due to the extremely stringent power and high availability requirements. Power consumption of wearable medical devices varies significantly based on the target function. Recent designs require calculations for signal processing and/or continuous monitoring via cloud-based healthcare services. Power consumption of such circuits is reported to be in the range between a couples of mWs to 30 W [5]. Several medical microsystems consume 10-30 mW or higher [6]. For instance, low power microprocessors integrated on chip for real-time epileptic seizure detection, can have maximum current consumption reaching 22.7 mA [7].

Wireless transcutaneous power converter (WPT) is a notable wireless power supply solution either for directly powering the device or for transcutaneously charging an implanted rechargeable battery [8]-[18]. Since WPT adopts AC power to transfer energy through magnetic coils, AC-DC conversion is essential. The conventional topology of AC-DC power path for medical devices usually has three stages: (1) rectifier, (2) DC-DC switching converter, and (3) linear regulator. To fit the small volume requirement, the rectifier stage can adopt the bootstrapping technique [8]. Because this kind of rectifier has no voltage or current regulation capability, authors in [9] - [11] propose to use linear regulators and voltage limiters to realize a continuous powering system for medical devices. Regarding the charging method, [12]-[15] use linear regulators after the rectifier to control the charging current and to reduce the voltage ripple. Since the efficiency of the linear regulator is highly dependent upon voltage drop between input and output, high efficiency is not easy to achieve for linear regulators to deal with high voltage drop. To further improve efficiency, DC-DC buck type converters were placed between the rectifier and the linear regulator [15], [16]. Although switching converters improve efficiency in DC-DC conversion, the power line still needs a rectifier to convert the AC source to a DC source. From the overall power line point of view, one more circuit stage implies more power conduction loss and larger circuit layout. Naturally, the idea of merging the three-stage structure into a single-state structure is attractive. However, there are challenges in designing this single-stage structure. One challenge is the design of the controller, which needs to coordinate multiple switches. The other challenge is the
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This paper is organized as follows. Section II presents the architecture of the proposed AC-DC converter and circuit operation. In Section III, the stability of the proposed architecture is investigated. Section IV presents the circuit implementation and the performance results. Section V draws concluding remarks for this work.

II. ARCHITECTURE AND CIRCUIT OPERATION

A. Conventional Power Path Structure for Medical Devices

Fig. 1 demonstrates an example of the conventional three-stage WPT topology. The 14 V_{pp} AC power is first transmitted through the wireless link by electromagnetic induction. The transmitted AC power is then rectified to 5 V DC power. The 5V DC value is supposed to be further regulated by a linear regulator for charging the battery whose voltage ranges between 3 V and 4.2 V. The overall efficiency of this circuit arrangement is low because of the high voltage-dropout between input and output of the linear-regulator-type charger. To solve this high dropout voltage issue, a step-down switching converter is supplemented between the rectifier and the linear regulator. The linear regulator consequently acts as a current source to control the charging current. This topology improves the efficiency of the linear regulator because it only regulates a very small voltage drop (about 0.3 V) in this case. The overall efficiency of the aforementioned structure is around 71.2%. This estimation is based on the 89% efficiency of the rectifier [20], 86% efficiency of a step-down switching converter[16], [21] and [22], and 93% efficiency of a linear regulator for the case of 4.2 V output voltage and 0.3 V dropout voltage.

The reason of using a linear regulator in Fig. 1 is its simple mechanism to control the charging current. If the Buck converter is able to regulate voltage and give CV and CC modes of charging, there is no reason to use the linear regulator. Thus, the linear regulator can be removed to boost the overall circuit efficiency. However, using the topology proposed in [16] for WPT purpose still needs a rectifier at its front end to convert AC to DC power. In order to further reduce the size of the overall power converter, this paper attempts to eliminate the rectifier stage and realize the rectification function in the Buck converter. To integrate rectifier and charger functions into a single Buck converter, more efforts should be made in the controller design. For example, a current sensing loop should be supplemented to deal with CC mode operation. Different switching mechanism to sustain either CV or CC mode should be carefully coordinated. Furthermore, to perform the rectifier function during battery charging, the coordination of the four switches inside the proposed Buck to rectify AC power to DC power is also a challenge.

B. Proposed single-stage AC-DC converter module

To reduce the multi-stage power conversion loss, Fig. 2 demonstrates a single-stage AC-DC converter that is connected to a battery and a single-inductor dual-output (SIDO) converter. Compared to the three-stage power path, the efficiency of the single-stage AC-DC converter is expected to be higher. The structure of the converter is also compact since several circuit components are eliminated including one capacitor from the rectifier, one switch as well as one diode from the step-down switching converter, and one MOSFET from the linear regulator.

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inductor L and a capacitor C. The output voltage \( V_{out} \) is filtered by the output filtering circuit, which includes MP1 and MP2. The average value of \( V_x \). The output voltage level can be adjusted by the switch to deliver power to the node \( V_x \). When MP2 turns off, the period. During the AC negative period, MP2 becomes the major action of the aforementioned action in every cycle.

### III. Stability Analysis

#### A. AC-DC Converter Power Stage Model

The single-stage structure can be modelled as a step-down switching converter because the frequency of the input AC power is much lower than the switching frequency. The slow-changing AC power is viewed as a DC input for the circuit. The operating principle shown in Fig. 3 is similar to that of a step-down switching converter. Inside the step-down switching converter, the inductor energy is transferred to the load. At that moment, MP7 and MP8 are turned on. When the energy is charged high enough for phase A, MP8 is turned off, and MP9 is then turned on. Charging would continue until the total energy for both outputs A and B is enough. Then, circuit alternates switches to discharge the inductor energy to the load. At that moment, MP7 is turned off, and MN3 is turned on. The circuit repeats the sequence of the aforementioned action in every cycle.

During the AC positive period, the MOSFET MP1 first turns on to pass the power from AC source to \( V_x \). This behavior is shown in Fig. 3(a). When MP1 turns off, MP2 turns on to keep the current flowing as shown in Fig. 3(b). These two actions keep on repeating several times until the end of the AC positive period. During the AC negative period, MP2 becomes the major switch to deliver power to the node \( V_x \). When MP2 turns off, MP1 turns on to keep current flowing. These two steps are shown in Fig. 3(c) and Fig. 3(d). The resulting voltage waveform \( V_x \) is shown in Fig. 3(e). The voltage \( V_x \) is then filtered by the output filtering circuit, which includes an inductor \( L \) and a capacitor \( C \). The output voltage \( V_{out} \) is the average value of \( V_x \). The output voltage level can be adjusted by changing the pulse-width of the PWM signal that drives MP1 and MP2.

A small resistor (\( R_s \)) is added in series with the inductor. The resistor is used to measure the charging current or the load current. The resistance is only 0.1 \( \Omega \) to avoid large voltage drop. The detected voltage is amplified by a gain designed in the controller. The measured current is used for current control in the charging mode. This design makes the single-stage circuit simultaneously rectify AC power to DC and charge a battery with current control.

#### C. Operating Principle of the SIDO

The ordered power-distributive control (OPDC) method is applied to this circuit. The OPDC is a type of non-time-multiplexing (non-TM) control scheme [23] - [24]. Its switching time is less than the time-multiplexing (TM) control [25] - [28], so its efficiency is better. The power stage of the SIDO is shown in Fig. 4a, and the inductor current waveform shown in Fig. 4b is used to explain the inductor current behavior of the power stage. At first, the inductor is charged at phase A. The slope of inductor current is the difference of the input and the output voltages of phase A (\( V_{out,A} \)) divided by inductor value. Its slope is marked in Fig. 4b as well. At that moment, MP7 and MP8 are turned on. When the energy is charged high enough for phase A, MP8 is turned off, and MP9 is then turned on. Charging would continue until the total energy for both outputs A and B is enough. Then, circuit alternates switches to discharge the inductor energy to the load. At that moment, MP7 is turned off, and MN3 is turned on. The circuit repeats the sequence of the aforementioned action in every cycle.

![Fig. 3](image-url) An equivalent circuit for the proposed structure showing the current flow in positive and negative periods.

![Fig. 4](image-url) The power stage of the SIDO.

- **A. AC-DC Converter Power Stage Model**
- **B. Inductor current waveform**

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**Fig. 4** The power stage of the SIDO.

- **Phase A**
- **Phase B**
- **AC V\(_{in}\)**
- **V\(_x\)**
- **I\(_L\)**
- **V\(_{out}\)**

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**Appendix A**

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**III. Stability Analysis**

- **A. AC-DC Converter Power Stage Model**
- **B. Inductor current waveform**

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**Appendix A**

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**III. Stability Analysis**

- **A. AC-DC Converter Power Stage Model**
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**Appendix A**

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values of the circuit are chosen as follows. The root-mean-square value of the current AC voltage, $V_{rms}$ = 5 V; the output voltage, $V_{out}$ = 3~4.2 V; the output current, $I_{load}$ = 1~200 mA; the switching frequency, $f_{sw}$ = 1 MHz; the input AC source frequency $f_{s}$ = 100 kHz; $C$ = 40 μF; $R_{str}$ = 125 mΩ; $L$ = 8.2 μH; $R_{s}$ = 100 mΩ.

$$G_{\text{id}}(s) = \frac{V_{\text{in,rms}}}{I_{\text{load}}} = \frac{1 + R_{\text{ESR}}C}{L\cdot C\cdot s^2 + \frac{1}{Q_{s0}^2} + 1}$$ (1)

$$\omega_0 = \frac{1}{\sqrt{LC}}$$ (2)

$$Q = \frac{1}{\omega_0} \cdot \frac{(R_1 + R_2) + R_{\text{load}}}{L + C \cdot R_1 (R_1 + R_2) + C \cdot R_2 R_{\text{load}} + C (R_1 + R_2) R_{\text{load}}}$$ (3)

The Bode diagram in Fig. 5 is the control-to-output gain under light load. The solid line is the loop gain without compensation, and the dashed line is the loop gain with compensation. There are two conjugate poles at 8 kHz. These two poles cause the phase to drop rapidly at that point. Although there is a zero in the system, its location is too far from the poles to have any influence. Generally, an integrator is always needed to eliminate the steady-state error because the DC gain is limited to have any influence. Generally, an integrator is always needed to eliminate the steady-state error because the DC gain is limited to have any influence. Generally, an integrator is always needed to eliminate the steady-state error because the DC gain is limited.

Regarding the battery charging function, the amount of current flowing into the battery is a major concern. In this case, the inductor current is measured and its average value is extracted as an indicator of current flow into the battery. The stability of the current control loop is related to the control-to-inductor-current transfer function. Since this single-stage structure is approximated to a step-down switching converter, the model of control-to-inductor-current can be expressed as of

\[
G_{\text{dc,step}}(s) = \frac{1}{R_1 C_1} \cdot \frac{s + \frac{1}{R_2 C_2}}{s(s + \frac{1}{R_2 C_2})}
\] (6)

It is composed of an operational amplifier, two resistors, $R_1$ and $R_2$, and two capacitors, $C_1$ and $C_2$. It provides one zero, one pole, and an integrator. The zero can be placed on the point to cancel the effect of one of these poles and boost the phase. To improve the DC gain, an integrator is included to provide a DC gain as high as the op-amp gain in the compensator. The

![Fig. 5 The Bode diagram of the control-to-output gain for voltage loop with/without compensation.](image)

![Fig. 6 The Bode diagram of the control-to-inductor-current gain for current loop with/without LPF before compensation.](image)
purpose of the high DC gain is to reduce the static error between the output voltage and the reference voltage. Increasing the switching frequency reduces the ripple but it also leads to high switching losses. To make the trade-off, the switching frequency of the proposed circuit is selected at 1 MHz. The AC input should be at least $10^4$ lower than the switching frequency to allow enough PWM signals in one AC period to regulate the AC voltage into DC voltage. To make a better wireless power transfer, the frequency of the AC source at the primary side should be carefully considered to match the coil design. To reduce the switching noise, the cross-over frequency is set at 18 kHz, which is less than one tenth of the switching frequency. The parameters for the compensator are as follows: $R_1 = 300 k\Omega; R_2 = 150 k\Omega; C_1 = 5 pF; C_2 = 1 nF$. In Fig. 5, the DC gain is increased from 8 dB to more than 40 dB at 10 Hz. The loop is stable with 50 degree phase margin when the cross-over frequency is 12 kHz.

C. The Current Control Loop Compensation for AC-DC converter

The same design process for the compensator is applied to the current control loop. The parameter in this type-II compensator are as follows: $R_1 = 90 k\Omega; R_2 = 100 k\Omega; C_1 = 10 pF; C_2 = 20 nF$. The compensated loop gains under light and heavy load conditions are shown in Fig. 7. The bandwidth is 5 kHz which is high enough for controlling charging current. The DC gain is largely boosted compared to the uncompensated loop gain. Therefore, charging current can be well controlled, and the loop is stable.

D. Line-to-output Analysis for AC-DC converter

In the proposed circuit, the AC side voltage is directly regulated to the required DC voltage level at output. Therefore, the line-to-output gain is a major concern in the design of the proposed converter. Based on the previous assumption, the open-loop line-to-output transfer function can be simplified as (7). It is very similar to the line-to-output transfer function in the uncompensated loop gain. Therefore, charging current can be well controlled, and the loop is stable.

E. Compensation for SIDO

To make stability analysis of the SIDO converter, a general model shown in Fig. 9 is analyzed. The transfer functions, $G_{dab}$, $G_{dib}$, $G_{dib}$, and $G_{ddc}$, are well derived in [24] and [35]. The loop gain can be derived as in equations (9) to (11).

$$T_2(s) = FM_a \cdot FM_b \cdot G_{ca}(s) \cdot G_{dab}(s) \cdot G_{dbb}(s) \cdot G_{das}(s)$$  \hspace{1cm} (9)

$$T_1(s) = FM_a \cdot G_{ca}(s) \cdot G_{dab}(s) - \frac{T_2(s)}{1+FM_b \cdot G_{cb}(s) \cdot G_{dbb}(s)}$$  \hspace{1cm} (10)

In Fig. 7, the DC gain is increased from 8 dB to more than 40 dB at 10 Hz. The loop is stable with 50 degree phase margin when the cross-over frequency is 12 kHz.
According to equations (10) and (11), the Bode diagram of these two loops is shown in Fig. 10. $T_1$ is the control loop for energy distribution, and $T_2$ is the control loop for the total energy income. Two type-III compensators are employed for $G_{ca}$ and $G_{cb}$ in each loop. Note that $T_1$ and $T_2$ affect each other if the compensator in one of the loops is adjusted. After careful tuning, the DC gain is improved to reduce the steady-state error after compensation, and the phase margin is enough for both loops.

$$T_2(s) = \frac{FM_b \cdot G_{cb}(s) \cdot G_{dbb}(s)}{1 + FM_a \cdot G_{ca}(s) \cdot G_{dab}(s)}$$

(11)

IV. CIRCUIT DESIGN AND IMPLEMENTATION

The proposed circuits are designed in the laboratory and then taped-out for testing. All the power MOS transistors and the controller are integrated on a single chip. The chip is manufactured by TSMC 0.25 µm high-voltage CMOS process. The circuit schematic of the chip is shown in Fig. 11, and the implementation details of the controller circuit are given in this section.

A. Implementation Concerns

There are several concerns on the circuit implementation. From the battery charging point of view, the fully charged voltage of a lithium-ion battery is 4.2 V. The voltage level of the power stage should be higher than 4.2 V, which means the duty ratio should have some additional margin to make sure that the voltage can reach 4.2 V when additional current is drained by other loads concurrently. It should be noted that all the components, i.e., all switches and L are required to be rated for peak power because the converter directly faces 100 kHz ac power. Therefore, the high-voltage manufacturing process is chosen for the chip implementation. In the high-voltage process, the power MOSFETs for AC-DC converter and their gate drivers are well suited to tolerate high voltage levels up to 12V for this particular process. The 5V devices are selected to construct the MOSFETs for the SIDO. To minimize power consumption, the controller is implemented using low-voltage devices whose supply voltage is 2.5V.
are the compensated signals from the input error. $V_{c_v}$ is used in inductor current, $I_L$. This signal is then compared with the reference voltage, $V_{ref}$, to produce the constant voltage mode, and $V_{c_c}$ is used in constant current load-current sensing circuit with a gain of 200 amplifies the reference signal, $I_{ref}$, in the compensation network to produce the control signal, $V_{c_c}$, for current control loop. $V_{c_v}$ and $V_{c_c}$ are the key components of the proposed circuit is the switching controller, which is shown in Fig. 13. The function of this block is to create two pulse-width modulated signals, duty1 and duty2, based on the input signal, duty, and to synchronize them with the AC power source. These two signals control MP1 and MP2 following the behavior shown in Fig. 3. The switch controller compares three signals; (1) the input AC voltage, $V_{in}$, (2) the feedback voltage, $V_{FB}$, and (3) the pulse-width modulation signal, duty. The signals, $V_{AC_P}$ and $V_{AC_N}$, are two nodes of the input AC voltage, $AC V_{in}$. When the AC voltage is positive and is higher than the feedback voltage, $V_{FB}$, duty1 follows duty and is passed to MP1. The signal duty2 is the complement of duty1 at that moment. In this process, MP1 switches at a high duty and duty_bar. Non-overlapping design is required to avoid a short circuit between the high side PMOS transistors, MP1 and MP2.

D. Zero Current Protection (ZCP)

When the input AC voltage is close to zero, the inductor current can go negative because current can flow back from output side through inductor to ground or to common node. The output capacitor would be discharged and induce the voltage drop on the output. To avoid this condition, the current path in Fig. 3(b) and Fig. 3(d) should be cut. Therefore, a zero current protection is needed to turn off the MOSFETs, MP1 and MP2 when inductor current approaches zero. To achieve this goal, the signal, duty_bar, is restrained by the ZCP circuit. When the inductor current, $I_{c}$, is larger than 0, duty_bar passes through to the switching controller. On the other hand, duty_bar stays low when $I_{c}$ is lower than 0.

E. Switching controller

A key component of the proposed circuit is the switching controller, which is shown in Fig. 12. The control behavior is shown in Fig. 12. The body cannot be connected to the source of the PMOS transistor because the source in this case is connected to AC power which changes periodically between positive and negative 7 V. A similar situation occurs for the drain terminal, so the PMOS transistor in step 2 in Fig. 12 may not turn-off properly and the current can leak through the source and the substrate to the output. Therefore, the dynamic body biasing technique is also applied to MP1 and MP2 which are shown in Fig. 2 and Fig. 11. The biasing circuit is constructed by four PMOS transistors, which are MP3, MP4, MP5, and MP6 in Fig. 11. These PMOS transistors turn on in an alternating fashion. For example, when the source of MP1 is at the highest positive voltage and the source of MP2 is at the lowest negative voltage, MP4 and MP5 turn on to provide the body voltage for the MP1 and MP2. For MP1, its body is connected to the highest positive voltage; the body of MP2 is connected to the inductor which has higher voltage potential than the lowest negative voltage. This way, the leakage current can be eliminated by keeping the higher voltage on the body of the PMOS transistor and vice versa.

C. The Control Circuit for AC-DC Converter

The block diagram of the proposed controller is shown in Fig. 11. The components connected to the op-amps, Op1 to Op4, form a type-II compensation network. Feedback should be applied to the input of the compensator ($V_{out}$ in Fig. 11) to control and stabilize the output voltage. The feedback voltage is compared with the reference voltage, $V_{ref}$, to produce the control signal, $V_{c_v}$, for voltage control loop. By the time, a load-current sensing circuit with a gain of 200 amplifies the voltage on node $I_{sense}$ and node $I_{sense}$ to obtain the value of inductor current, $I_{c}$. This signal is then compared with a current reference signal, $I_{ref}$, in the compensation network to produce the control signal, $V_{c_c}$, for current control loop. $V_{c_v}$ and $V_{c_c}$ are the compensated signals from the input error. $V_{c_v}$ is used in constant voltage mode, and $V_{c_c}$ is used in constant current mode.

B. Dynamic Body Biasing

The control of dynamic body biasing is prone to latch-up and substrate problems on the high-side switches. The circuit behavior is shown in Fig. 12. The body cannot be connected to the source of the PMOS transistor because the source in this case is connected to AC power which changes periodically between positive and negative 7 V. A similar situation occurs for the drain terminal, so the PMOS transistor in step 2 in Fig. 12 may not turn-off properly and the current can leak through the source and the substrate to the output. Therefore, the dynamic body biasing technique is also applied to MP1 and MP2 which are shown in Fig. 2 and Fig. 11. The biasing circuit is constructed by four PMOS transistors, which are MP3, MP4, MP5, and MP6 in Fig. 11. These PMOS transistors turn on in an alternating fashion. For example, when the source of MP1 is at the highest positive voltage and the source of MP2 is at the lowest negative voltage, MP4 and MP5 turn on to provide the body voltage for the MP1 and MP2. For MP1, its body is connected to the highest positive voltage; the body of MP2 is connected to the inductor which has higher voltage potential than the lowest negative voltage. This way, the leakage current can be eliminated by keeping the higher voltage on the body of the PMOS transistor and vice versa.

The two aforementioned control signals, $V_{c_v}$ and $V_{c_c}$, would be selected by an analog MUX to pass as $V_c$. The selection is based on the measured output voltage, $V_{out}$. When $V_{out}$ reaches the upper threshold, which is set at 4.2 V in this case, analog MUX passes $V_{c_v}$ to $V_c$. On the other hand, analog MUX passes through $V_{c_c}$ to $V_c$ when $V_{out}$ reaches the lower threshold, which is 3V in this design. The voltage control signal, $V_c$, is then compared with a sawtooth signal, $V_{ramp}$, to determine the pulse width of the PWM signal. This signal, which is the main signal to control the power MOSFETs, passes through the non-overlap controller to generate two non-overlapping signals, duty and duty_bar. Non-overlapping design is required to avoid a short circuit between the high side PMOS transistors, MP1 and MP2.
frequency to regulate the passing power. MP2 stays complement to MP1 to keep continuous current for inductor in the positive period. During negative period, similar to the positive period (but in opposite polarity), the duty2 follows duty while keeping duty1 signal the complement of duty2. These two signals, duty1 and duty2, are then inverted because MP1 and MP2 are p-type MOSFETs.

F. SIDO circuit

The SIDO circuit is shown on the right side of Fig. 11. The voltage of output port B, $V_{out_b}$, is the feedback to the compensator to generate the control signal $V_{cb}$. This signal is then compared with a saw-tooth wave to determine the PWM signal for controlling the total energy of two phases. Likewise, port A output voltage, $V_{out_a}$, is the feedback to the compensator to generate control signal, $V_{ca}$ for PWM signal $D_a$. The PWM signals, $D_a$ and $D_b$, are used to distribute energy to both phases.

G. Circuit Implementation

The micro-photograph of the controller for the proposed circuit is shown in Fig. 14. The chip size is 1.9 mm $\times$ 2.3 mm including pads. The chip area includes an AC-DC converter for the size of 1.1 mm $\times$ 0.794 mm, and a SIDO for the size of the 0.739 mm $\times$ 0.692 mm. This chip can be integrated into a 3 cm $\times$ 3 cm medical module containing passive (capacitor and inductor) and network components.

To differentiate controllers from the components in the power stages, MOSFETs and controllers are labelled in accordance with those shown in Fig. 11. The controller includes saw-tooth generators, bias circuits, operational amplifiers, comparators, current protection, and logic elements. Those circuits are well deployed in different areas because their voltage specifications are different.

H. Experimental Results

The AC-DC converter is expected to work either with or without a battery. Such design can extend the operating reliability once battery power is low or in need of replacement. Two tests were conducted to evaluate the performance of the designed AC-DC converter. When the AC-DC converter is not connected to a battery, it works under constant-voltage mode (CV mode). In CV mode, load transient is measured and the results are shown in Fig. 15. In the beginning, the 2.1V DC output voltage is regulated from 12 $V_{pp}$ (peak to peak) AC induced voltage to serve to a 90 mA load. The 100 kHz AC power is generated by a Class-E Amplifier and then transmitted through a transformer. The reason to use a transformer is to eliminate the uncertainties from the coil in this case. When load current steps down from 90 mA to 1 mA within 1 $\mu$s, the output voltage shows 260 mV voltage overshoot and then recovers within 1.2 ms. When load current steps up from 1 mA to 90 mA within 1 $\mu$s, the output voltage has 220 mV undershoot and then recovers within 1.2 ms. The test result validates the regulation capability of the AC-DC converter in response to load changing.

In Fig. 16, the waveform shows that the 14 $V_{pp}$ AC source is converted to a 3.6 V DC voltage. The DC voltage is then connected to SIDO converter for further voltage regulation. The voltages of the two output ports are 1.2 V and 1.5 V. Note that the ripple at the output of the SIDO converter is further reduced to 50 mV that could serve as a standalone power supply without
battery support. Another load transient is tested at the output ports of the SIDO converter. Testing result is shown in Fig. 17. When one of the output current steps up or down rapidly between 0 mA and 30 mA, the output voltage, $V_{out_a}$, shows that the voltage spikes are less than 180 mV and the transient recovers within 190 μs. Although there is minor cross regulation between phase A and phase B, the figure shows that the two outputs of the SIDO are stable.

To evaluate the battery charging operation, the output of the AC-DC converter is connected to both SIDO and the battery in parallel for the following tests. The controller detects battery voltage to determine the control loop. The behavior can be explained by the internal signal in the simulation result in Fig. 18. A large capacitor is used to model the battery to shorten the simulation time. In the beginning, the circuit is started from constant-current mode to charge the battery. The current mode compensator is selected to generate control signal. The control signal is compared with saw-tooth wave, $V_{ramp}$, to generate PWM signal. During the charging period, the battery voltage increases at a constant slope until the battery voltage reaches the upper threshold. Afterwards, charging current drops. At that moment, the voltage compensator is selected instead to enable constant-voltage mode operation.

![Fig. 17](image1.png)
Fig. 17 The load transient test for both output of SIDO converter.

![Fig. 18](image2.png)
Fig. 18 The internal signal when charging a battery in the simulation.

![Fig. 19](image3.png)
Fig. 19 The current and voltage in battery charging test with the proposed AC-DC converter.

![Fig. 20](image4.png)
Fig. 20 The efficiency and temperature in battery charging test with the proposed AC-DC converter.
The experimental result of battery charging is shown in Fig. 19. A 110mAh lithium-ion battery is used in the experiments. When the charging current is 70 mA, it takes about 95 minutes to charge the battery from 3 V to 4.2 V. When the charging current is 100 mA, it takes about 61 minutes to charge the same battery from 3 V to 4.2 V. The charging current is kept constant until charging is complete. The voltage does not increase with a constant slope due to the variation of the equivalent resistance and inductance during charging with a real battery. When battery charging is complete, the voltage becomes constant and current drops to 0.

For biomedical applications, the temperature is also a big concern. For the temperature measurement, the battery is sealed with an electrical thermometer by an insulation sleeving. Through the charging progress, records of efficiency and temperature are shown in Fig. 20. The efficiency is around 70% under 70 mA charging current during the charging period, and the temperature of the battery only rises by 1°C. For the second test, the efficiency is around 75% under 100 mA charging current during the charging period with the temperature rise less than 1°C. The efficiency of the proposed circuit is also measured in the load current range between a light load of 6 mA, to a heavy load of 250 mA. The results show that the maximum efficiency is 79.4% under 150 mA load current. The experiments are repeated 10 times to average a value for each dot in Fig. 19 and Fig. 20.

In TABLE I, circuit functionality and performance indices of the proposed circuit are compared with those of state-of-the-art AC-DC converters. It is noted that the proposed circuit provides dual functions by the delicate control for AC-DC power conversion, which includes battery charging control as well as constant voltage control.

Regarding the power efficiency issue on battery charging path, most counterparts need a pre-stage rectifier to provide DC input voltage. The optimal efficiency of the rectifier is 89% [8] [36]. In [16], the authors adopt a step-down DC-DC converter following the rectifier to regulate charger voltage, which needs to count another 86% efficiency in the overall power path. Therefore, this two-stage topology has a combined efficiency of 76%.

Both [10] and [12] use linear regulators to enhance the battery charging capability, which do not include the pre-stage power conversion circuit (rectifier+Buck). To make a proper estimation of the overall power path, 76% pre-stage efficiency is taken into account.

Table I shows that the performance of the proposed circuit is comparable to those state-of-arts in terms of efficiency, functionality, and size. Structure-wise speaking, the physical circuit proposed in this paper is simpler and smaller than that of the conventional cascading system. This structure is pretty appealing to the application of mobile electronic devices because mobile devices usually require smaller size and lighter weight. Note that this circuit is combined with inductive power transfer (IPT) circuit to achieve resonant condition and make the power transfer. There are some cases discussing about the impedance matching between receiver and active rectifier to reach high efficiency [11]. However, the optimization of the IPT is not within the scope of this paper. This issue could be a future work for the next phase research.

<table>
<thead>
<tr>
<th>Year</th>
<th>Topology</th>
<th>AC frequency (MHz)</th>
<th>Secondary $V_{pp}$ (V)</th>
<th>$V_{rectifier}$ (V)</th>
<th>$V_{out}$ (V)</th>
<th>MAX. I$_{load}$ (mA)</th>
<th>Max. Output power (W)</th>
<th>This Work</th>
</tr>
</thead>
<tbody>
<tr>
<td>2011</td>
<td>linear-type battery charger</td>
<td>27.3</td>
<td>20</td>
<td>9</td>
<td>3.3, 5</td>
<td>40 mA</td>
<td>0.36 W</td>
<td>100 kHz</td>
</tr>
<tr>
<td>2011</td>
<td>DC-DC Buck or Buck-type</td>
<td>--</td>
<td>--</td>
<td>4.3</td>
<td>4.2</td>
<td>3 mA</td>
<td>0.012 W</td>
<td>--</td>
</tr>
<tr>
<td>2012</td>
<td>--</td>
<td>5-10</td>
<td>2-2.2</td>
<td>2.1-4.2</td>
<td>2.1-4.2</td>
<td>900 mA</td>
<td>4 W</td>
<td>--</td>
</tr>
<tr>
<td>2016</td>
<td>--</td>
<td>2-2.2</td>
<td>4.2</td>
<td>0.7-1.2</td>
<td>2-2.2</td>
<td>700 mA</td>
<td>0.84 W</td>
<td>--</td>
</tr>
<tr>
<td>2016</td>
<td>--</td>
<td>--</td>
<td>--</td>
<td>--</td>
<td>2-2.2</td>
<td>760 mA</td>
<td>--</td>
<td>--</td>
</tr>
</tbody>
</table>

* The total efficiency is measured or counted from the secondary coil (Receiving) to the battery end.
stage AC-DC converter is that it can serve loads either without a battery by CC-CV dual modes, which can be detected automatically by the output condition. The SIDO structure is integrated with AC-DC converter as well to provide additional voltage levels for the loads. All the experimental results show that the proposed power module can perform high-efficient and robust voltage regulation for medical circuits or battery charger in the wearable devices.

REFERENCES


