

INDUCTIVE COUPLING MECHANISMS FOR CHARGING ELECTRIC VEHICLES

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Declaration

No portion of the work referred to in this thesis has been submitted in support of an application for another degree or qualification of this or any other university, or other institution of learning.

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Abstract

This thesis examines the problem of transferring power from the roadside to a stationary battery-powered electric vehicle across a large (upto 75mm) air gap.

Electric vehicles (EVs) have been considered as a means of reducing vehicle exhaust pollution and improving inner-city air quality. The operation of EVs however can involve complex battery charging procedures in order to ensure maximum safety and to prolong equipment life. The limited energy storage of the EV battery often results in the charge connection being made very frequently to partially recharge the vehicle's battery during periods of vehicle inactivity. This process is termed "opportunity charging" and it is becoming common in EV fleets with predefined operating schedules such as passenger buses. Operating an opportunity charging regime can lead to complex battery management procedures to ensure adequate vehicle range and battery life. The battery management procedures can be automated with an automatic battery charger which will be able to recharge the EV without manually plugging it in.

A novel method for transferring power from the wayside to a stationary electric vehicle is proposed and studied. The transmitter and receiver of power are separated by a 75mm air gap. The device power rating of 10kW is applicable to two different electric bus fleets. Power is transferred by electromagnetic induction. A frequency control solution to compensate for misalignment between the roadside and vehicle components is investigated. This control method has not previously been used to compensate for the misalignment of inductive couplings.

The process of transferring power across the large air gap is investigated by simulation and experimental methods. Finite element analysis software has been used to produce a magnetic model of the inductor cores. The power circuits, inductor cores and lead acid battery have been electrically simulated using simple ac analysis and computer simulations. Awareness of the underlying principles enables development of a unique prototype contact-less electric vehicle charging system. The prototype is subsequently evaluated experimentally. Results show that the required power can be transferred to the vehicle batteries across a significant air gap at efficiencies matching those of conventional conductive battery chargers.

The application of the system to an actual electric vehicle is considered. A design procedure is developed in order to apply this new technique to any electric vehicle installation.

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This thesis is concerned with the transfer of power from the utility grid to the vehicle traction battery across an inductive coupling. The inductive coupling must have a large air gap to allow the mounting of the secondary inductor on the underside of the vehicle while the primary inductor can be fitted flush with the road surface. To allow for vehicle positioning error the system must be tolerant of some physical misalignment whilst maintaining an acceptable rate of power transfer and operating efficiency.

2.5 Summary

The possibility of an automatic electric vehicle battery charger has been examined. The lead acid battery has been identified as the target battery for the inductive charging mechanism. Although possibly the poorest performing battery in terms of specific energy and specific power it is by far the cheapest and this has led to its widespread proliferation. It is indeed the battery technology chosen by a considerable number of advanced EV manufacturers. Regardless of battery type the EV has a much lower range than a conventionally fuelled vehicle.

It has been proposed that opportunity charging could be used to overcome the poor range of the EV. Opportunity charging takes advantage of working breaks and rest periods to partially recharge the on-board battery of a battery electric vehicle. An automatic charging system will require no intervention from the user and therefore the frequent connection which will be made between the vehicle and the utility grid will be transparent to the operator. The target vehicle for the charger has been identified as a small electric bus such as the Oxford electric bus. It has been calculated that a charger capable of delivering 10kW of power will be suitable for the Oxford electric bus requirements. The system developed within this thesis will therefore be aimed at delivering a power of 10kW.

In order to offer a totally automated battery charging system the charge connection needs to be made without the intervention of the user. Inductive coupling offers a solution where power can be transferred across air gaps hence eliminating the need to 'plug in' the battery operated device for charging. Inductive charging using closely coupled inductors to replace plugs and sockets has been proven to offer safety advantages such as the ability to operate in all weathers and inherent isolation of the EV electrical system from the utility network. Other systems offer the possibility of hands free operation with small air gap couplers (8mm to 10mm). The systems require very accurate parking and any misalignment results in a significant loss of power transfer

capability. A third system of inductive power transfer to EVs has been developed to transfer power to vehicles moving along tracks or inductively powered roadways. Both suffer from poor efficiency, high weight and misalignment problems. Inductive coupling technology has been identified as the preferred technology for an automatic battery charger but none of the available devices to date are suitable. The requirements for a totally automatic inductive charging system have been identified.

Key standards that will be applicable to the automated inductive charging device have been identified. Various international standard setting bodies are involved in implementing EV standards. Collaboration between the USA and Europe has seen common standards such as the Mode (Level) one to three charger definitions (see section 2.3.1). As with any electrical product EMC is an important issue. There are no European EMC standards relating specifically to inductive charging equipment therefore the Generic standards have to be consulted. Another important area is human exposure to electromagnetic fields. A large air gap inductive charger will inherently produce magnetic fringing fields. Although Europe and the UK have no specific standards there are guidelines in place which are likely to be adopted as standards and therefore the device must conform to these.

The assumptions made in calculating the device inductance are summarised as follows:

- The windings are constructed from a single turn of copper whose area is equal to the available area in the active length of the device.
- The end windings occupy the same area as the windings in the active length and they are situated immediately along side the inductor core.
- The inductors are located in free space with no external magnetic field to influence them.
- The results are calculated for the active length of both the primary and secondary windings and the orthogonal end windings as shown in Figure 3-4. The effects of the winding "corners" is neglected. The effects of the ferrite outside of the windings of the active length is neglected in the calculation of the end winding inductance.

Figures 3-5 and 3-6 shows the effects of core alignment in three dimensions. In each case both windings consist of 20 turns which was the maximum number of turns for the device and selected cable (See section 3.1.4).

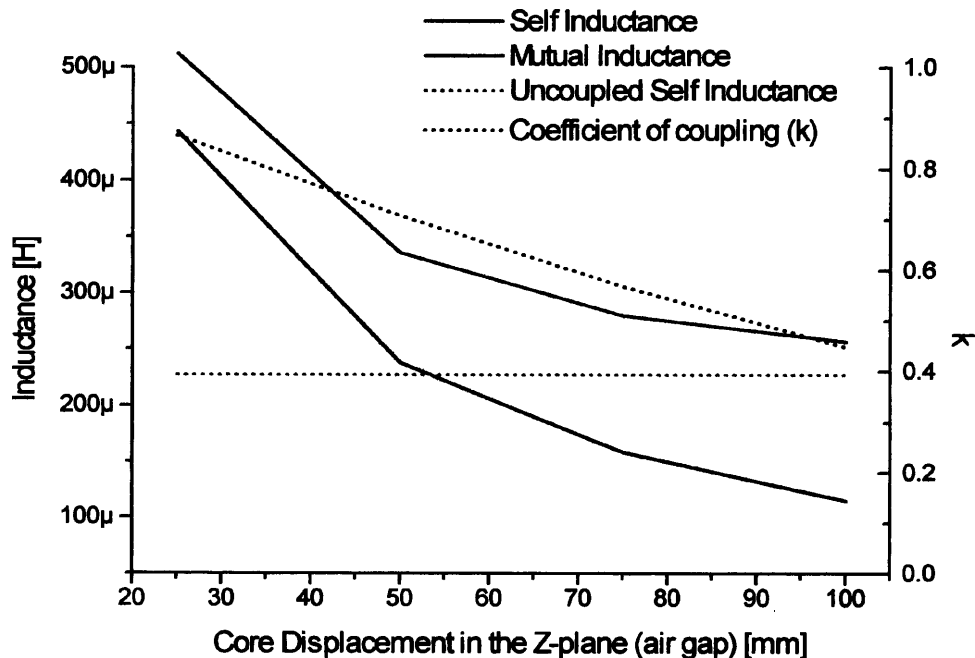
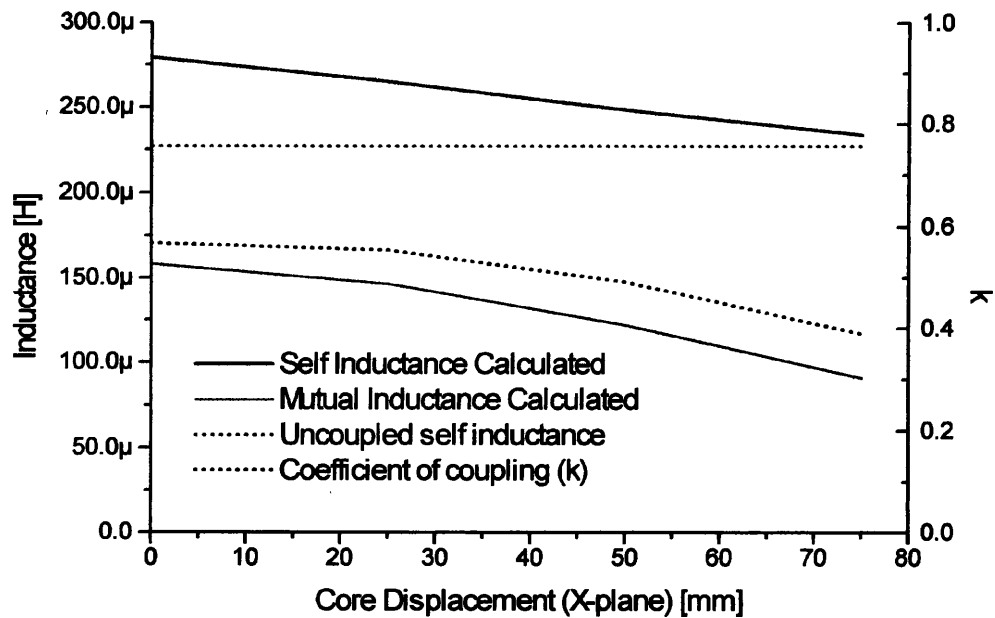


Figure 3-5 Effect of changing air gap on device inductance

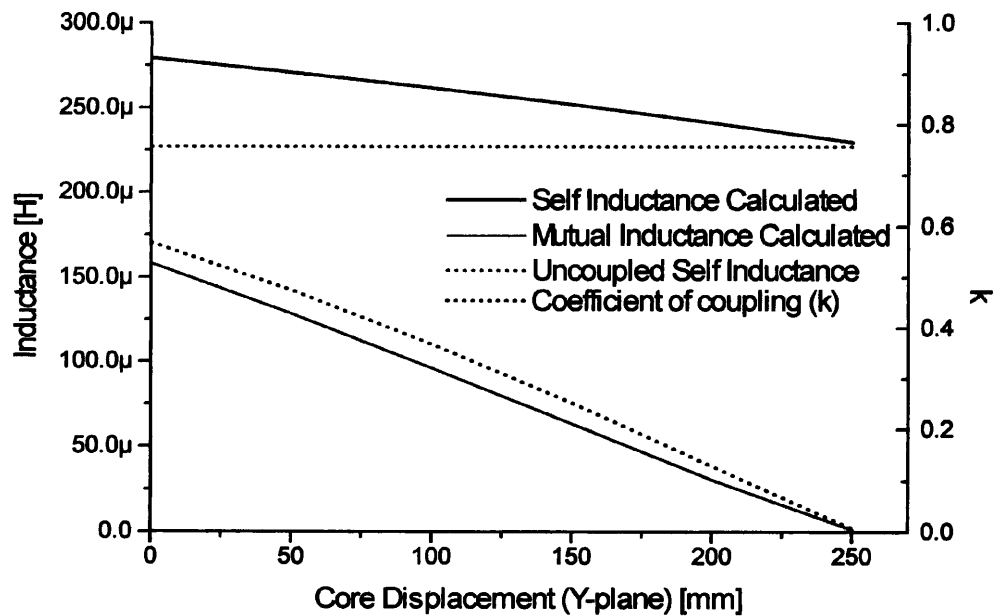
For a small air gap the windings are tightly coupled and the inductance of the windings is high. As the air gap increases the coupling factor (k) reduces and both the self and mutual inductance of the coils reduces. At large air gaps the self inductance of both coils tends towards the air-cored inductance, i.e. the self inductance of the coil behaves as if the second coil did not exist. Also as the air gap increases the mutual inductance tends towards zero, i.e. the air gap acts as if it were infinite. As the devices misalign the self and mutual inductance decrease. It is clear that the electrical characteristics of the coupling will vary considerably within the physical operating requirements.

For the development of the power circuit the effect of misalignment has only been analysed in one dimension, since it is clear that misalignment in either the X or Y plane or both simultaneously will yield similar results. As the device misaligns the self inductance of each winding will decrease and tend towards the air-cored self inductance. The mutual inductance will decrease and tend towards zero.

The number of turns for each winding depends upon the load, the required power transfer, the mains interface and the size of the coupling, all of which are related to the application vehicle. The number of turns has been set to 20 for both the primary and secondary winding of the universal coupling. This is the maximum number of turns possible for the cable specified in the next section and therefore yields the maximum mutual inductance.



(a) Core misalignment X-Plane



(b) Core misalignment Y-Plane

Figure 3-6 Effect of core misalignment on device inductance
(air gap $Z=75\text{mm}$)

3.1.4 Calculation of winding resistance

To complete the model of Figure 3-3 the winding resistance needs to be calculated. The resistance of a solid or simple un-insulated stranded conductor increases significantly with frequency. The increase in resistance significantly increases cable losses. The apparent increase in resistance is due to a phenomenon known as skin effect. The skin effect is a function of frequency and causes a greatly increased resistance in the centre of the conductor decreasing towards the surface of the conductor. A measure of the skin effect is given by the 'skin depth'. The skin depth is the width from the outer skin of the conductor where the magnitude of the current has reduced to 1/e or 37% of the current magnitude at the surface. The skin depth is given by the following equation:

$$\delta = \frac{1}{\sqrt{\pi f \mu_r \mu_0 \sigma}} \quad \text{Equation 3-19}$$

Where,

δ = The skin depth [m]

f = frequency [Hz]

μ_r = relative permeability

μ_0 = permeability of free space

σ = conductivity of the cable material

In a cylindrical wire the skin effect results in most of the current flowing within the outer region of the wire. Many high frequency conductors are made of hollow tubes since the middle region of copper is carrying little or no current. However to achieve an acceptable current density within the copper the tubes have to have a very large diameter.

In order to overcome the skin effect conductors whose diameter is smaller than the skin depth can be chosen. In order to increase the current carrying capability several of these thin insulated strands may be paralleled and twisted so that each strand occupies an equal length in each part of the conductor. Cable constructed in this way is called Litz wire. The average ac impedance of each strand is equal since the current will be forced to distribute itself evenly within the conductor, Weinberg [75].

In overcoming the skin effect Litz wire introduces another ac loss mechanism; the proximity effect, Murgatroid, [76]. The proximity effect occurs if a solid conductor is in

impedance of the resonant circuit behaves like a pure resistance. The inverter current is in phase with the inverter square wave output voltage. The power switches will switch on and off at zero current and therefore have zero switching losses. Operation at resonance is utilised by Green^[56-58] however the fixed frequency nature of the circuit suffers from poor tolerance to load changes and device misalignment.

Operation above the resonant frequency means the impedance of the resonant circuit will be inductive, that is the load current will lag the applied square wave voltage. The IGBTs will switch on at zero current however they must switch off a current and are therefore subject to turn-off losses. Operation below the resonant frequency results in a capacitive impedance. The load current will lead the applied square wave voltage. The IGBTs will switch off at zero current however they must switch on a current and are therefore subject to turn-on losses.

Operating above resonance has many advantages. The IGBT free wheel diode (FWD) need only to be rated for the maximum switching frequency of the inverter since there are no switching stresses applied to it. Below resonance the FWD needs to be of the fast recovery type to minimise switching stress and turn off losses. During turn off low loss snubbers can be employed to reduce device over voltages. The low loss snubber is just a capacitor connected across the switching device. Any energy stored within the capacitor is not dissipated in an external resistance but is dissipated within the IGBT the next time it is turned on. Below resonance turn on snubbers consisting of a resistor, capacitor and diode (RCD) are required to minimise turn on switching stress. The switching energy stored in the capacitor is dissipated in the snubber resistor therefore adding to losses. Operation above resonance imposes a definite minimum operating frequency therefore all of the filter components can be designed with little overhead. Operation above resonance also allows a higher operating frequency for a predefined inductor / capacitor resonant circuit.

3.4 Summary

A novel universal inductive coupling with a large air gap has been developed for transferring power from the wayside to an electric vehicle. Practical aspects such as operating frequency, core material and cabling have been addressed. The electrical characteristics of the coupling have been modelled using finite element analysis techniques. The effects of device misalignment on the electrical equivalent circuit of the coupling have been examined.

Deliberately introducing a large air gap into a transformer results in a poor coupling factor (k). The leakage inductances of both the primary and secondary windings are large. Under no load the voltage induced in the secondary winding as a result of a primary winding current is greatly reduced by the volt drop across the primary winding leakage reactance. When loaded the load voltage is considerably lower than the secondary induced voltage due to the voltage drop across the secondary winding leakage reactance.

The low magnetising inductance of the coupling results in a large primary magnetising current which leads to high copper losses. The primary magnetising current can be reduced by increasing the operating frequency however under load the effects of the leakage reactances becomes greater at higher frequencies and the load voltage reduces with increasing operating frequency.

The large leakage reactances can be compensated for by the addition of a secondary side parallel capacitor. The capacitor effectively offers a reactance opposite to the inductive leakage reactance which results in reactive cancellation and a high secondary load voltage at high frequency. The amount of compensation is determined by the Q factor of the secondary side of the coupling. The Q factor is defined as the ratio of real to reactive power in the circuit. Too high a Q like the configuration selected by Green^[56-58] results in high circulating currents and inefficient operation with a need for significantly higher circuit component rating. With a Q set too low the reactance will not be effectively compensated.

As the coupling device misaligns the inductance and coupling factor of the coupling will change. Misalignment significantly increases leakage reactance and lowers magnetising inductance. This causes a reduction in the power transfer capability. A simple control solution is required to regulate the power transfer under a variety of alignment conditions.

Resonant inverters have been suggested to supply the circuit with high frequency current. The resonant inverter consists of either a half bridge or full bridge IGBT converter. The IGBT switches can be operated with a 50% duty cycle to supply the circuit with a square wave voltage. The inductance of the coupling forms the only inductive element in the resonant tank of the inverter which is completed with the simple addition of a primary side capacitor. The primary capacitor can be either parallel or series connected to form either a parallel or series resonant converter.

Perfect alignment will require an operating frequency of 8.3kHz. If the primary capacitor is fixed at $3\mu\text{F}$ the inverter current phase angle will be $+60^\circ$. For $X=75\text{mm}$ misalignment the operating frequency will be 6.1kHz, if the primary capacitor is left at $3\mu\text{F}$ the inverter current phase angle will be -26° .

To reduce the frequency variation the secondary tuning capacitor can be chosen to give a greater Q . This will result in a steeper frequency response and greater selectivity at the expense of much higher primary and secondary side circulating currents. Alternatively a complex system of individual capacitors for each possible misalignment condition could be employed. Appendix 7 details the ac analysis of the parallel resonant inverter driven coupling.

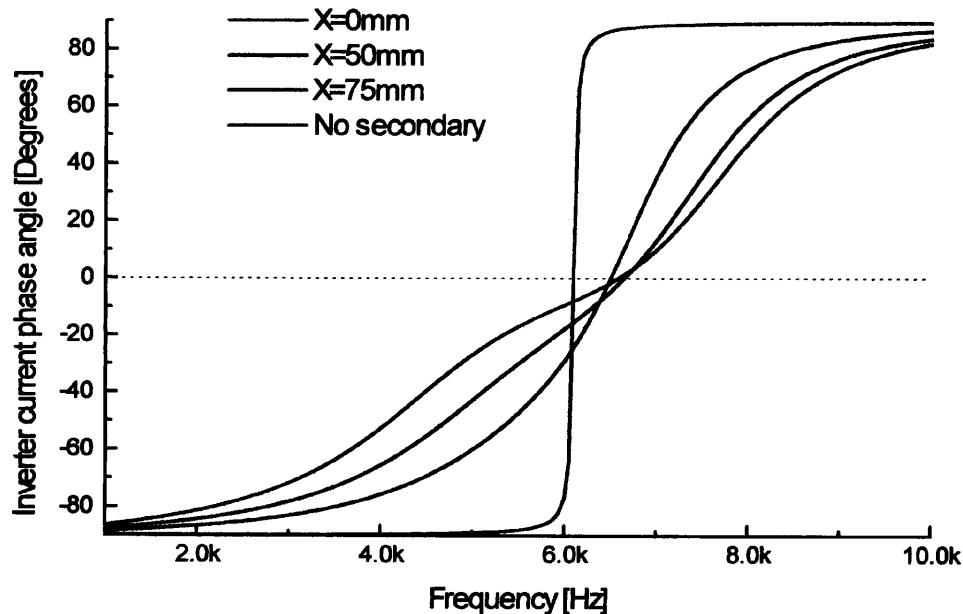


Figure 4-12 Inverter current phase angle for parallel resonant inverter driven coupling ($C_p=3\mu\text{F}$, $C_s=3\mu\text{F}$, $R_L=10\Omega$, $Z=75\text{mm}$, $Y=0\text{mm}$)

4.2.3 Series resonant inverter driven coupling

The series resonant inverter is shown in Figure 4.6(b). It is used by Bleijs in the Praxitele IPT system^[29,30]. In Bleijs' work he uses an uncompensated secondary winding therefore suffering from the effects of the secondary leakage reactance. The primary tuning capacitor, C_p , is placed in series with the primary winding to cancel the primary side leakage reactance.

In this thesis the combination of the series resonant inverter driven coupling with secondary side parallel capacitor compensation is proposed and examined. The resulting frequency response is shown in Figure 4-13 for the case where $C_p = C_s$ ($3\mu\text{F}$). Like the parallel primary capacitor the series primary capacitor compensates for the effects of the primary leakage reactance. The important difference is that the series primary capacitor has the effect of increasing the primary terminal voltage and therefore raising the voltage induced in the coupling. The secondary leakage reactance is compensated by the secondary parallel capacitor. As the device is misaligned the coupling factor (k) decreases. If the circuit is operated at a frequency near to 6.25kHz the secondary voltage and the rate of power transfer will rise. This phenomena is a result of the resonant network.

The circuit is actually two coupled resonant circuits, the primary and secondary. The Q factor of the secondary circuit (Q_2) is given by Equation 3-28. As the device misaligns L_2 increases and Q_2 reduces provided the load remains constant. This tends to reduce V_2 . The characteristics of the primary side resonance are defined by the series capacitor (C_p) and the input impedance to the coupling device, Z_i . Z_i contains a real component R_{Zi} and a reactive component X_{Zi} . We know from Chapter 3 X_{Zi} is always inductive and increases with misalignment.

The quality factor of the primary side can be approximated by Equation 4-9 where $L_{Zi} = X_{Zi}/\omega$.

$$Q_1 = \frac{1}{R_{Zi}} \cdot \sqrt{\frac{L_{Zi}}{C_p}} \quad \text{Equation 4-9}$$

Misalignment causes L_{Zi} to increase which in turn increases Q_1 . The result is an increase in the voltage induced in the coupling leading to an increase in secondary terminal voltage. Careful component selection allows frequency control misalignment regulation over a small frequency band. The operating frequency can be either above or below the resonant frequency. Operation above the resonant frequency is preferred since the higher frequency operation results in a lower flux system and the operating frequency can coincide with the resonant frequency of the secondary side. By selecting C_p to be larger than C_s the resonant frequency of the primary side can be shifted to a lower frequency than the secondary side. This is shown in Figure 4-14. This offers similar frequency compensation with a lower Q_1 resulting in a lower primary voltage

and current. Appendix 8 details the ac analysis of the series resonant inverter driven coupling.

The main advantage of the series resonant inverter is the inherent increase in primary winding current as the device misaligns. This allows a much smaller frequency band for misalignment frequency control. The circuit also maintains a low Q therefore minimising circulating current and improving efficiency. The main disadvantage of both types of inverter used in this application with frequency control for misalignment is the primary current phase angle. In the parallel inverter the phase angle changes considerably as the device misaligns and the frequency is varied. For the series resonant inverter the phase angle remains almost constant however lags the applied voltage. Both circuits will lead to IGBT switching losses within the inverter.

After consideration of both inverter types the series inverter driven coupling has been selected as the preferred topology.