Analysis, Design and Control of a Transcutaneous Power Regulator for Artificial Heart

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Abstract - Based on commonly used parameters for a generic transcutaneous transformer model, a remote power supply using resonant topology for artificial heart is analyzed and designed for easy controllability and high efficiency. Primary and secondary windings of the transcutaneous transformer are positioned outside and inside human body respectively for energy transfer. The two large leakage inductances and the mutual inductance of the transformer are varying parameters of the coupling-coefficient which varies with transformer alignment and gap due to external positioning. Varying resonant-frequency resonant-tank circuits are formed using the transformer inductors and external capacitors to obtain a load insensitive frequency for the voltage transfer function at given range of coupling coefficients and loads. Previous researches usually use frequency modulation which may require a wide control frequency range well above the load insensitive frequency. In this paper, fundamental frequency study of the input-to-output voltage transfer function is carried out. Using the proposed control method, the switching frequency can be locked at just above the load insensitive frequency at heavy load for best efficiency. Specifically, above resonant operation in driving the resonant circuits when varying the coupling-coefficient is maintained using a digital-phase-lock-loop (PLL) technique to achieve zero-voltage switching of a full-bridge switches configuration which is also programmed to provide pulse-width-modulation (PWM) in controlling the output voltage. A prototype transcutaneous power regulator is built and found to have good efficiency and regulation in responding to changing alignment or gap of the transcutaneous transformer, load and input voltage dynamically.

I. INTRODUCTION

Advances in technology allow artificial organs or monitoring devices to be implanted into human body for the extension and improvement of human lives. The implants must operate inside the body for a considerable period of time and communicate with the outside world wirelessly with medical data and commands. Among power sources by chemical, nuclear or bio-chemical methods, chemical rechargeable batteries are by far the safer and popular reusable power for the implant inside the human body. The rechargeable battery is recharged remotely through the human skin using inductive link, like the one used in rechargeable toothbrushes with separate wireless chargers.

Totally implantable systems minimize the risk of infection and improve patient's mobility. Reliable and safe methods in providing power for the implants are the key factors in achieving totally implantable systems. Inductive link technology is widely used for powering different kinds of implants

This work was supported by Hong Kong Research Grants Council under project CERG PolyU 5306/06E.

[1]–[23]. Pure air coil type transformer has been used and studied in [1], [2], [4], [5], [6], [8], [10], [13] and [23], within which [1], [2] and [8] use class E power converter, for low power applications and space saving. Discussion on the importance of battery power source can be found in [14]. For higher power applications, such as powering the artificial heart for a power level of 10–60 W, magnetic cores are used [3], [7], [9], [11], [12], [15]–[22]. To compensate for the large leakage inductances due to transcutaneous separation, resonant type converter of class D [9], [13], [17], [20], [21] is used. More detailed studies in using resonant converter for compensating the leakage inductances have been conducted by Cho et. al [17], [20], [21]. However, their studies uses a wide frequency range for frequency modulation in controlling the converter at a less than ideal output-to-input voltage transfer ratio of 0.45 [17] and have not fully demonstrate an effective control method for output voltage regulation dealing with changing transformer coupling coefficient, input voltage and load.

In this paper, we will present a PWM-PLL control method for transcutaneous power regulator using a generic transformer model whose coupling coefficient is changing with transformer alignment and gap. The transformer model and its equivalent parameters commonly used in transcutaneous transformer [9], [13], [17], [20], [21] is given in Section II. In Section III, resonant tank circuits using external capacitors and the inductances of the transcutaneous transformer are studied. Fundamental frequency analysis is preformed for the input-to-output voltage transfer function to obtain an operation frequency upon which the voltage transfer function is least sensitive to both load and transformer variations. Design and control of a transcutaneous power converter in regulating output voltage against changing coupling coefficient, load and input voltage is presented in Section IV. A prototype transcutaneous power regulator is built and its static and dynamic performances are evaluated in Section V. Finally, we conclude in Section VI.

II. EXAMPLE LEAKAGE INDUCTANCES OF TRANSCUTANEOUS TRANSFORMER

We use the transformer model in Fig. 1(a) to represent a transcutaneous transformer where all components are loss-less. Using the power level and size of cores used in [9], [13], [17], [20], [21], we have built our own a transcutaneous transformer using Planar Core E64/10/50 for an air gap length (skin depth) ranging from 10 mm to 20 mm. The parameters

Fig. 1. Transformer model and compensation circuits.

TABLE I
TRANSCUTANEOUS TRANSFORMER PARAMETERS

Air gap (mm)	Mis- alignment (mm)	L_P (μH)	L_S (μH)	L_1 (μH)	L_2 (μH)	L_M (μH)	k
20	10	42.60	38.13	30.32	28.30	12.27	0.273
20	0	42.63	38.14	30.03	28.06	12.56	0.279
10	0	49.41	43.53	21.85	21.46	27.56	0.532

the example transformer emulated represent the trend of leakage inductances and mutual inductance of other transformer designs [3], [7], [9], [11], [12], [15]–[22]. Therefore, as our target is on the design and control of a power regulator for transcutaneous converter application, the parameters of the example transformer should be adequate. Some of the parameters measured for various coupling coefficients k are shown in Table I, where

$$\begin{aligned} L_P(k) &= L_1(k) + L_M(k), \\ L_S(k) &= L_2(k) + n^2 L_M(k), \\ k &= \frac{n L_M}{\sqrt{L_P L_S}}, \end{aligned} \quad (1)$$

and the turn-ratio of the transformer is $n = 0.895$. As indicated in Table I, all inductance values change with k , i.e. air gap or misalignment length. Normally, external capacitances C_P and C_S are added to form resonant tanks with the transformer inductances [17]. The frequency of resonant tanks is thus changed with changing k .

III. FUNDAMENTAL FREQUENCY ANALYSIS OF RESONANT TANK CIRCUITS USING TRANSCUTANEOUS TRANSFORMER MODEL

We consider the fundamental frequency equivalent circuit in Fig. 1(b) where the transformer has been compensated using external capacitors C_P and C_S to transfer energy from the ac equivalent source v_{AB} to the equivalent load R_E . The components in the secondary side of the ideal transformer are transformed to their equivalents in the primary side shown in Fig. 1(c) which is further transformed to its Thévenin's equivalent shown in Fig. 1(d) for the calculation of input-to-output voltage transfer function $\frac{v_{OS}}{v_{AB}}(\omega)$. We have from Fig. 1(c) and (d):

$$v_T = v_{AB} \frac{j\omega L_M}{Z_P}, \quad (2)$$

$$Z_T = -\frac{\omega L_M \left(\omega L_1 - \frac{1}{\omega C_P} \right)}{Z_P}, \quad (3)$$

$$\begin{aligned} v_{OP} &= v_T \frac{R_E/n^2}{Z_T + Z_S/n^2 + R_E/n^2 - j\omega L_M} \\ &= v_{AB} \frac{\omega L_M}{Z_P} \frac{R_E}{n^2 Z_T + Z_S + R_E - jn^2 \omega L_M}, \end{aligned} \quad (4)$$

where

$$Z_P = j \left(\omega L_P - \frac{1}{\omega C_P} \right), \quad (5)$$

$$Z_S = j \left(\omega L_S - \frac{1}{\omega C_S} \right). \quad (6)$$

The input-to-output voltage transfer function is determined from (4) as

$$\begin{aligned} G_v(\omega) &= \frac{v_{OS}}{v_{AB}} = \frac{nv_{OP}}{v_{AB}} \\ &= \frac{\omega n L_M R_E}{(Z_S + R_E) Z_P + n^2 \omega L_M^2} \\ &= \frac{n}{\frac{Z_P}{\omega L_M} + \frac{\Delta}{\omega^3 C_S C_P R_E L_M}} \end{aligned} \quad (7)$$

where

$$\begin{aligned} \Delta &= \omega^4 C_P C_S (n^2 L_M^2 - L_P L_S) + \omega^2 (L_P C_P + L_S C_S) - 1 \\ &= \omega^4 L_P C_P L_S C_S (k^2 - 1) + \omega^2 (L_P C_P + L_S C_S) - 1. \end{aligned} \quad (8)$$

It can be readily seen in (7) that if $\Delta = 0$, then G_v is independent of R_E . Solving for roots of (8), we obtain

$$\omega_L = \sqrt{\frac{\omega_P^2 + \omega_S^2 - \sqrt{(\omega_P^2 + \omega_S^2)^2 - 4(1 - k^2)\omega_P^2\omega_S^2}}{2(1 - k^2)}}, \quad (9)$$

$$\omega_H = \sqrt{\frac{\omega_P^2 + \omega_S^2 + \sqrt{(\omega_P^2 + \omega_S^2)^2 - 4(1 - k^2)\omega_P^2\omega_S^2}}{2(1 - k^2)}}, \quad (10)$$

where

$$\omega_P = \frac{1}{\sqrt{L_P C_P}}, \quad (11)$$

$$\omega_S = \frac{1}{\sqrt{L_S C_S}}. \quad (12)$$

TABLE II
PERCENTAGE CHANGE OF INDUCTANCE RATIO WHEN CHANGING k

	$\frac{L_S}{L_P}$	$\frac{L_2}{L_1}$	k
	0.895	0.933	0.273
	0.895	0.934	0.279
	0.881	0.982	0.532
Maximum change	-1.56%	+4.99%	

From Fig. 1(c), the input impedance Z_I is given as

$$Z_I = Z_P + \frac{\omega^2 n^2 L_M^2}{Z_S + R_E}, \quad (13)$$

which after some manipulations can be expressed as

$$Z_I = \frac{Z_P R_E}{Z_S + R_E} + \frac{\Delta}{\omega^2 C_P C_S (Z_S + R_E)}. \quad (14)$$

A. Leakage Inductance Compensation

It has been shown [17] that when the two leakage inductances are compensated using external capacitances satisfying $L_1 C_P = L_2 C_S$, $1/\sqrt{L_1 C_P}$ and $1/\sqrt{(L_S + L_M)C_P}$ are two roots of Δ and $|G_v| = n$ in (7). However, the compensation technique can only be optimized at one particular k at which the equality $L_1 C_P = L_2 C_S$ holds. When k changes, L_1 and L_2 change as well, leaving $|G_v|$ deviating from n .

We propose here to use capacitances satisfying

$$L_P C_P = L_P C_S \quad (15)$$

as an additional method in compensating the leakage inductances. In doing so, using the approximations

$$\frac{L_M}{L_P} \approx k \approx \frac{n^2 L_M}{L_S}, \quad (16)$$

equations (9), (10) and (7) reduce to

$$\omega_L = \frac{\omega_P}{\sqrt{1+k}} \approx \frac{1}{(L_P + L_M)C_P}, \quad (17)$$

$$\omega_H = \frac{\omega_P}{\sqrt{1-k}} \approx \frac{1}{L_1 C_P}, \quad (18)$$

$$G_v(\omega_L) \approx n, \quad (19)$$

$$G_v(\omega_H) \approx n. \quad (20)$$

The required ratio of $\frac{C_P}{C_S}$, indicated on columns $\frac{L_S}{L_P}$ and $\frac{L_2}{L_1}$ in compensating the inductances are given in Table II, showing better accuracy of the proposed compensation for the particular transformer at ω_H and ω_L in achieving $|G_v| \approx n$.

For the subsequence analysis, the compensation method (15) and approximation (16) will be used. We define

$$\omega_0 = \omega_P = \omega_S, \quad (21)$$

$$\omega_r = \frac{\omega}{\omega_0}, \quad (22)$$

$$Q = \frac{\sqrt{\frac{L_P}{C_P}}}{R_E}, \quad (23)$$

then

$$Q_S = n^2 Q \quad (24)$$

$$|v_{CS}| = \left| \frac{n^2 Q v_{OS}}{\omega_r} \right| \quad (25)$$

$$|v_{CP}| = \left| \frac{v_{OS}}{n k \omega_r^2} \left| j n^2 Q \left(\omega_r - \frac{1}{\omega_r} \right) + 1 \right| \right| \quad (26)$$

TABLE III
DESIGN SPECIFICATION OF THE TRANSCUTANEOUS CONVERTER

DC input voltage	DC output voltage	Load Resistance	Output power
30–60 V	24 V	9.6–48 Ω	12–60 W

Fig. 2. Transcutaneous Converter.

Fig. 3. Calculated frequency response of input-to-output voltage gain of the transcutaneous converter.

Fig. 4. Measured frequency response of input-to-output voltage gain of the transcutaneous converter.

Equations (25) and (26) are useful for determining capacitor voltage stress in deciding Q .

IV. DESIGN AND CONTROL OF TRANSCUTANEOUS POWER CONVERTER

A transcutaneous power converter shown in Fig. 2 has been built according to the specification shown in Table III. The components used are shown in Table IV. The calculated and measured frequency responses of input-to-output voltage gain without the control circuitry in Fig. 2 are shown in Figs. 3 and 4 respectively, indicating the lossless ideal model matches well with the experimental circuit. The frequencies f_{H_kmin} and f_{H_kmax} in Figs. 3 and 4 are $\omega_H/2\pi$ in (10) for minimum and maximum k 's respectively.

In previous design [17], frequencies well above the maximum of ω_H in (10) for all k 's were used for frequency modulation in controlling the output voltage. Therefore, the design did not make use of the benefits, such as load

TABLE IV
POWER COMPONENTS USED IN THE TRANSCUTANEOUS CONVERTER

Circuit component	value
Power MOSFET Q_1 – Q_4	IRF540
Capacitance C_P	12.82 nF / 200V
Capacitance C_S	14.57 nF / 200V
Schottky Diode D_1 – D_4	MBR2535L
Capacitance C_O	330 μ F
Transcutaneous transformer	Ferrocube 3F3, PlanarEcore64/10/50 N_P =19 turns of Litz wire AWG38# $\times$ 32 $\times$ 3 N_S =17 turns of Litz wire AWG38# $\times$ 32 $\times$ 3
Switching frequency	250–320 kHz

Fig. 5. Phase angle between the primary inductance current and the driving voltage when using the proposed PWM and PLL control.

insensitivity and small resonant tank circulation currents, in operating the converter near and just above ω_H .

We use a control technique of pulse wide modulation and phase lock loop shown in Fig. 5 in achieving operation of the converter in a frequency above and close to ω_H . The operation of the control circuit is briefly explained as follows. The switching frequency f_S of the converter is clamped between $f_{Smin}(=f_{H.kmin})$ and $f_{Smax}(=320\text{kHz} > f_{H.kmax})$ shown in Fig. 4. A digital signal i_D is produced using the zero crossing of the primary inductor current i_{L1} shown in Fig. 5. Phase difference of i_D and v_S is calculated and is locked to a fixed phase angle of θ_0 by modulating the switching frequency of the converter. The angle θ_0 is adjusted to be large enough for zero voltage switching of the full bridge switches. The operation frequency of the converter is therefore locked just above ω_H for full load while changing k . When there are changes for either a higher input voltage or lighter load, pulse width modulation will kick in, following Fig. 5, where the converter will operate at a frequency a bit higher than ω_H and phase angle θ .

From Fig. 5, we have

$$\theta = \theta_0 + \frac{\pi(1-D)}{2}, \quad (27)$$

$$|v_{AB}| = \frac{2\sqrt{2}}{\pi} V_{IN} \sin \frac{\pi D}{2} \quad (28)$$

$$|v_{OS}| = \frac{2\sqrt{2}}{\pi} V_O \quad (29)$$

where D is the duty cycle of the PWM control. The input-to-output voltage gain is modified to

$$G_d(D) = G_v(2\pi f_\theta) \sin \frac{\pi D}{2}, \quad (30)$$

where the phase lock loop circuitry will find a frequency f_θ satisfying (27) from which $G_v(2\pi f_\theta)$ in (7) is obtained for calculating G_d in (30). Fig. 6 shows the relation between

Fig. 6. Calculated duty cycle to input-to-output voltage gain transfer function for different loading conditions.

Fig. 7. Block diagram of control circuitry.

D and G_d for k_{max} and k_{min} at different loading conditions. Fig. 7 shows a block diagram of the control circuitry, implementing the PWM and PLL control of transcutaneous power converter. In the next section, we will evaluate the effectiveness of this control method.

V. EVALUATION OF TRANSCUTANEOUS POWER CONVERTER

A prototype shown in Fig. 8 of the proposed transcutaneous power regulator composed of converter in Fig. 2 and control circuit in Fig. 7 is built. Figs. 9, 10 and 11 give the transient responses of the transcutaneous power regulator undergone step changes in load, input voltage and coupling coefficient respectively. Fig. 12 gives the efficiency of the regulator at full load.

VI. CONCLUSION

Transcutaneous power regulator for artificial heart has been studied in this paper. Compensation methods have been

Fig. 8. Prototype transcutaneous power regulator.

Fig. 9. Output voltage transient response of the transcutaneous regulator under a step change of load.

studied using fundamental frequency analysis on the input-to-output voltage transfer function of a capacitor compensated transcutaneous transformer power converter. Based on the results of the analysis, new control method using both pulse width modulation and phase lock loop control has been proposed. The control method has been implemented in a transcutaneous power regulator. The power regulator has been evaluated for step changes in input voltages, output loads and transformer gaps as well as power conversion efficiency, giving excellent results.

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Fig. 10. Output voltage transient response of the transcutaneous regulator at full load under a step change of input voltage V_{IN} .

Fig. 11. Output voltage transient response of the transcutaneous regulator at full load under a change of transcutaneous transformer gap between 10mm and 20mm, i.e. between k_{max} and k_{min} .

Fig. 12. Measured converter efficiency of the transcutaneous power regulator at full load for different input voltages and coupling coefficients.

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