Phase-Locked Loop Using Complex-Coefficient Filters for Grid-Connected Inverter

Akihiro Ohori Member (DAIHEN Corporation, a-ohori@daihen.co.jp)
Nobuyuki Hattori Member (DAIHEN Corporation)
Tsuyoshi Funaki Member (Osaka University)

Keywords: photovoltaic power system, grid-connected inverter, phase-locked-loop, complex coefficient filter

Recently, photovoltaic (PV) power systems have attracted considerable attention in an attempt to mitigate global warming. In a PV power system, it is necessary to synchronize the grid voltage when a PV inverter is interconnected with a grid. Conventionally, sophisticated phase-locked loops (PLL) were proposed for a three-phase system. However, it was difficult to establish the high-performance PLL for a single-phase system, because it is difficult to obtain instantaneous orthogonal vectors of a grid voltage. This paper proposes a high-speed and high-precision PLL using complex-coefficient filters for a single-phase grid-connected inverter. The instantaneous orthogonal vectors of grid voltage for a single-phase system can be calculated easily by using the complex-coefficient filters.

The difference between a real-coefficient filter and a complex-coefficient filter is shown in Fig. 1. Fig. 1(a) shows a bode diagram for a real-coefficient band-pass filter, and Fig. 1(b) shows a bode diagram for a complex-coefficient band-pass filter. The real-coefficient filter has a symmetric characteristic at both negative and positive frequency values, but the complex-coefficient filter has different characteristics at positive and negative frequency values. The complex-coefficient band-pass filter can extract the instantaneous orthogonal vectors (2) from the grid voltage in a single-phase system (1).

\[ V = \sum_{n=1}^{\infty} V_n \cos(2\pi nf_ot + \theta_o) \]

\[ V_r + jV_i = \frac{V_1}{2} \cos(2\pi f_ot + \theta_o) + \frac{jV_1}{2} \sin(2\pi f_ot + \theta_o) \]

Fig. 2 shows the proposed single-phase PLL. It has a simple configuration, where only the complex-coefficient filters are inserted in front of a standard PLL. The numerical result is shown in Fig. 3. Fig. 3(a) shows the grid voltage waveform, Fig. 3(b) shows the instantaneous orthogonal vectors calculated using the complex-coefficient filter, and Fig. 3(c) shows the detected phase error of the proposed PLL. These show the effectiveness of the proposed method.

![Fig. 1. Bode diagram of band-pass filter](image1)

![Fig. 2. Proposed single-phase PLL](image2)

![Fig. 3. Simulation results](image3)
A Phase-Shift-Controlled Switched-Capacitor-Based Resonant Converter
— — Interleaving Technique for Reducing the Switching Voltage/Current Ripples— —

Takuro Arai Student Member (Tokyo Institute of Technology)
Hideaki Fujita Member (Tokyo Institute of Technology)

Keywords: conversion efficiency, interleaving technique, phase-shift control, power losses, resonant converters, switched capacitors

This paper deals with interleaving technique for switched-capacitor-based resonant converters (SCRCs) with a phase-shift control method. The current ripples produced by two SCRCs cancel each other in case of the interleaving operation, and thus the output current ripple can be significantly reduced. This paper also presents that a new connection of filter capacitors makes it possible to suppress the output voltage ripples, leading to a great reduction in capacitance value of the output filter capacitor. Moreover, the theoretical analysis clarifies that current sharing between two SCRCs strongly depends on the stray resistance in conventional control and that the phase-shift control can equally share the input and output currents irrespective of the stray resistance. A 2.6-kW experimental setup has confirmed that the interleaving operation can reduce the current ripples to 44% and the voltage ripples to 5% as compared with those in conventional synchronous operation.

Fig. 1 shows two SCRCs connected in parallel, which consist of the same switching devices and resonant circuits. The output filter capacitor $C_{out}$ is connected in series to the input filter capacitor $C_{in}$. SCRC-A and SCRC-B are operated at the same switching frequency, and the operating phase angle of SCRC-B lags by $180^\circ = \frac{T_s}{2}$ from that of SCRC-A. Therefore, output odd-order ripple currents included in $i_{1A}$ and $i_{1B}$ cancel each other if the amounts of the ripple currents in $i_{1A}$ and $i_{1B}$ are the same. The remaining even-order ripple currents cause the ripple voltages not only in $v_{C_{out}}$ but also in $v_{C_{in}}$. The voltage ripples in $v_{C_{out}}$ and $v_{C_{in}}$ also cancel if the phase angles of the voltage ripples are opposite.

Theoretical analysis developed in this paper derives the amplitude of each frequency component in the ripple current in both conventional and phase-shift control methods. Moreover, the analysis also clarifies that the current sharing between SCRC-A and SCRC-B depends on the circuit resistance in the conventional control method, whereas it is decided by inductance of the resonant inductors $L_r$ in the phase-shift control method. It is impossible to adjust the circuit resistance for balancing the current sharing, because it is the sum of the on-state resistance in $S_1$-$S_4$, the equivalent series resistance in $C_r$, and so on. On the other hand, the resonant inductors are easily adjusted by the air gap in the magnetic cores.

Fig. 2 shows the relationship between the averaged output current $I_{out}$ and the rms ripple current $\tilde{i}_1$. The phase-shift control method has a reduced current ripple, especially at a lower load range.

Fig. 3 shows experimental waveforms at a rated power of 2.6 kW in the case of the interleaving operation. The voltage ripples in $v_{C_{out}}$ and $v_{C_{in}}$ are sawtooth waveforms having different polarities. Thus, a small ripple only remains, and the rms voltage ripples in the $v_{out}$ are reduced to 0.1 V.
Multifunctional Control of a Two-Wheel Driven Wheelchair Considering Comfort of a Passenger

Takuro Kawamura  Student Member (Keio University, takuro@sum.sd.keio.ac.jp)
Toshiyuki Murakami  Senior Member (Keio University, mura@sd.keio.ac.jp)

Keywords: two-wheel driven wheelchair, uprising and landing control, trajectory tracking control

This paper describes a control scheme for the multifunctional control of a two-wheel driven wheelchair. Fig. 1 shows the overview of the experimental system in this research.

Within forty years, the number of people around the world aged 60 or over is going to be triple of the number in 2007. It has been reported that the aging ratio in Japan is now over 20% and is still increasing. Therefore, having a senior-friendly society is necessary. Because of such social circumstances, various welfare devices have been developed for elderly and handicapped people. The wheelchair is a highly-popularized welfare device. However, typical wheelchairs have two problems. One is low mobility, and the other is the difficulty faced while going over bumps. These problems are caused by the front casters.

Recently, a two-wheel driven wheelchair with no casters has been proposed. Although the above issues are solved by using such vehicle structures, a serious and important issue pertaining to stability arises. To resolve this issue, some effective approaches have been proposed. Nevertheless, further research is required in order to put two-wheel driven wheelchairs to practical use. The aim of this research is to achieve the multifunctional control of the two-wheel driven wheelchair. In particular, this paper focuses on the process of getting on and off and trajectory tracking control.

Fig. 2 shows the entire block diagram of the proposed control system. First, the process of getting on and off is focused on. In conventional method, a passenger gets on and off while the wheelchair stabilized. Uprising and landing control are proposed to achieve the getting on and off motions efficiently. The proposed approach allows the wheelchair to uprise and land while the passenger gets on. In addition, a variable command generation is proposed to decrease vibration. The experimental results indicate that comfort of a passenger is improved by the proposed method.

Next, trajectory tracking control is focused on. Trajectory tracking in the conventional method is not sufficiently robust. In this research, a method for achieving robust trajectory tracking control is also proposed. For a desired target trajectory, the two-wheel driven wheelchair can track the trajectory correctly by using the proposed method. The experimental results indicate that robust trajectory tracking control is achieved without decreasing a passenger’s comfort.

Through the application of the proposed approaches, the multifunctional control of a two-wheel driven wheelchair is achieved, and comfort of a passenger is improved. The results obtained in this study show that two-wheel driven wheelchairs are highly feasible for practical use in future society.

Tsuyoshi Suzuki Non-member (Tokyo Denki University, tskz@ieee.org)
Takafumi Kobayashi Non-member (Tokyo Denki University)
Kei Sawai Non-member (Tokyo Denki University)
Kuniaki Kawabata Member (RIKEN)
Fumiaki Takemura Member (Okinawa National College of Technology)
Naoko Isomura Non-member (Okinawa National College of Technology)
Hideyuki Yamashiro Non-member (Okinawa National College of Technology)

Keywords: wireless sensor network, sensor node, underwater environment, monitoring, information gathering support

Coral reefs are a very important environment that maintains a great variety of life-forms. In recent years, there have been growing concerns about the disappearance of coral reefs because of the impact of environmental change. Therefore, there is an urgent need to preserve coral reefs. In the present situation, in order to preserve coral reefs, experts such as divers and researchers have been collecting data periodically on environmental conditions. However, these monitoring tasks are hampered by critical problems such as diving time limitations, decompression sickness, and accidents due to the hazardous environment.

Therefore, we are developing an underwater monitoring system applying wireless sensor network technologies to support the monitoring tasks aimed at preserving the coral reef environment (Fig. 1). Here, we study the functions of the wireless sensors that make up the underwater monitoring wireless sensor network for gathering information on underwater environmental conditions. For this purpose, a prototype sensor node was developed based on our previous experiments and the feedback of coral reef researchers. Figure 2 shows the layout of the prototype sensor node.

This sensor node consists of two units: a wireless communication unit and a sensor unit. The wireless communication unit is equipped with an airtight container with a float (weight capacity 69 [kg]) for waterproofing and a laptop PC with a wireless LAN adaptor and an extension antenna. IEEE802.11b is used for the communication protocol. A GPS sensor and two lithium-ion rechargeable batteries are also included. The sensor unit is equipped with a USB camera for observation of the coral, a USB temperature sensor for underwater temperature measurement, and an acceleration sensor for detection of the sensor unit posture. A motor controller is also included to drive the underwater thruster. These devices are contained in a watertight acrylic hollow cylinder. Three underwater thrusters are mounted on the outer frame of the sensor unit to control the posture angle on a horizontal plane.

We conducted practical field experiments to observe a colony of coral reefs along the Okinawa coast by using the prototype. The results are described here (Fig. 3). Then, we verified that the captured images were useful for observation of the coral from feedback given by researchers on coral ecology and environmental engineering and underwater monitoring expert divers. Finally, we discuss the experimental results and the required functions of the sensor node for our future research.
Reference Trajectory Design That Overlays Motion Modes of Laser-positioning Servomechanism of Galvanometric Mirrors

Daisuke Matsuka Member (Hitachi, Ltd.)
Haruaki Otsuki Non-member (Hitachi, Ltd.)
Tetsuya Sano Non-member (Hitachi Via Mechanics, Ltd.)

Keywords: high speed and high precision control, galvanometric mirror, trepanning

A laser-positioning servo mechanism with galvanometric mirrors is required for high speed and high precision for high-density and multi-layer printed wiring board (PWB) drilling. This paper proposes an efficient control method for galvanometric mirrors that move a laser beam to drill a sequence of holes larger than the beam size. The method is applicable to a microprocessor-based controller, and it reduces drilling time. It combines point-to-point (PTP) motion between holes with continuous path (CP) motion at each hole, producing the reference trajectory in which the circular CP orbit is superimposed onto the PTP trajectory immediately after the movement between the holes starts.

Fig. 1 shows a structural diagram of the positioning system, where \((x_o, y_o)\) is the center of the hole, \((x_L, y_L)\) is the PTP trajectory, \((x_R, y_R)\) is the circular CP orbit, and \((x, y)\) is the trajectory reference generated by an accumulator. The sampling time of the positioning system for the laser drilling machine is too short to calculate the optimum trajectory on time. Therefore, the proposed method uses the circular orbit data stored in memory and adds them to the PTP trajectory. Then, this system changes the additional point of orbit data in memory depending on the time for movement between the holes; the entry trajectory is tangent to the circumference of the hole, so the proposed method reduces vibration in the tangent response.

The effectiveness of the proposed approach has been verified by experiments using a prototype of the servomechanism of galvanometric mirrors. Fig. 2 shows the beam locus obtained by smoothly changing the mirror position from the position corresponding to the PTP mode to that corresponding to the CP mode. The method thus enables a laser drilling machine to start immediately after reaching the orbit, and it shortens the processing time. In addition, no vibration is observed in the transient response.

In the experiment, the proposed method reduces the process time by more than 10%, without loss of precision. Experimental results confirm the effectiveness of the proposed method.
Miniaturization of a Transformer-Linked Multi-Phase Boost Chopper Circuit Using Boost Ratio Adjustment

Takahiro Kawashima Member (Shimane Institute for Industrial Technology, kawashima@shimane-iit.jp)
Masayoshi Yamamoto Member (Shimane University, yamamoto@ecs.shimane-u.ac.jp)

Keywords: boost chopper circuit, multi-phase converter, miniaturization, coupled inductor, tapped inductor

This paper presents an approach to miniaturize an inductor of a boost chopper circuit. The size of a switching power converter strongly depends on that of a capacitor and an inductor. As an approach to miniaturize the size of these elements, we propose to increase the switching frequency. However, high frequency switching leads to an increase in the switching loss. Moreover, this increase often causes electromagnetic interference problems. A multi-phase converter using a coupled inductor is effective for achieving miniaturization, avoiding an increase in the switching frequency. This paper applies a transformer-linked multi-phase boost chopper circuit as a specific example of the multi-phase converter using a coupled inductor.

It is important to adjust the relationship between the boost ratio and the duty ratio because the effect of miniaturizing the transformer-linked multi-phase boost chopper circuit varies depending on its duty ratio. A adjustment of the relationship between the boost ratio and the duty ratio of the boost chopper circuit is realized by using a tapped inductor. Fig. 1 shows a two-phase transformer-linked multi-phase boost chopper circuit comprising a tapped coupled inductor. The boost ratio $G_v$ of a circuit shown in Fig. 1(a) is expressed as follows:

$$G_{v(a)} = \frac{V_o}{V_i} = \frac{1}{1 - d} \cdot \left(1 + \frac{N_2}{N_1} \cdot d\right) \quad \cdots \cdots \cdots \cdots \cdots \cdot (1)$$

The $G_v$ of a circuit shown in Fig. 1(b) is expressed as follows:

$$G_{v(b)} = \frac{V_o}{V_i} = \frac{1}{1 - d} \cdot \left(1 - \frac{N_2}{N_1} \cdot \frac{N_2}{N_1 + 1} \cdot d\right) \quad \cdots \cdots \cdots \cdots \cdots \cdot (2)$$

The boost ratio of the boost chopper circuit comprising a tapped inductor is adjusted on the basis of the winding turn ratio $N_2/N_1$ of the inductor tap. In the case of a two-phase transformer-linked multi-phase boost chopper circuit, 0.5 is the optimal duty ratio for miniaturization. Therefore, if boost ratio is higher or lower than 2, $N_2/N_1$ is adjusted so that the chopper circuit is operated near the 0.5 duty ratio.

The prototypes of the coupled inductor and the conventional non-coupled inductors are shown in Fig. 2. The total weights of the non-coupled inductors are 1030 g and 1035 g. The weight of the tapped coupled inductor is 512 g. The weight of the tapped coupled inductor is one-fourth that of non-coupled inductors.

The experimental results at minimum load are summarized in Table 1 and those at maximum load are listed in Table 2. The experimental results confirm the reduction of the output smoothing capacitor current ripple.

The above results confirm that the proposed method is very effective in miniaturizing a converter.

---

Table 1. Performance characteristics under minimum load condition ($V_i$: 50 V, $V_o$: 140 V, 300 W output)

<table>
<thead>
<tr>
<th></th>
<th>Non-coupled</th>
<th>Tapped coupled</th>
</tr>
</thead>
<tbody>
<tr>
<td>Inductor current ripple (p-p value)</td>
<td>$i_{icr}: 4.36 A$</td>
<td>$i_{icr, wc}: 6.30 A$</td>
</tr>
<tr>
<td>Inductor current RMS</td>
<td>4.21 A</td>
<td>4.31 A</td>
</tr>
<tr>
<td>Output smoothing capacitor current RMS</td>
<td>1.94 A</td>
<td>0.83 A</td>
</tr>
</tbody>
</table>

Table 2. Performance characteristics under maximum load condition ($V_i$: 40 V, $V_o$: 140 V, 1.5 kW output)

<table>
<thead>
<tr>
<th></th>
<th>Non-coupled</th>
<th>Tapped coupled</th>
</tr>
</thead>
<tbody>
<tr>
<td>Inductor current ripple (p-p value)</td>
<td>$i_{icr}: 3.72 A$</td>
<td>$i_{icr, wc}: 11.2 A$</td>
</tr>
<tr>
<td>Inductor current RMS</td>
<td>21.1 A</td>
<td>23.55 A</td>
</tr>
<tr>
<td>Output smoothing capacitor current RMS</td>
<td>10.1 A</td>
<td>4.69 A</td>
</tr>
</tbody>
</table>
Multivariable Control System Design and Application for Energy Saving of Temperature and Humidity Chamber

Masanobu Obika Non-member (ADAPTEX Co., Ltd., obika@adaptex.co.jp)
Masaki Oyama Non-member (ADAPTEX Co., Ltd., oyama@adaptex.co.jp)
Koki Sakane Non-member (ESPEC Corp., k-sakane@espec.co.jp)
Katsuhiko Watabe Non-member (ESPEC Corp., k-watabe@espec.co.jp)
Tetsuya Shimada Non-member (ESPEC Corp., te-shimada@espec.co.jp)
Toru Yamamoto Senior Member (ESPEC Corp., k-watabe@espec.co.jp)

Keywords: energy-saving, multivariable control, temperature and humidity chamber

This paper reports the development of a multivariable control system for energy saving of a thermo-hygrostat. A thermo-hygrostat is shown in Fig. 1. This device is used to inspect various products and to create the desired atmosphere by controlling the temperature and humidity in the chamber.

The controlled values are those of temperature and humidity, and manipulated values are those of the heater, humidifier and refrigerator. A conventional multi-loop PID controller is known to have sometimes caused energy loss owing to the canceling out of the manipulated values because the system gains of both the heater and the refrigerator are opposite. In this case, the equilibrium point of the manipulated values is underspecified.

It is important to fix the equilibrium point of the manipulated values to a reasonable point. In order to achieve this objective, one of the three manipulated values must be led to the desired point that yields energy saving. In other words, the proposed scheme includes the refrigerator output into state variables in order to control the refrigerator output. Further, by designing the multivariable control system instead of multi-loop controllers, the control performance can be drastically improved, although there are strong interaction properties.

Here, the method to determine the desired refrigerator output is shown in Fig. 2. The desired equilibrium point might differ according to the operation points. Therefore, the desired refrigerator output should be changed according to the operating conditions. However, determining the desired refrigerator outputs in advance is a reckless idea. In this paper, the auto-generator of the desired refrigerator output is also proposed. This generator is able to achieve usual energy saving.

The result of this study shows that 40% reduction in energy consumption was achieved. In addition, the control performance was drastically improved by the proposed scheme, as described in Fig. 3. The improvement in both energy saving and the control performance was clearly achieved.
Separation Measurement of Parasitic Inductance and Stray Capacitance using the TDR Method

Satoshi Hashino Member (Tokyo Metropolitan University)
Toshihisa Shimizu Senior Member (Tokyo Metropolitan University)

Keywords: parasitic inductance, stray capacitance, TDR measurement method

This paper proposes measurement method to separate the parasitic inductance and stray capacitance in the components of a power electronics circuit. This study focuses on time domain reflectometry (TDR). In many cases, components mounted on power electronics circuits have parasitic parameters such as a parasitic inductance in series with a stray capacitance in parallel. It is important to separate these parameters into components of inductance and capacitance, as well as to determine these values in order to consider EMI prevention of particularly high power density mounted converters. This paper describes a method to measure the values of the inductance and capacitance individually by changing the characteristic impedance of a micro strip line (MSL) used in TDR measurement method from 50 Ω to other values. The MSL is called a MSL test fixture in this paper.

Figure 1(a) shows the conventional TDR measurement setup in which the first recovery diode (FRD: FFPF30U60S, Fairchild Semiconductor products) is inserted between 50 Ω MSL test fixtures. The photograph shown on the left indicates a TDR measurement setup when the FRD is situated away from the MSL test fixture’s ground layer. In contrast, the photograph shown on the right indicates a setup when the FRD is attached to the MSL test fixture’s ground layer. Fig. 1(b) shows the TDR measurement waveforms when an incident step signal is inserted from the left-hand side of the MSL test fixture shown in Fig. 1(a). The curve in red shows the reflection waveform when the measurement setup shown in the left of Fig. 1(a) is used. This waveform indicates an inductive reflection. In contrast, the curve in black indicates the reflection waveform when the measurement setup shown in the right of Fig. 1(a) is used. It can be seen from the black curve that an inductive reflection and a capacitive reflection occurs continuously because capacitance $C_p$ is formed in parallel between the FRD and MSL’s ground layer. It is difficult to extract $C_p$ by using the conventional TDR measurement method.

In order to separate the inductance and capacitance from the reflection waveform shown in Fig. 1(b), the characteristic impedance of the MSL test fixture is changed from 50 Ω to other values. Fig. 2(a) shows the TDR measurement setups when the characteristic impedance of the MSL test fixture is reduced from 50 Ω to 18 Ω by changing the width of the transmission line. Fig. 2(b) shows the TDR measurement waveforms when an incident step signal is inserted from the left-hand side of the MSL test fixture shown in Fig. 2(a). The curve in red and the curve in black indicate the reflection waveforms when the measurement setup shown in the left and right of Fig. 1(a) are used, respectively. Approximately 7 ns after insertion of the 0.25 V step signal, the amplitude of the reflection waveform is changed to 0.12 V. Then, the inductive reflection of the FRD is obtained. It can be seen from Fig. 2(b) that the shape of the reflection waveforms of both the curves in red and black are consistent. Fig. 3 shows the TDR measurement setup when the characteristic impedance of the MSL test fixture is increased from 50 Ω to 88 Ω as well as its measurement waveform. Fig. 3(b) shows only capacitive reflection. Therefore, the inductance and capacitance can be separated by using the proposed TDR measurement method.
Method of Torque Ripple Suppression in Current Source Inverter for IPMSM with Concentrated Winding Based on Spatial MMF Distribution

Yuu Kawai Student Member (Nagaoka University of Technology)
Hitoshi Haga Member (Nagaoka University of Technology)
Seiji Kondo Senior Member (Nagaoka University of Technology)

Keywords: torque ripple suppression, concentrated winding, spatial harmonics, current source inverter

1. Introduction

In this paper, a method of torque ripple suppression for an interior permanent magnet synchronous motor (IPMSM) with concentrated winding based on spatial magnetomotive force (MMF) distribution is proposed. The proposed superimposed harmonic method can suppress the 6th torque ripple to a greater extent than the conventional sine-wave method.

Fig. 1 shows a model of armature winding with rotor. This paper discussed MMF distribution about the concentrated winding. The proposed method works on the amount of three-phase current with spatial harmonics. Eqs. (1)–(3) represent the three-phase current, MMF distribution factor and MMF distribution, respective, where \( N \) denotes number of turns. Eq.(4), expresses the phase of \( dq \)-axis on Fig. 1, and Eq. (5) gives the amount of \( dq \)-axis current with spatial harmonics. The \( dq \)-axis current with spatial harmonics is the cause of a torque ripple. Therefore, a torque ripple can be suppressed by increasing this three-phase current. The proposed method realizes proper three-phase current. The current source inverter can directly control three-phase current. Fig. 2 shows the circuit configuration of the motor drive system for the current source inverter.

Fig. 3 shows the experimental results of the superimposed harmonic method and conventional sine-wave method. The superimposed harmonic method achieves better suppression of the 6th torque ripple than the conventional sine-wave method; the 6th torque ripple is suppressed from 30\( \% \) to 5\( \% \). Experimental results confirm the validity of the proposed method.

\[
\begin{align*}
    i_u &= I_m \cos \omega t \\
    i_v &= I_m \cos (\omega t - \frac{2\pi}{3}) \\
    i_w &= I_m \cos (\omega t - \frac{4\pi}{3}) \\
    R_{u,CON} &= \sum_{n=0}^{\infty} k_{2n+1} \cos [(2n+1) \cdot \theta] \\
    R_{v,CON} &= \sum_{n=0}^{\infty} k_{2n+1} \cos [(2n+1) \cdot (\theta - \frac{2\pi}{3})] \\
    R_{w,CON} &= \sum_{n=0}^{\infty} k_{2n+1} \cos [(2n+1) \cdot (\theta - \frac{4\pi}{3})] \\
    F_{CON}(t, \theta) &= \sqrt{\frac{2}{3}} N (i_u \cdot R_{u,CON} + i_v \cdot R_{v,CON} + i_w \cdot R_{w,CON}) \\
    \theta_d &= \omega t - \frac{\pi}{6} \\
    \theta_q &= \omega t + \frac{\pi}{2} \\
    i_{d,CON}(t) &= \frac{F_{CON}(t, \theta_d)}{N} \\
    i_{q,CON}(t) &= \frac{F_{CON}(t, \theta_q)}{N} \\
\end{align*}
\]
Energy Transmission System for an Artificial Heart
—- Novel Method for Reducing Fluctuations in Output Voltage by Uniformization of Magnetic Coupling Coefficient—-

Shumpei Taguchi Non-member (Department of Applied Electronics, Tokyo University of Science, Japan)
Kenji Shiba Member (Department of Applied Electronics, Tokyo University of Science, Japan)

Keywords: artificial heart, energy transmission, output voltage, electromagnetic induction

Transcutaneous energy transmission systems (TETSs) use electromagnetic induction to transfer power to an artificial heart or to an left ventricular assist device (LVAD). An air-core-type TETS achieves high-energy transmission efficiency and can stably supply energy. However, the primary coil and secondary coil must be placed on the skin. To make this method practical, placing the primary coil on the skin should be avoided because such placement would enable the primary coil to be moved by sweat and each type of motion. Therefore, in this study, we developed a novel TETS for an LVAD, where in the primary coil is placed on the clothing rather than on the skin. In this case, a fluctuation in the relative position between the primary and secondary coils causes a large change in the output voltage. To drive an LVAD stably, it is necessary to control the output voltage using feedback control. Feedback control, however, requires the use of an additional information transmission system in the TETS circuits. Therefore, we developed a novel method for reducing the fluctuation in the output voltage without using feedback control; this method involved uniformizing the coupling magnetic coefficient $k$.

The results of our experiments show that the gap in the position of the transcutaneous transformer is maximum when the distance between the coils, $d$, is 3 cm (considering the thickness of the skin) and the distance between the coil axes, $l$, is 4 cm. Therefore, we designed a transcutaneous transformer so that $\Delta k$ has a low value when $d$ is varied from 1 to 3 cm and when $l$ is varied from 0 to 4 cm. Fig. 1 shows the analysis results of $\Delta k$ as function of the outer diameter of the primary coil when the secondary coil diameter is fixed at 7 cm. $\Delta k$ denotes the difference between $k$ in the initial state ($d = 1$ cm, $l = 0$ cm) and $k$ in the maximum-position-gap state ($d = 3$ cm, $l = 4$ cm). As shown in Fig. 1, $\Delta k$ decreases with an increase in the outer diameter of the primary coil. Similarly, $\Delta k$ decreases with a decrease in the outer diameter of the secondary coil.

Photographs of the air-core-type transcutaneous transformer designed in this study are shown in Fig. 2. $T_1$, $T_2$ are primary coils, $R_1$ and $R_2$ are secondary coils, and $T_1$-$R_1$ is a conventional transcutaneous transformer. $T_2$-$R_2$ is a proposed transcutaneous transformer. $T_2$ is designed to have an outer diameter larger than that of $T_1$, $R_2$ is designed to have an outer diameter smaller than that of $R_1$.

The experiment measured the rotational speed of a DC brushless motor when $d$ was varied from 2 to 3 cm and when $l$ was varied from 0 to 4 cm. Fig. 3 shows the rotational speed of the transcutaneous transformer as functions of $d$ and $l$. The rotational speed of a conventional transcutaneous transformer, $T_1$-$R_1$, was 0 rpm when $d$ was 3 cm and $l$ was 4 cm. Thus, the fluctuation in the output voltage was 19.2 V. Conversely, the rotational speed of the proposed transcutaneous transformer, $T_2$-$R_2$, was 1300 rpm when $d$ was 3 cm and $l$ was 4 cm. Thus, the fluctuation in the output voltage was 11.0 V.

These results suggest that the fluctuation in the output voltage can be maintained at a sufficiently low value so as to drive the LVAD stably without the use of feedback control.
A Precise Photolithography Process Control Method using Virtual Metrology

Hidetaka Tsuda Member (Fujitsu Limited, tsuda.hidetaka@jp.fujitsu.com)
Hidehiro Shirai Non-member (Fujitsu Semiconductor IT Systems Limited, shirai.hidehiro@jp.fujitsu.com)
Eiichi Kawamura Non-member (Fujitsu Semiconductor Limited, kawamura.eiichi@jp.fujitsu.com)

Keywords: semiconductor, virtual metrology, tool data, inspection data, data mining

Virtual metrology, which enables both precise controllability and economic efficiency for manufacturing processes, has recently attracted interest. Nowadays, there are many application examples of virtual metrology in CVD (chemical vapor deposition) processes, but only a few examples in photolithography processes. In both cases, the employed virtual metrology models are established by conventional hypothesis verification methods that depend on engineers’ skills. In addition, it is necessary to update the models to reflect situational changes in tools and processes. Therefore, the models have problems in maintaining accuracy. Therefore, establishing or discovering high-accuracy methods for virtual metrology models is necessary for precise process control.

We propose a precise processes control method using virtual metrology in photolithography processes. Figs. 1(a) and 1(b) show the correlations between inspection data \( p_2 \) related to focus control and tool data \( p_1 \) related to the tool environment. Although Fig. 1(a) composed of 20 records does not show a remarkable correlation, Fig. 1(b) composed of 5 records selected by our proposed data mining method shows a remarkable correlation that is buried in the vast data. We adopt Fig. 1(b) itself as the virtual metrology model that holds good regardless of situational changes and make updation unnecessary. Whenever abundant tool data \( p_1 \) is obtained, virtual metrology data \( p_2v \) equivalent to inspection data \( p_2 \) is calculated by the virtual metrology model, and the process recipe is updated to apply process control based on the latest situation.

Fig. 2 shows the trend of the tool management parameter which is a process control result. Before time \( t_s \), it shows the process control result by measured inspection data \( p_2 \). The duration of the inspection data collection is so long that used process recipes do not always reflect the latest situation. As a result, the tool management parameter frequently exceeds the control limits, UCL (upper control limit), LCL (lower control limit). After \( t_s \), on the other hand, the figure shows the simulated process control result by \( p_2v \). Whenever tool data \( p_1 \) is collected, the process recipe is updated and process control based on the latest situation is applied. S. D. (standard deviation) of the tool management parameter has reduced to less than 10% after \( t_s \). This proves the validity of our method.

![Fig. 1. Correlations between \( p_1 \) and \( p_2 \)](image1)

![Fig. 2. Trend of the tool management parameter](image2)
Double-Switch Series-Resonant Cell Voltage Equalizer Using Voltage Multiplier for Series-Connected Energy Storage Cells

Masatoshi Uno Member (Japan Aerospace Exploration Agency, uno.masatoshi@jaxa.jp)
Akio Kukita Member (Japan Aerospace Exploration Agency, kukita.akio@jaxa.jp)

Keywords: voltage equalizer, electric double-layer capacitor (EDLC), series-resonant inverter, voltage multiplier

Series-connected energy storage cells, such as lithium-ion cells and electric double-layer capacitors (EDLCs), suffer from a voltage imbalance that stems from nonuniform individual cell characteristics, such as capacity and the self-discharge rate. The voltage imbalance is very likely to cause premature irreversible deterioration and shorten the service life of cells. Hence, cell voltage equalizers are necessary to eliminate cell voltage imbalance, thereby ensuring years of operation.

Various types of cell voltage equalization techniques have been proposed. However, conventional equalizers require multiple switches, magnetic components, and/or secondary windings of a multi-winding transformer in proportion to the number of series connections of cells. Therefore, conventional equalizers are usually complex, expensive, bulky, and less extendable as the number of series connections increases.

A double-switch series-resonant equalizer using a voltage multiplier is proposed in this paper, as shown in Fig. 1. The double-switch operation without a multi-winding transformer offers simplified circuitry and good modularity at reduced size and cost, compared to conventional equalizers.

Mathematical analyses were separately performed for the voltage multiplier and the series-resonant inverter. The voltage multiplier was redrawn as a dc equivalent circuit, whereas the series-resonant inverter was described using current models. By combining the dc equivalent circuit and the current models, a dc equivalent circuit for the proposed equalizer was derived, as shown in Fig. 2.

The simulation analyses were separately performed for both the original and the derived dc equivalent circuits. The simulation results of the derived dc equivalent circuit were in good agreement with those of the original circuit, even under parameter-mismatched conditions, thus verifying the derived dc equivalent circuit. Given that the dc equivalent circuit does not contain high-frequency switching devices, even simulation analyses of an hour’s duration can be completed in an instant, thereby dramatically reducing the simulation time and burden.

A 5-W prototype of the proposed equalizer was built for eight cells connected in series. An experimental equalization test was performed for series-connected EDLCs from an initially voltage-imbalanced condition. The voltages of cells with high initial voltages decreased because they provided power to the equalizer, whereas the voltages of cells with low initial voltages increased because they received power from the equalizer. Voltage imbalance was gradually eliminated in course of time. The standard deviation of the cell voltages decreased to approximately 5 mV at the end of the experiment, thus demonstrating the equalization performance of the proposed equalizer.

Fig. 1. Double-switch series-resonant equalizer using a voltage multiplier for series-connected energy storage cells

Fig. 2. Derived dc equivalent circuit of the proposed equalizer

Fig. 3. Equalization profiles of eight EDLCs connected in series