

Investigating an LCL Load Resonant Inverter for Inductive Power Transfer Applications

Chwei-Sen Wang, Grant A. Covic, *Member, IEEE*, and Oskar H. Stielau

Abstract—A inductor/capacitor/inductor load resonant inverter is investigated for inductive power transfer applications. The inverter uses a variable frequency controller and operates in discontinuous current mode. The steady state operation of this system is determined by a power flow balance between the inverter and the resonant tank. The results are used to design a system to achieve maximum power transfer.

Index Terms—Inductively coupled power transfer (ICPT) systems, inductor/capacitor/inductor (LCL).

I. INTRODUCTION

INDUCTIVELY coupled power transfer (ICPT) systems are designed to deliver power efficiently from a stationary primary source to one or more movable secondary loads over relatively large air gaps via magnetic coupling. Modern power electronics have enabled many new applications such as contact-less power supplies for professional tools [1], contact-less battery charging across large air-gaps for electric vehicles [2], compact electronic devices [3], mobile phones [4], and medical implants [5]. Other examples include material handling systems [6] and public transport systems [7] where the secondary systems are electrically isolated and move along a long track. The advantages of such systems are safety, reliability, low maintenance and long product life.

The mutual coupling of ICPT systems is generally weak. To deliver the required power and ensure equipment size remains manageable, it is necessary to operate at high frequency. In addition, resonant circuits are normally employed in the primary and/or secondary networks to boost the power transfer capability, while minimizing the required VA rating of the power supply.

Load resonant inverters are very popular for applications such as induction heating [8]. Induction heating inverters in the 10 to 100-kHz frequency range can be used for ICPT applications, but the design considerations are very different due to the difference in load characteristics. When analysing load resonant converters, the load is normally represented as a resistor in series with the primary inductance, but in many practical ICPT systems, the reflected reactance due to secondary resonance is comparable to this primary inductance [1]–[3], [5]. To analyze ICPT systems accurately, the secondary resonant circuit thus needs to be solved simultaneously with the primary resonant network.

Manuscript received August 21, 2003; revised November 20, 2003. Recommended by Associate Editor J. A. Ferreira.

The authors are with the Department of Electrical and Computer Engineering, The University of Auckland, Auckland 1309, New Zealand (e-mail: h_f_power@yahoo.com.au).

Digital Object Identifier 10.1109/TPEL.2004.830098

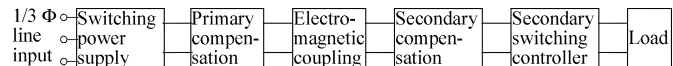


Fig. 1. Block diagram of an inductive power transfer system.

An inductor/capacitor/inductor (LCL) load resonant inverter can be operated with continuous or discontinuous current. Continuous current operation has been reported in detail for induction heating applications using a variable frequency controller [8]. Discontinuous current operation, however, has not been investigated thoroughly in the literature. In this paper, discontinuous current operation with a variable frequency controller is investigated for ICPT applications.

Sinusoidal ac analysis (fundamental mode analysis) is commonly used to analyze load resonant inverters [2], [5], [6], [8]. This approach is suitable for determining steady state characteristics, but is unable to identify the transient response or nonlinear features. Alternatively, time domain analysis using differential equations can be used [9]. However, this approach becomes too complex to be solved analytically as the number of resonant components is increased. In this paper, the inverter is modeled using time domain differential equations, while sinusoidal ac analysis is used to model the resonant tank. A power flow balance between the inverter and the resonant tank determines the steady state operation, enabling the inverter to be designed to achieve maximum power transfer to the load. The analysis is verified using a design example for contact-less electric vehicle battery charging.

II. FUNDAMENTAL STRUCTURE

A. Block Diagram of an ICPT System

An ICPT system essentially comprises two magnetically coupled electrical systems driven by a high frequency switching power supply. A block diagram of a typical system is shown in Fig. 1. The primary winding (stationary track or coil) is normally compensated in order to minimize the VA rating of the supply. Compensation of the secondary winding (movable pickup) enhances the power transfer capability of the system. A switched mode controller may be used to control the power flow from the pickup to the load. In more complex systems many individual pickups can exist, supplied by a single track.

B. ICPT System Driven by an LCL Load Resonant Inverter

The schematic in Fig. 2 shows an ICPT system using an LCL load resonant inverter. The subscripts “i,” “p,” “s,” and “R” stand for “inverter,” “primary,” “secondary,” and “load,” respectively. Various topologies can be used, with the advantages of

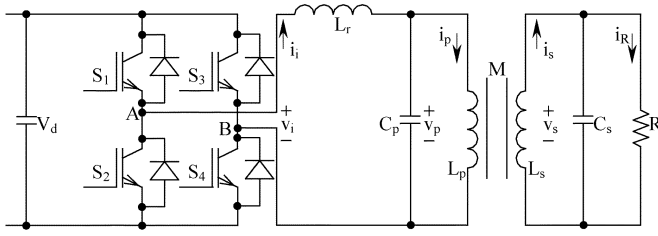


Fig. 2. Circuit diagram of an ICPT system driven by an LCL load resonant inverter.

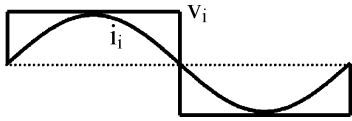


Fig. 3. Inverter voltage and current waveforms in continuous current mode.

the different topologies discussed in [10]. In this paper, parallel compensation was chosen for both the primary and secondary compensation. The switching power supply used in this system is a full-bridge inverter fed by a dc voltage V_d . The inductance of the primary winding is shown as L_p , and primary parallel compensation is implemented by connecting the primary compensation capacitor C_p in parallel with the primary winding. Together with the series inductor L_r , these three components form the LCL resonant circuit.

The load on this LCL resonant circuit is the reflected impedance from the secondary due to the magnetic coupling between the primary and the secondary, denoted as M . The secondary winding L_s is compensated by the secondary compensation capacitor C_s connected in parallel with the winding. An equivalent load resistor R represents both the secondary switching controller and the load.

III. CONTROL

A. Basic Modes of Operation

The inverter can be controlled using either a fixed or a variable frequency controller. Fixed frequency controllers have the advantage that a compensated pickup will always run at its correct tuned frequency. Variable frequency controllers usually control the operating frequency so that the inverter operates close to unity power factor. In this paper only variable frequency controllers will be discussed.

There are two distinct groups of variable frequency controllers, namely controllers that operate with a continuous inverter current and those that operate with a discontinuous inverter current.

B. Continuous Current Mode

In continuous current mode, the inverter is controlled to always ensure a path for the inverter current. This can be achieved by using the voltage amplification effect described in [8]. With this design the resonant tank voltage is much larger than the output inverter voltage, and the inverter current becomes nearly sinusoidal as shown in Fig. 3.

For optimal inverter efficiency, the frequency is controlled so that the inverter current is in phase with the inverter output

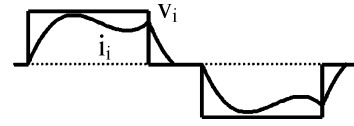


Fig. 4. Inverter voltage and current waveforms in discontinuous current mode.

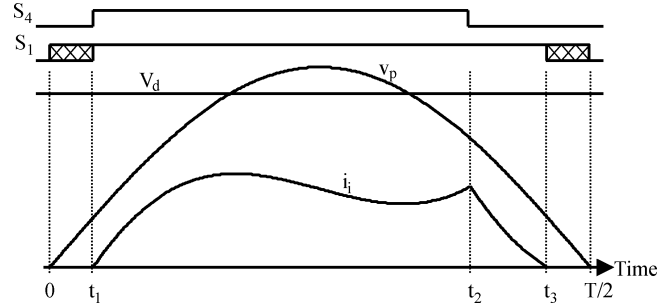


Fig. 5. Switching control in discontinuous current mode.

voltage. Switching losses are virtually nonexistent, and very high operating frequencies can be achieved. Control can be achieved by varying the frequency as described in [8], but the resulting switching losses need to be considered carefully.

C. Discontinuous Current Mode

In discontinuous current mode the inverter is switched so that the inverter current drops to zero and stays there for a finite period during each half cycle as shown in Fig. 4. The biggest advantage is that turn-on losses are eliminated, and only conduction and turn-off losses need to be considered.

It is further possible to control the switches so that inverter current is minimized. This is achieved by ensuring the inverter voltage always has the same polarity as the resonant tank voltage. With v_p positive, only S_1 and S_4 are allowed to turn on, while S_2 and S_3 are only turned on if v_p is negative. Fig. 5 shows current and voltage waveforms for half a cycle.

To ensure discontinuous current, t_2 has to be early enough to ensure that the current will drop back to zero before the end of the half cycle. Since the resonant tank voltage v_p provides the voltage to drive the inverter current back to zero, t_2 is a strong function of v_p (t_2 must reduce as v_p reduces).

Power control is achieved by varying t_1 . As t_1 increases, the inverter current drops resulting in lower power. With only turn-off and conduction losses, a trade-off between the switching and conduction losses is necessary. Using modern IGBT modules, conduction losses are equal to turn-off losses at a switching frequency of about 30 kHz. With such devices, this would be the maximum practical operating frequency.

IV. MODELLING THE DISCONTINUOUS CURRENT MODE

To analyze the LCL resonant inverter, a four-step process is followed:

Firstly an inverter model is developed in which the electromagnetic structure, primary and secondary compensation as well as the load are all modeled as a sinusoidal voltage source. With this simplification, the inverter current waveform as well as the inverter output power can be determined as a function of the resonant tank voltage.

Once the inverter current waveform is known, the fundamental component of the inverter current is calculated. Of particular interest is the phase relationship between the inverter current and the primary resonant tank voltage.

Next a resonant tank model consisting of the electromagnetic structure, primary and secondary compensation as well as the load is developed. A magnetic coupling model is used which calculates the effect of the secondary circuit on the primary circuit. If all parameters are known, the model can be used to calculate the power and phase relationships of the complete resonant tank.

Finally the two models are combined, and the steady state operating point is determined.

A. Inverter Model

In the inverter model shown in Fig. 6, all primary and secondary resonant tank components are replaced by a single sinusoidal voltage source v_p . This is a valid simplification for high Q resonant systems, where the resonant tank voltages are very close to sinusoidal.

The resonant tank voltage is given by

$$v_p(t) = \hat{V}_p \sin \omega t. \quad (1)$$

The inverter voltage depends on the states of the switches, and is equal to

$$v_i(t) = \begin{cases} v_p(t) & 0 < t < t_1 \\ V_d & t_1 < t < t_2 \\ 0 & t_2 < t < t_3 \\ v_p(t) & t_3 < t < \frac{T}{2} \end{cases} \quad (2)$$

With this inverter model, the inverter current i_i can be determined from

$$v_i(t) - v_p(t) = L_r \frac{di_i(t)}{dt}. \quad (3)$$

During the time interval from t_1 to t_2 , the inverter current i_i solved from (3) is

$$i_i(t) = \frac{V_d}{L_r}(t-t_1) + \frac{\hat{V}_p}{\omega L_r} [\cos \omega t - \cos \omega t_1] \quad t_1 < t < t_2. \quad (4)$$

During the freewheeling period after t_2 , the inverter current solved from (3) is

$$i_i(t) = \frac{V_d}{L_r}(t_2 - t_1) + \frac{\hat{V}_p}{\omega L_r} [\cos \omega t - \cos \omega t_1] \quad t_2 < t < t_3. \quad (5)$$

Since the inverter current drops to zero at t_3 , the time t_3 can be calculated as

$$t_3 = \frac{\cos^{-1} \left(\frac{\omega V_d}{\hat{V}_p} (t_1 - t_2) + \cos \omega t_1 \right)}{\omega}. \quad (6)$$

To ensure discontinuous current mode operation, t_3 has to be less than $T/2$. The maximum value of t_2 is thus

$$t_{2,\max} = t_1 + \frac{\hat{V}_p}{\omega V_d} (1 + \cos \omega t_1). \quad (7)$$

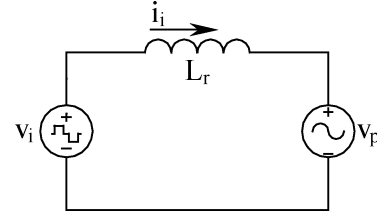


Fig. 6. Inverter model.

The average power supplied by the inverter is equal to

$$P_i = \frac{2}{T} \int_0^{\frac{T}{2}} v_i(t) i_i(t) dt. \quad (8)$$

Before t_1 and after t_3 , the inverter current is zero and the inverter supplies no power. During the freewheeling period from t_2 to t_3 , the inverter voltage is zero, and again the inverter supplies no power. The inverter only supplies power from t_1 to t_2 , when the inverter voltage is equal to the dc bus voltage V_d . Equation (8) can thus be rewritten as

$$P_i = \frac{2V_d}{T} \int_{t_1}^{t_2} i_i(t) dt. \quad (9)$$

Substituting (4) into (9) results in

$$P_i = \frac{\omega V_d^2}{2\pi L_r} (t_2 - t_1)^2 + \frac{V_d \hat{V}_p}{\pi \omega L_r} (\sin \omega t_2 - \sin \omega t_1 - \omega(t_2 - t_1) \cos \omega t_1). \quad (10)$$

This equation shows that the inverter power can be controlled by varying t_1 and t_2 .

B. Fundamental Component of the Inverter Current

Using Fourier analysis, the inverter current can be expressed as the sum of its Fourier components:

$$i_i(t) = \sum_{h=\text{odd}} \{A_h \cos(h\omega t) + B_h \sin(h\omega t)\} \quad (11)$$

where

$$A_h = \frac{4}{T} \int_0^{\frac{T}{2}} i_i(t) \cos(h\omega t) dt$$

$$B_h = \frac{4}{T} \int_0^{\frac{T}{2}} i_i(t) \sin(h\omega t) dt.$$

Substituting (4) and (5) into (11), the fundamental component of the inverter current is

$$i_{i1}(t) = A_1 \cos(\omega t) + B_1 \sin(\omega t) = C_1 \sin(\omega t + \Phi_1) \quad (12)$$

where

$$A_1 = \frac{2}{\pi\omega L_r} \times \left\{ V_d [(\omega t_1 - \omega t_2) \sin(\omega t_3) + \cos(\omega t_2) - \cos(\omega t_1)] \right. \\ \left. + \hat{V}_p \left[\cos(\omega t_1) \sin(\omega t_3) - \cos(\omega t_1) \sin(\omega t_2) \right] \right\} \\ B_1 = \frac{2}{\pi\omega L_r} \times \left\{ V_d [(\omega t_1 - \omega t_2) \cos(\omega t_3) + \sin(\omega t_2) - \sin(\omega t_1)] \right. \\ \left. + \hat{V}_p \left[\cos(\omega t_1) \cos(\omega t_3) - \cos^2(\omega t_1) \right] \right\} \\ C_1 = \sqrt{A_1^2 + B_1^2} \\ \Phi_1 = \tan^{-1} \left(\frac{A_1}{B_1} \right).$$

The phase angle Φ_1 is the angle between the resonant tank voltage and the fundamental of the inverter current, and gives a good indication of how well the inverter matches the load.

The analysis that follows can be simplified by expressing t_1 to t_3 as a ratio of the period to give

$$t_{1n} = \frac{t_1}{T} = \frac{\omega t_1}{2\pi} \\ t_{2n} = \frac{t_2}{T} = \frac{\omega t_2}{2\pi} \\ t_{3n} = \frac{t_3}{T} = \frac{\omega t_3}{2\pi}. \quad (13)$$

If t_{1n} and t_{2n} are controlled to be constant, it can be seen from (6) that t_{3n} will also be fixed, and all terms of the form ωt thus become constants. The phase angle Φ_1 in (12) thus becomes independent of the frequency and is only a function of V_p .

C. Resonant Tank Model

The electromagnetic structure, primary and secondary compensation as well as the load are now considered in more detail. The components are all shown in Fig. 7, using a mutual inductance coupling model [6] to combine the primary and secondary systems. The load is assumed to be linear and resistive.

The induced voltage in the secondary due to the primary current, and the reflected voltage in the primary due to the secondary current, are specified in terms of the mutual inductance M , the operational frequency ω , and the primary and secondary currents I_p and I_s .

For the parallel-compensated secondary shown in Fig. 7, the secondary impedance Z_s is equal to

$$Z_s = \frac{j\omega M I_p}{I_s} = j\omega L_s + \frac{1}{j\omega C_s + \frac{1}{R}}. \quad (14)$$

The reflected impedance of secondary as seen by the primary is found by dividing the reflected voltage $-j\omega M I_s$ by the primary current I_p , and is equal to

$$Z_r = \frac{-j\omega M I_s}{I_p} = \frac{\omega^2 M^2 I_s}{j\omega M I_p} = \frac{\omega^2 M^2}{Z_s}. \quad (15)$$

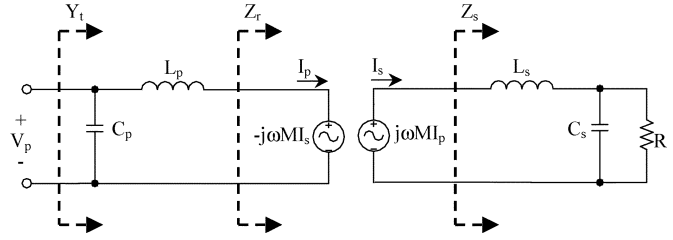


Fig. 7. Resonant tank model.

The total impedance of the circuit can now be calculated. It is, however, more convenient to express this in terms of an admittance Y_t , and can be determined from Fig. 7 as

$$Y_t = j\omega C_p + \frac{1}{j\omega L_p + Z_r}. \quad (16)$$

Finally the power transferred via the electromagnetic coupling structure can be calculated as

$$P_c = (\text{Re} Y_t) \frac{\hat{V}_p^2}{2} \quad (17)$$

where $\text{Re} Y_t$ represents the real component of the admittance.

D. Steady State Analysis

In the steady state analysis, the sinusoidal voltage source in the inverter model is replaced by the resonant tank model. Since the power and phase relationships have been calculated for both models, the steady state operating point can be determined by setting the phase and power of each model equal to each other.

Under steady state conditions it is assumed that the controller keeps the parameters t_{1n} and t_{2n} fixed. As has been shown, under these conditions the inverter phase angle becomes independent of frequency, and is only a function of the resonant tank voltage V_p . From the resonant tank model, on the other hand, the phase was determined as a function of frequency ω (16). Since the phase relationships have to be equal, the operational frequency ω can be found as a function of the resonant tank voltage V_p .

This relationship of frequency versus resonant tank voltage can now be used to determine the power transfer characteristic of both the inverter and the resonant tank. The power delivery characteristics of the inverter are very different to the power absorption characteristics of the resonant tank, and the inverter needs to be matched to the resonant tank for efficient power transfer. This is done by the correct choice of the series inductor L_r .

As L_r is decreased, the power delivery capability of the inverter increases at the expense of a larger inverter current. The aim is to choose L_r as large as possible (to minimize the inverter current) while still being able to deliver the required power to the resonant tank.

By plotting the inverter power for different values of L_r against the resonant tank power, a good choice of L_r can be made. An example of this procedure is shown in Section VI. In the example it can be seen that for large tank voltages the inverter limits the power transfer, while for low tank voltages the resonant tank limits the power transfer. A good choice for L_r is where the crossover point is at the maximum power absorption capability of the resonant tank.

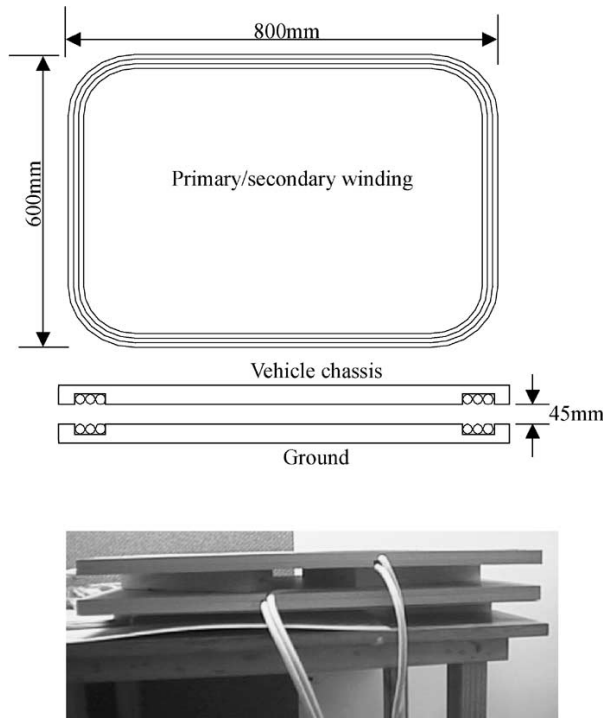


Fig. 8. Primary and secondary coils with 45 mm air gap.

TABLE I
PARAMETERS OF THE CONTACT-LESS ELECTRIC VEHICLE BATTERY CHARGER

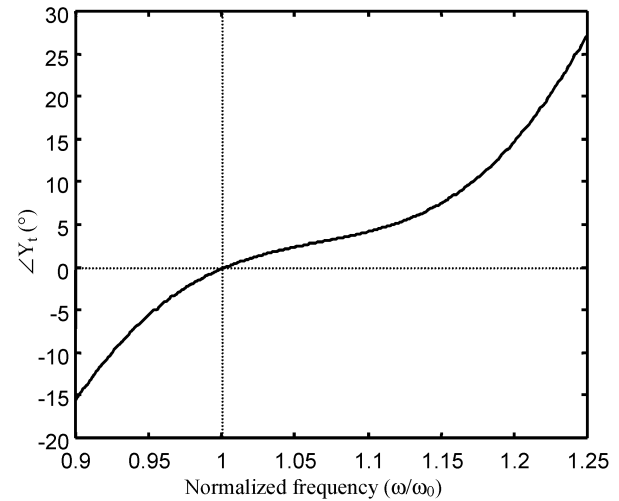
Nominal resonant frequency	20kHz
Rated power	30kW
Primary rated current	150A
Rated load	6Ω
Primary inductance	29.6μH
Primary capacitance	2.28μF
Mutual inductance	12.7μH
Secondary inductance	26.9μH
Secondary capacitance	2.42μF

Once the optimum value of L_r has been obtained, the power equations for the inverter and resonant tank can be set equal to each other. Since both equations are functions of ω and V_p , unique solutions for both the operating frequency and resonant tank voltage can be obtained.

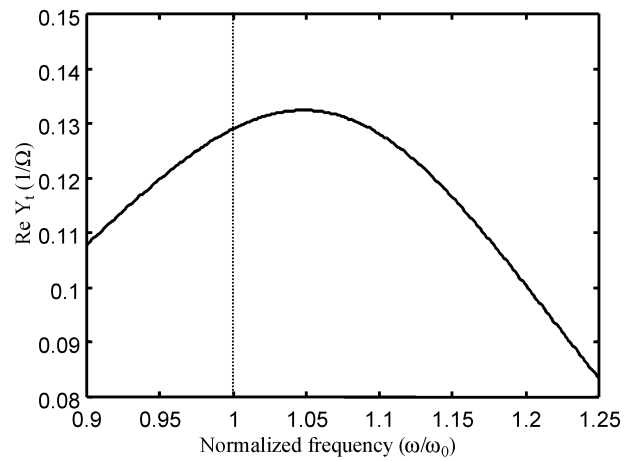
In the next section the above methodology is used to determine the operating conditions of a practical circuit.

V. DESIGN EXAMPLE

For the design example, a flat coil ICPT system for contact-less electric vehicle battery charging was chosen [11]. This system was designed to deliver 30 kW across a 45-mm air gap at a nominal frequency of 20 kHz with a primary current I_p of 150 A. The electromagnetic coupling structure is shown in Fig. 8



(a)



(b)

Fig. 9. Properties of the resonant tank admittance at rated load: (a) phase angle and (b) real component.

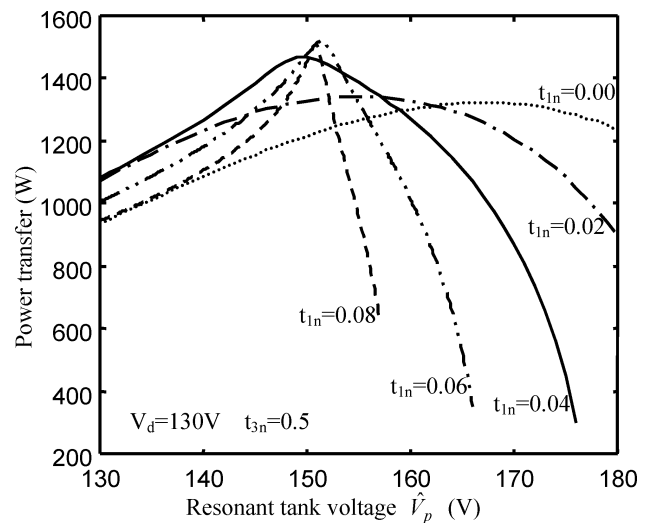


Fig. 10. Power transferred into the resonant tank.

and consists of identical primary and secondary windings. The secondary coil is attached to the underside of the electric vehicle, while the primary coil is buried in the ground. When an

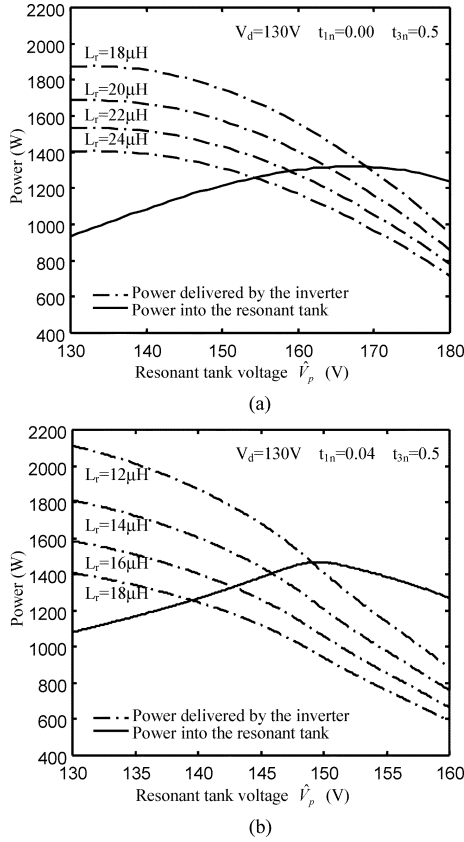


Fig. 11. Inverter and resonant tank power: (a) $t_{1n} = 0.00$ and (b) $t_{1n} = 0.04$.

electric vehicle stops over the charging bay, the primary is energised and electric power is transferred across the air gap via magnetic coupling between the primary coil in the ground and the secondary coil on the vehicle.

The design of the resonant tank follows the design procedure described in [10], [11], with the values of the measured coupling parameters as well as other relevant components given in Table I.

The operating frequency is allowed to vary around the design frequency of

$$\omega_0 = \frac{1}{\sqrt{L_s C_s}}. \quad (18)$$

The phase angle of the resonant tank admittance at rated load is shown in Fig. 9(a) as a function of the normalized operational frequency (ω/ω_0). As expected, the phase angle is zero at $\omega/\omega_0 = 1$.

The real component of the resonant tank admittance at rated load is shown in Fig. 9(b) as a function of the normalized operational frequency. The admittance reaches a maximum at a frequency slightly higher than the secondary resonant frequency. Assuming a constant resonant tank voltage, maximum coupling power can be achieved at this operating point, where the resonant tank runs at a small leading phase angle.

VI. VERIFICATION

The developed theory was compared with experimental results taken on a system with the dc bus voltage V_d set at 130 V.

TABLE II
STEADY STATE OPERATIONS FOR MAXIMUM POWER WITH SELECTIONS OF t_{1n}

V_d	130V				
R	6Ω				
t_{3n}	0.5				
t_{1n}	0.00	0.02	0.04	0.06	0.08
L_r	19μH	16μH	12μH	6.5μH	3.2μH
$\hat{V}_p$	166.4V	154.8V	149.1V	151.3V	150.7V
P	1323W	1342W	1467W	1518W	1507W
I_p	38.5A	36.2A	34.1A	34.8A	33.9A
Peak i_i	18.8A	18.9A	21.1A	25.9A	38.5A
f	24.0kHz	23.0kHz	21.1kHz	20.9kHz	20.5kHz
Φ_1	17.7°	9.1°	3.1°	1.8°	2.1°

TABLE III
SYSTEM BEHAVIOUR AT LIGHT LOAD

V_d	130V			
t_{1n}	0.04			
t_{3n}	0.5			
L_r	12μH			
R	6Ω	4Ω	2Ω	0Ω
$\hat{V}_p$	149.1V	158.5V	167.7V	177.1V
P	1467W	1006W	541W	0W
I_p	34.1A	34.9A	34.9A	39.5A
f	21.1kHz	22.5kHz	23.2kHz	21.7kHz

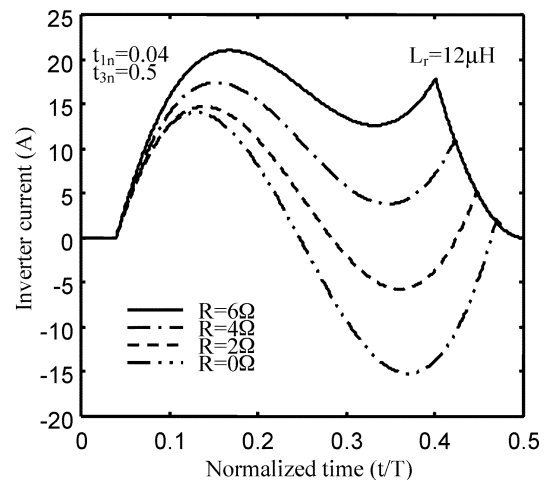


Fig. 12. Inverter current waveforms at light loads.

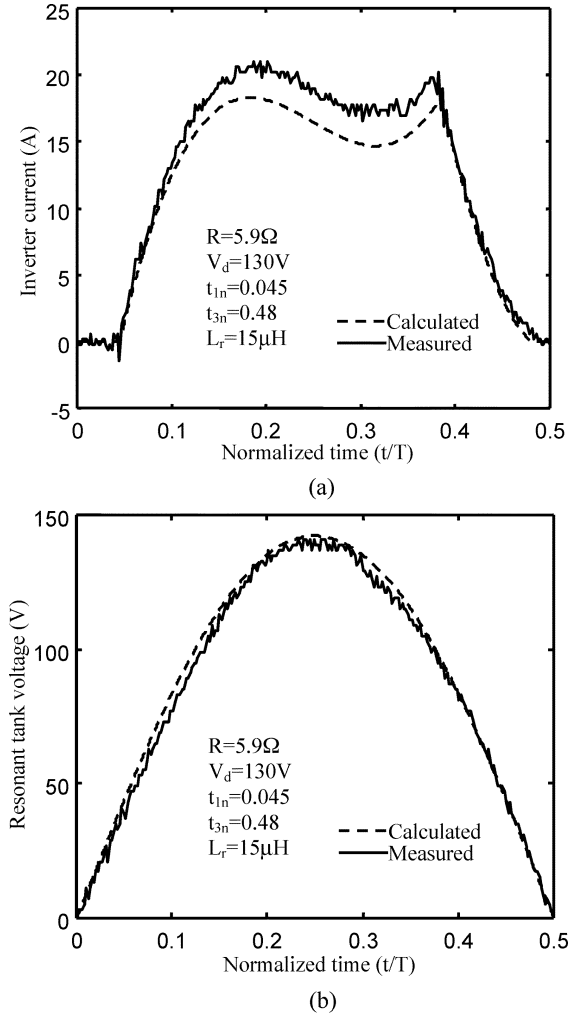


Fig. 13. Calculated and measured steady state operation.

TABLE IV
MEASURED FREQUENCY, PRIMARY CURRENT AND POWER

Frequency	20.9kHz
Primary current	31.5A
Input power into the DC link	1485W
Output power into the load	1275W
System efficiency	85.8%

At this voltage the primary current is much lower than the rated current of 150 A, and the power levels need to be scaled down correspondingly.

The power transfer at rated load is shown in Fig. 10 as a function of the resonant tank voltage. Here graphs are drawn for various selections of t_{1n} assuming the inverter current drops to zero at the very end of each half cycle ($t_{3n} = 0.5$). These graphs were obtained by first calculating the phase angle for each value of $\hat{V}_p$ using (12), equating this phase angle to the resonant tank phase angle determined in (16) to obtain the operating frequency ω , and then using this frequency in (17) to determine the power absorption capability of the resonant tank.

The graphs show a very nonlinear nature. As t_{1n} is increased, the maximum power transfer increases until t_{1n} is equal to about 0.06. Increasing t_{1n} further results in a drop in the maximum power transfer. In addition the range of tank voltages over which power can be transferred is a strong function of t_{1n} . Low values of t_{1n} give good power transfer over a wide range of resonant tank voltages, while larger values of t_{1n} result in a high power transfer capability over only a narrow range.

For Fig. 11(a) and (b), t_{1n} was chosen as 0.00 and 0.04, respectively, and the power delivery capability of the inverter plotted on the same graph as the power absorption capability of the resonant tank. The effect of the series inductor L_r can be seen clearly, with the inverter power delivery capability increasing as L_r is made smaller. The price to pay for a smaller series inductor, however, is an increase in the inverter current, which directly increases the required VA rating of the inverter. From an inverter perspective, the largest possible value for L_r is thus desirable. As described in Section IV-D, the optimum value of L_r should be chosen where the inverter power curve intersects the peak of the resonant tank power curve. From Fig. 11 it can be seen that both the inverter power delivery and the resonant tank power absorption depend on the choice of t_{1n} . Consequently, the optimum value of L_r is also dependent on t_{1n} . With t_{1n} equal to 0.00, the optimum inductance is about 19 μH , with the optimum inductance decreasing to about 12 μH when t_{1n} is increased to 0.04.

Various maximum power operating points for different values of t_{1n} are compared in Table II along with the optimum series inductance L_r . As t_{1n} is increased, the maximum power initially increases while the primary current reduces as the operating frequency moves closer to the secondary resonant frequency. However, the required series inductance also becomes smaller as t_{1n} is increased, causing the peak inverter current i_i to increase. As t_{1n} is increased further to 0.06, very little power gain is achieved, but the inverter current increases by more than 20%. A further increase in t_{1n} results in a reduced power transfer capability but a much larger inverter current, and should be avoided. For this application t_{1n} should be controlled close to 0.04 in order to maximize the power and minimize the primary current, while still keeping an acceptable peak inverter current level.

The analysis and design above assumes rated load. The same approach can be used to investigate the behavior of the system if the load is not at the nominal value.

The steady state operation at light loads are compared to rated load conditions in Table III assuming t_{1n} and t_{3n} are controlled at optimum values determined for rated load. The variation of the primary current at reduced load is small, a desirable characteristic for ICPT systems.

Inverter current waveforms are compared for different loading conditions in Fig. 12. As the load reduces, the current that needs to be turned off decreases. At loads of about 3 Ω and less, zero turn-off losses can be achieved by turning off the switches while the current is flowing through the freewheeling diodes.

SEMIKRON IGBT modules SKM150GB125D driven by SEMIKRON IGBT drivers SKHI22B are used in the test system. These IGBT modules have low turn-off losses but high conduction losses, for the chosen example conduction losses

are higher than turn-off losses. By choosing an IGBT with a lower on-state voltage, on-state losses can be traded off against switching losses.

The calculated and measured waveforms of the inverter current and the resonant tank voltage at steady state are shown in Fig. 13. The inverter is switched so that t_{1n} is equal to 0.04 and t_{3n} slightly less than 0.5. This ensures that the inverter current drops back to zero before the start of the next negative half cycle. The series inductor of 15 μ H is slightly larger than the optimal theoretical value of 12 μ H.

The measured operating frequency, primary current, and input and output powers are given in Table IV. These results compare well with the theoretical calculations in Table II despite the loss-less assumption of the analytical model.

VII. CONCLUSION

In this paper a LCL resonant inverter used in an inductive power transfer application is investigated. The variable frequency control philosophy chosen reduces the required inverter VA rating, but results in a complex design. This design is simplified by deriving the steady state operation using a power flow balance analysis between the inverter and the resonant tank. The theory predicts the inverter operating point and enables the optimum value of the series inductance to be found. It was found that the system exhibits very nonlinear characteristics, requiring special care when designing a controller. The theoretical analysis and design considerations were confirmed using a contactless electric vehicle battery charger.

REFERENCES

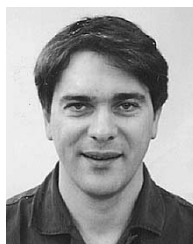
- [1] T. Bieler, M. Perrotet, V. Nguyen, and Y. Perriard, "Contactless power and information transmission," in *Proc. IEEE Industry Applications Conf.*, vol. 1, 2001, pp. 83–88.
- [2] R. Laouamer, M. Brunello, J. P. Ferrieux, O. Normand, and N. Buchheit, "A multi-resonant converter for noncontact charging with electromagnetic coupling," in *Proc. 23rd Int. Conf. Industrial Electronics, Control and Instrumentation (IECON'97)*, vol. 2, 1997, pp. 792–797.
- [3] H. Abe, H. Sakamoto, and K. Harada, "A noncontact charger using a resonant converter with parallel capacitor of the secondary coil," *IEEE Trans. Ind. Applicat.*, vol. 36, pp. 444–451, Mar./Apr. 2000.
- [4] Y. Jang and M. M. Jovanovic, "A contactless electrical energy transmission system for portable-telephone battery chargers," in *Proc. Telecommunications Energy Conf. (INTELEC'00)*, 2000, pp. 726–732.
- [5] G. B. Joun and B. H. Cho, "An energy transmission system for an artificial heart using leakage inductance compensation of transcutaneous transformer," *IEEE Trans. Power Electron.*, vol. 13, pp. 1013–1022, Nov. 1998.
- [6] J. T. Boys, G. A. Covic, and A. W. Green, "Stability and control of inductively coupled power transfer systems," *Proc. Inst. Elect. Eng.*, vol. 147, no. 1, pp. 37–43, Jan. 2000.
- [7] G. A. Covic, G. Elliott, O. H. Stielau, R. M. Green, and J. T. Boys, "The design of a contact-less energy transfer system for a people mover system," in *Proc. 2000 Int. Conf. Power System Technology*, vol. 1, Dec. 2000, pp. 79–84.
- [8] S. Dieckerhoff, M. J. Ruan, and R. W. De Doncker, "Design of an IGBT-based LCL-resonant inverter for high-frequency induction heating," in *Proc. IEEE Industry Applications Conf.*, vol. 3, 1999, pp. 2039–2045.
- [9] J. G. Hayes, M. G. Egan, J. M. D. Murphy, S. E. Schulz, and J. T. Hall, "Wide-load-range resonant converter supplying the SAE J-1773 electric vehicle inductive charging interface," *IEEE Trans. Ind. Applicat.*, vol. 35, pp. 884–895, July/Aug. 1999.
- [10] O. H. Stielau and G. A. Covic, "Design of loosely coupled inductive power transfer systems," in *Proc. IEEE Power System Technology Conf. (PowerCon'00)*, vol. 1, pp. 85–90.
- [11] C.-S. Wang, G. A. Covic, and O. H. Stielau, "General stability criterions for zero phase angle controlled loosely coupled inductive power transfer systems," in *Proc. IEEE Industrial Electronics Conf. (IECON'01)*, vol. 2, 2001, pp. 1049–1054.



Chwei-Sen Wang received the B.E. degree in mechanical engineering from the National Chiao Tung University, Hsinchu, Taiwan, R.O.C., the M.E. degree in mechanical engineering from National Taiwan University, Taipei, Taiwan, R.O.C., and the M.E. degree (with honors) in electrical and computer engineering from The University of Auckland, Auckland, New Zealand.

He is currently a Doctoral Fellow with the Foundation of Research, Science and Technology, New Zealand. His research area covers automatic

production systems, computer graphics, CAD/CAM, refrigerant compressors, room air conditioners, power electronics, and inductively coupled power transfer systems.



Grant A. Covic (M'88) received the B.E. (with honors) and the Ph.D. degrees from The University of Auckland, Auckland, New Zealand, in 1986 and 1993, respectively.

He is a Senior Lecturer in the Department of Electrical and Computer Engineering, The University of Auckland. His current research interests include power electronics, ac motor control, electric vehicle battery charging and inductive power transfer. He also has a strong interest in improved delivery methods for electronics and control teaching.

Dr. Covic received the IEE John Hopkinson Premium Award in 1999.



Oskar H. Stielau received the B.Eng., M.Eng., and D.Eng. degrees from the Rand Afrikaans University, Johannesburg, South Africa, in 1986, 1988, and 1991, respectively.

He currently consults in the high frequency power electronic field specializing in inductive technologies. Prior to that he spent two years with the Inductive Power Transfer Research Group, University of Auckland, New Zealand, and also seven years working in industry, mainly in the induction heating field.