

Wide-Load-Range Resonant Converter Supplying the SAE J-1773 Electric Vehicle Inductive Charging Interface

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Abstract—The recommended practice for electric vehicle battery charging using inductive coupling (SAE J-1773), published in January 1995 by the Society of Automotive Engineers, Inc., outlines values and tolerances for critical vehicle inlet parameters which must be considered when selecting a coupler driving topology. The inductive coupling vehicle inlet contains a significant discrete capacitive component in addition to low magnetizing and high leakage inductances. Driving the vehicle interface with a variable-frequency series-resonant converter results in a four-element topology with many desirable features: unity transformer turns ratio; buck/boost voltage gain; current-source operation; monotonic power transfer characteristic over a wide load range; throttling capability down to no load; high-frequency operation; narrow modulation frequency range; use of zero-voltage-switched MOSFET's with slow integral diodes; high efficiency; inherent short-circuit protection; soft recovery of output rectifiers and secondary dv/dt control and current waveshaping for the cable, coupler, and vehicle inlet, resulting in enhanced electromagnetic compatibility. In this paper, characteristics of the topology are derived and analyzed using two methods. Firstly, the fundamental mode ac sine-wave approximation is extended to battery loads and provides a simple, yet insightful, analysis of the topology. A second method of analysis is based on the more accurate, but complex, time-based modal approach. Finally, typical experimental results verify the analysis of the topology presented in the paper.

Index Terms— Battery charging, electric vehicle, inductive charging, resonant converters.

I. INTRODUCTION

INDUCTIVE coupling is a method of transferring power magnetically rather than by direct electrical contact and the technology offers advantages of safety, power compatibility, connector robustness, and durability to the users of electric vehicles. An off-vehicle high-frequency power converter feeds the cable, coupler, vehicle charging inlet, and battery load. The electric vehicle user physically inserts the coupler into the vehicle inlet where the ac power is transformer coupled,

rectified, and fed to the battery. The technology is presently being researched and productized at levels ranging from a few kilowatts to hundreds of kilowatts [2]–[5]. A recommended practice for inductive charging of electric vehicles, SAE J-1773, has been published by the Society of Automotive Engineers, Inc. (SAE) [1]. The specifications, as outlined in SAE J-1773, for the coupler and vehicle inlet characteristics must be considered in selecting a driving topology. Among the most critical parameters are the frequency range, the low magnetizing inductance, the high leakage inductance, and the significant discrete parallel capacitance.

This paper discusses the applicability of a variable-frequency superresonant, series-resonant converter driving the vehicle inlet. The combination of the two passive elements of the series-tank impedance with the two significant parallel elements of the inductive coupling vehicle inlet results in a four-element series-parallel *LCLC* (SP-LCLC) converter with a capacitive output filter. The large leakage inductance forms part of the inductance of the series-resonant converter.

Two-element and three-element *LLC* and *LCC*-type resonant circuits have been comprehensively discussed and analyzed. Four-element and higher order topologies provide many variations with unique characteristics, but have not been treated comprehensively in the literature due to their complexity and number. However, higher order models must be considered if, as discussed in this paper, the impact of the magnetizing inductance in the transformer of the inductive coupler is considered. Consequently, the three-element series-parallel *LCC* type becomes a four-element series-parallel *LCLC* converter if this magnetizing inductance is included in the analysis.

The four-element SP-LCLC converter has previously been analyzed with an inductive output filter [7]. The use of a capacitive output filter has been discussed in [8] and [9] for the simple *LC* parallel resonant circuit and in [10] and [11] for the three-element *LCC* circuit. The analysis of such converters is complicated, because the rectifier and capacitive output filter stage are decoupled from the resonant stage for a significant period during the switching cycle. As a result, the number of active resonant elements changes during the switching cycle, giving so-called multiresonant operation.

Two methods are used to analyze the characteristics of the four-element resonant topology. Firstly, an extension of the Fundamental Mode ac sine-wave Approximation (FMA) [12] is developed which gives a simple, yet insightful, analysis of

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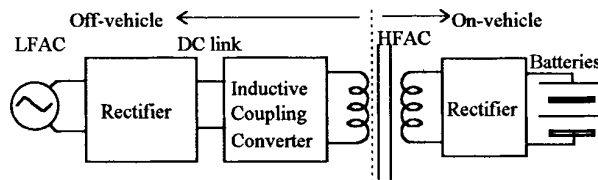


Fig. 1. Block diagram of power flow from the utility to the electric vehicle batteries.

the topology. In particular, this approach results in useful equations which describe the current-source charging characteristic of the converter.

A second method of analysis is based on the more accurate, but complex, time-domain modal approach. In this technique, the dynamic equations of the converter are solved based on a judicious state variable transformation [6], [7]. This analysis generates accurate topology design curves, demonstrates key converter characteristics, and vindicates the simpler FMA approach, while highlighting areas of operation where discrepancies between the two methods exist.

The objectives of this paper are firstly to describe the inductive coupling application, as defined in SAE J-1773, together with the characteristics required of the off-board converter topology. These issues are discussed in Section II. The basic SP-LCLC converter topology and its operation are described in Section III. Section IV presents the fundamental mode analysis, while the modal analysis is presented in Section V. The results from an experimental prototype and from the two analyses are presented and discussed in Section VI.

II. INDUCTIVE COUPLING

A. Basic Principles

The basic principle underlying inductive coupling is that the two halves of the inductive coupling interface are the primary and secondary of a take-apart transformer. When the charge coupler (i.e., primary) is inserted into the vehicle inlet (i.e., the secondary), power can be transferred magnetically with complete electrical isolation, as with a standard transformer.

The simplified power flow diagram for inductive charging of electric vehicles is shown in Fig. 1. The charger rectifies the low-frequency 50/60-Hz ac (LFAC) utility power to dc for conversion to high-frequency ac (HFAC) power. The HFAC is transformer coupled at the vehicle inlet, rectified, and supplied to the batteries. Battery charging information is fed back to the primary side using a radio-frequency communication link contained within the coupler and inlet.

According to SAE J-1773, the inductive coupler can be represented by the equivalent circuit model shown in Fig. 2(a). For analysis, the transformer model can be simplified further to include only the topologically significant components, as shown in Fig. 2(b). These are the magnetizing inductance L_P , the reflected parallel capacitance C_P , and the lumped leakage inductance L_L , equal to the sum of the primary and the reflected secondary leakage inductances L_{pri} and $(n/4)^2 L_{sec}$.

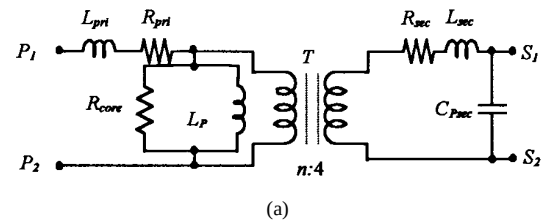
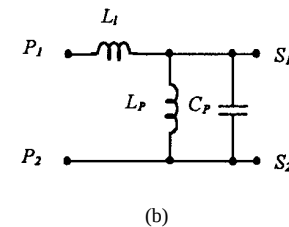


Fig. 2. Simplified electrical equivalent circuits of inductive coupler.

TABLE I
COMPONENT VALUES AND TOLERANCES AS DEFINED IN SAE J-1773

Component	L_P	L_L	C_P
Value	45 μ H $\pm$ 10%	1.6 μ H $\pm$ 10%	40 nF $\pm$ 5%

Component values and tolerances are shown in Table I for unity transformer turns ratio.

The vehicle inlet secondary has four turns. The SAE J-1773 standard specifies the low number of turns so that vehicle charging inlets are compatible for different power levels, e.g., a Level 1 (120 V, 15 A) or Level 2 (230 V, 40 A) charge module can charge a Level 3 (25–160 kW) vehicle inlet. For Level 3 charging, copper losses can significantly dominate core losses in the charge vehicle inlet transformer. Thus, the secondary is wound with a low number of turns to minimize vehicle inlet losses and the resulting on-vehicle weight and cost [5].

The combination of the low number of transformer turns, isolation, spacing requirements, and the large air gap in the center leg of the ferrite core to ensure mechanical fit results in a relatively low magnetizing inductance. For the same reasons, the vehicle inlet has a significant leakage inductance.

The secondary circuit has an additional discrete capacitor C_P . This capacitor acts as the fourth element in an SP-LCLC resonant converter and confers on it many desirable features and characteristics for inductive charging.

B. Inductive Coupling Converter Requirements

The optimum topology for the DC-HFAC power converter driving the SAE J-1773 vehicle inlet can be deduced based on the system description and requirements [4].

1) *Unity Transformer Turns Ratio*: The rectified ac utility voltage will be in the 200–400-V range with a nominal 230-V input. The exact voltage level depends on whether or not a power-factor-correction (PFC) stage is used. The dc-link voltage will typically be in the 370–400-V range if a boost-regulated PFC stage is used. The required battery charging voltage will be in the 200–500-V range. Thus, a unity transformer turns ratio is desirable to minimize primary-side voltage and current stresses.

2) *Buck/Boost Voltage Gain*: Based on Section II-B-1 above, buck/boost operation is required to regulate over the full input and output voltage range.

3) *Current-Source Capability*: A desirable feature of the converter is that it should operate as a controlled current source. The vehicle inlet has a capacitive rather than inductive output filter stage to reduce on-vehicle cost/weight and additionally desensitize the converter to operating frequency and power level.

4) *Monotonic Power Transfer Curve Over a Wide Load Range*: Wide load operation is required for the likely battery voltage range. The output power should be monotonic with the controlling variable over the range.

5) *Throttling Capability Down to No Load*: All battery technologies require a wide charging current range, varying from a rated power charge to a trickle charge for battery equalization.

6) *High-Frequency Operation*: High-frequency operation is required to minimize vehicle inlet weight and cost.

7) *Full-Load Operation at Minimum Frequency for Variable-Frequency Control*: Optimizing transformer operation over the load range suggests that full-load operation takes place at low frequencies.

8) *Narrow Frequency Range*: The charger should operate over as narrow a frequency range as possible for two principal reasons. Firstly, the passive components can be optimized and secondly, operation into the AM radio band is avoided as electromagnetic emissions in this region must be tightly controlled.

9) *Soft Switching*: Soft switching of the converter stage is required to minimize semiconductor switching loss for high-frequency operation. Zero-voltage switching of a power MOSFET with its slow integral diode can result in efficiency and cost advantages. Similarly, soft recovery of the output rectifiers results in reduced power loss and EMC benefits.

10) *High Efficiency*: The power transfer must be highly efficient to minimize heat loss and to maximize the overall fuel economy of the electric vehicle charging and driving cycle.

11) *Secondary dv/dt Control*: Relatively slow dv/dt 's on the cable and secondary result in reduced high-frequency harmonics and less parasitic ringing in the cable and secondary, minimizing electromagnetic emissions.

12) *Continuous Conduction Mode Quasi-Sinusoidal Current Waveforms*: Quasi-sinusoidal current waveforms over the load range result in reduced high-frequency harmonics and less parasitic ringing in the cable and secondary, minimizing electromagnetic emissions.

III. CONVERTER DESCRIPTION AND OPERATION

The full-bridge SP-LCLC resonant dc-dc power converter with simplified inductive coupler and capacitive output filter is shown in Fig. 3. The vehicle inlet leakage inductance L_l is significantly less than the series inductance L_S and can be included with L_S , thus utilizing the parasitic element in the topology and avoiding the possible detrimental effects which transformer leakage inductance can have on many other topologies.

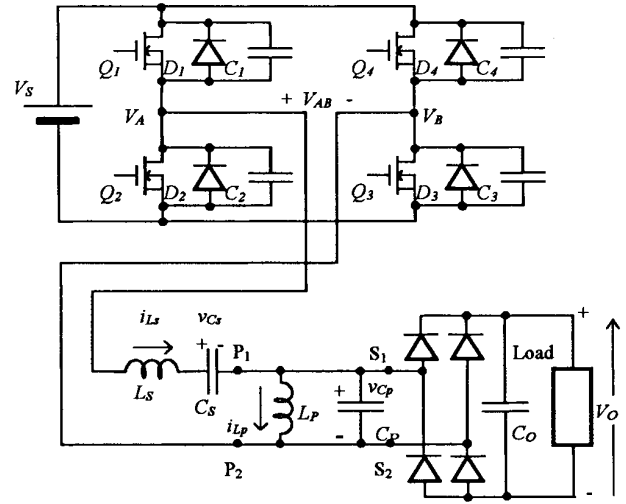


Fig. 3. Full-bridge SP-LCLC dc-dc converter.

The full bridge consists of the controlled switches Q_1, Q_2, Q_3 , and Q_4 , their intrinsic antiparallel diodes D_1, D_2, D_3 , and D_4 , and the snubber capacitors C_1, C_2, C_3 , and C_4 , to implement zero-voltage switching. The resonant network consists of two separate resonant tank circuits, the L_S - C_S series tank and the L_P - C_P parallel tank, as defined in the SAE J-1773 standard. Switches Q_1 and Q_3 are gated together in a complementary fashion to switches Q_2 and Q_4 .

The converter operates in a superresonant mode and regulates power to the battery by increasing the operating frequency to reduce output current for a given voltage.

Several principal state-plane trajectories exist, each of which consists of a sequence of characteristic modes. However, only two trajectories need to be considered for this four-element topology, operating in superresonant variable-frequency mode. The voltage and current waveforms in the resonant tank for these two trajectories, designated Trajectories 1 and 2, are shown in Figs. 4 and 5, respectively, for operation above the resonant frequency where the switches can transition at zero voltage. Trajectory 2 occurs at the higher frequencies and lower load conditions.

The two sets of trajectory waveforms show similar characteristics. The switches experience zero-voltage transitions for both turn-on and turn-off. Both trajectory waveforms show that the switches turn on at zero voltage as the inverse diode is conducting. At the turn-off instant, t_3 , in each case, a significant current flows in the switch allowing the snubber and parasitic switch capacitance to be resonantly charged and discharged, resulting in a zero-voltage turn-off. In addition, the diode has a soft turn-off, resulting in significantly reduced reverse-recovery loss. Thus, MOSFET's with slow intrinsic diodes can be used. Hence, losses in the inverter bridge are predominantly due to on-state conduction losses.

The detailed circuit operation for Trajectory 1, as shown in Fig. 4, can be understood as follows. Mode $M1$ occurs from time t_0 to t_1 . At time t_0 , transistors Q_2 and Q_4 are gated off and Q_1 and Q_3 can be gated on after a short deadtime, but

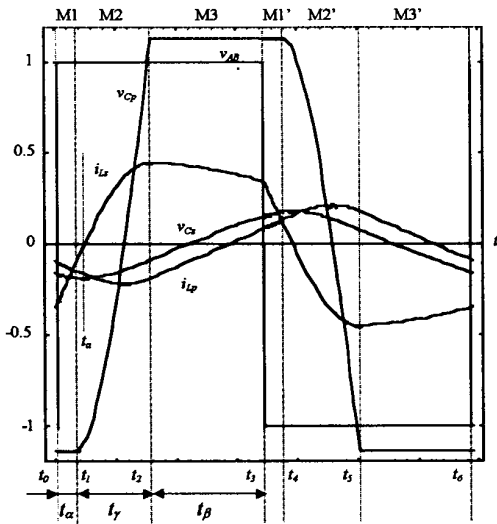


Fig. 4. Typical normalized waveforms for Trajectory 1, v_{AB} , v_{CS} , i_{LS} , v_{CP} , i_{LP} versus time.

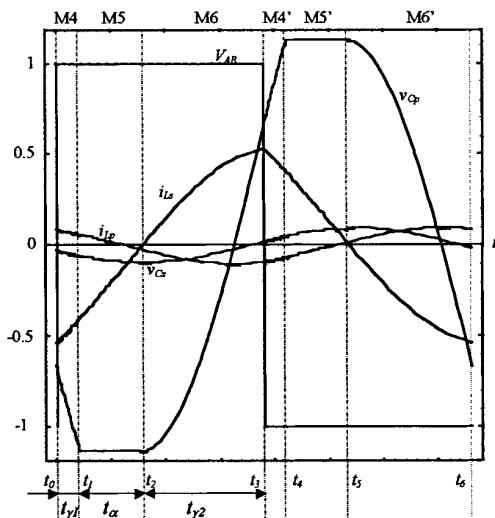


Fig. 5. Typical normalized waveforms for Trajectory 2, v_{AB} , v_{CS} , i_{LS} , v_{CP} , i_{LP} versus time.

they do not conduct because their inverse diodes D_1 and D_3 are conducting. At time t_1 , current i_{LS} equals current i_{LP} and the output rectifiers commutate with reduced reverse recovery, thus decoupling the capacitively filtered load from the resonant tank. During Mode M2, from time t_1 to t_2 , capacitor C_P is charged from $-V_O$ at t_1 to $+V_O$ at t_2 . At instant t_a during Mode M2, current i_{LS} changes polarity, commutating diodes D_1 and D_3 with reduced reverse recovery, and transistors Q_1 and Q_3 now begin to conduct. At t_2 , the output rectifiers again become forward biased and the parallel tank is once more clamped to the output voltage.

Mode M3, from t_2 to t_3 , is the main power transfer mode. At t_3 , Q_1 , and Q_3 are gated off and Q_2 and Q_4 can be gated on after a short deadtime, beginning the complementary half cycle. These devices can be gated off at zero voltage,

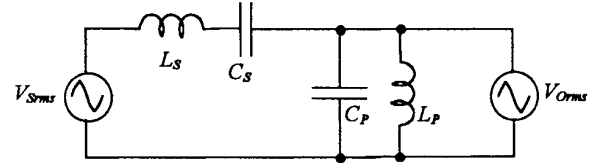


Fig. 6. Fundamental mode equivalent circuit.

as current i_{LS} is sufficient to completely charge the parallel capacitances C_1 and C_3 and discharge C_2 and C_4 prior to gating on transistors Q_2 and Q_4 . Modes $M1'$, $M2'$, and $M3'$ are the complementary modes of $M1$, $M2$, and $M3$, respectively. Clearly, in this trajectory, the switches turn on at zero voltage and zero current and turn off at zero voltage. The inverse diodes and the output rectifier diodes turn off without reverse recovery.

Similarly, the detailed circuit operation for Trajectory 2, as shown in Fig. 5, can be understood as follows. Mode M4 occurs from time t_0 to t_1 . At t_0 , transistors Q_2 and Q_4 are gated off and Q_1 and Q_3 can be gated on after a short deadtime, but they do not conduct because their inverse diodes D_1 and D_3 are conducting. During Mode M4, capacitor C_P is finally discharged and its voltage reaches $-V_O$ at t_1 . At t_1 , the output rectifiers again become forward biased and the parallel tank is once more clamped to the output voltage.

Mode M5, from t_1 to t_2 , is the only power transfer mode for Trajectory 2. At t_2 , current i_{LS} equals current i_{LP} and the output rectifiers commutate, thus decoupling the capacitively filtered load from the resonant tank. During Mode M6, from t_2 to t_3 , capacitor C_P is charged from $-V_O$. At t_3 , Q_1 and Q_3 are gated off and Q_2 and Q_4 can be gated on after a short deadtime, beginning the complementary half cycle. These devices can be gated off at zero voltage as current i_{LS} is sufficient to completely charge the parallel capacitances C_1 and C_3 and discharge C_2 and C_4 prior to gating on transistors Q_2 and Q_4 . Modes $M4'$, $M5'$, and $M6'$ are the complementary modes of $M4$, $M5$, and $M6$, respectively. As in the previous case, the switches turn on at zero voltage and zero current and turn off at zero voltage. The inverse diodes and output rectifier diodes turn off without reverse recovery.

IV. FMA

FMA [12] provides a relatively simple, but elegant, approach for the solution of complex multi-element topologies [4], [10]. An extension of this analysis is developed here to reveal the current-source behavior of the SP-LCLC converter topology. For constant output voltage applications, the FMA approach described in [12] can be extended so that both input and output ports of the resonant converter can be modeled by fundamental frequency sine-wave voltage sources, as shown in Fig. 6.

Given this approximation, the relationship between the rms values of the fundamental components representing the source and output voltages and their corresponding dc amplitudes can

be assumed to be

$$V_{S_{rms}} = \frac{2\sqrt{2}}{\pi} V_S \quad (1)$$

$$V_{O_{rms}} = \frac{2\sqrt{2}}{\pi} V_O \quad (2)$$

where V_S and V_O are the dc-bus voltage and dc-battery voltage, respectively.

Equation (2) neglects the open-circuit interval of the parallel capacitor voltage $v_{CP}(t)$ at the input to the rectifier bridge and assumes that it is a square wave. Furthermore, it is also assumed that the current $i_O(t)$ into the rectifier bridge is in phase with $v_{CP}(t)$, giving unity power factor at this point.

From this fundamental mode equivalent circuit, the relationship between the dc input voltage V_S , the dc output voltage V_O , and the dc output current I_O can be derived. This relationship, which is the approximate charging characteristic of the converter, can be shown to be the ellipse

$$V_S^2 = \frac{V_O^2}{\left(\frac{1}{1 - Z_S(\omega)Y_P(\omega)}\right)^2} + \frac{I_O^2}{\left(\frac{8}{\pi^2} \frac{1}{Z_S(\omega)}\right)^2} \quad (3)$$

where Z_S is the impedance of the series tank and Y_P is the admittance of the parallel branch. The parallel tank, composed of L_P and C_P , has a resonant frequency ω_{OP} . Above ω_{OP} , the impedance of this branch is capacitive or negative. The series tank, composed of L_S and C_S , has a natural resonant frequency ω_{OS} . Above this frequency, the series tank impedance is inductive or positive. Thus, a frequency ω_C exists, greater than ω_{OS} and ω_{OP} , at which

$$1 - Z_S(\omega_C)Y_P(\omega_C) = 0. \quad (4)$$

Equation (3) reveals clearly the special charging characteristic of the SP-LCLC converter at this particular frequency. Substituting (4) into (3) generates a new expression for the output current at the current source frequency

$$I_O(\omega_C) = \frac{V_S}{Z_S(\omega_C)} \frac{8}{\pi^2}. \quad (5)$$

At this frequency, the output voltage V_O has been eliminated from (5), and the output current magnitude is determined only by the source voltage and the series tank impedance. Clearly, at frequency ω_C , the converter acts as a current source.

Equation (5) can be used to solve for the approximate normalized converter characteristics for specified values of input voltage, output voltage, and frequency. The subscript "n" indicates that the quantity is normalized with respect to the resonant angular frequency ω_{OP} or the characteristic impedance Z_{OP} of the parallel tank. Voltages are normalized with respect to the source voltage V_S . Currents are normalized with respect to the source voltage divided by the parallel tank characteristic impedance V_S/Z_{OP} , where

$$Z_{OP} = \sqrt{\frac{L_P}{C_P}}. \quad (6)$$

The normalized dc output current versus normalized angular frequency is plotted in Fig. 7 for two values of inductance

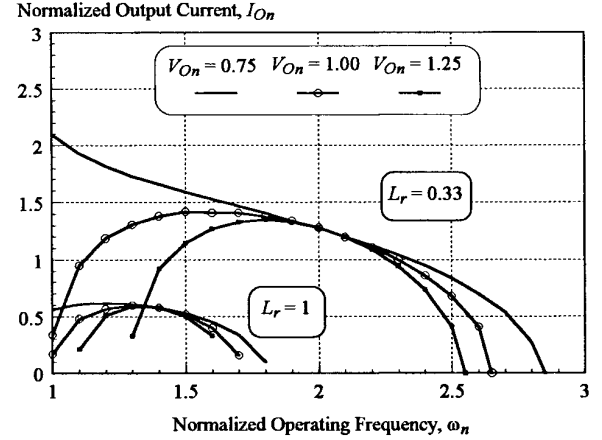


Fig. 7. Normalized output current I_{On} versus normalized operating frequency ω_n for $\omega_{OS} = 0.5\omega_{OP}$, calculated using FMA.

ratio L_r defined as L_S/L_P , and a fixed series tank natural resonant frequency $\omega_{OS} = 0.5\omega_{OP}$. For subsequent analyses, the following normalizing quantities are used: $V_S = 380$ Vdc, $Z_{OP} = 34 \Omega$, $\omega_{OP} = 2\pi f_{OP}$ where $f_{OP} = 119$ kHz is the natural resonant frequency of the parallel tank. The associated normalizing current is thus calculated to be 11.2 A.

The two sets of curves clearly show the current-source frequency where the output current is a constant for a fixed input voltage, and independent of output voltage. For the curve set with $L_r = 0.33$, the current source frequency ω_{Cn} occurs at approximately $\omega_n = 2.1$ and, for the curve set with $L_r = 1$, the frequency is at approximately $\omega_n = 1.4$.

As a first approximation, the converter can be designed using this analysis to operate in a frequency range around the current-source point where the charging current is determined solely by the dc input voltage and is independent of the output voltage at that frequency.

V. MODAL ANALYSIS

Accurate analysis of the SP-LCLC topology requires the solution of the time-domain state variable equations governing the operation of the converter. This approach is termed modal analysis. Thus, the modal analysis is more accurate, if considerably more complicated, than the FMA analysis and is used here to demonstrate the principal operating characteristics of the converter. The modal analysis is used as a design tool to accurately solve for any instantaneous, peak, maximum, rms, or average voltage or current in the topology.

Two different types of modes can be distinguished in the earlier trajectories, as shown in Column 7 of Table II. Type-A modes are defined as modes in which the capacitively filtered load is decoupled from the resonant elements and the input source voltage supplies only the four passive elements, as shown in Fig. 8.

Type-B modes are defined as modes in which the load is diode coupled to the resonant tanks and the voltage across the parallel capacitor C_P is clamped to the output voltage. Consequently, only the two series elements are resonating,

TABLE II
CHARACTERISTICS OF THE 12 MODES DEFINING TRAJECTORIES 1 AND 2

Traj.	Mode	V_s	V_o	V_{E1}	V_{E2}	Type	Time
1	M1	+	-	$+V_s+V_o$	$-V_o$	B	t_α
	M2	+	O.C.	$+V_s$	N/A	A	t_γ
	M3	+	+	$+V_s+V_o$	$+V_o$	B	t_β
	M1'	-	+	$-V_s+V_o$	$+V_o$	B	t_α
	M2'	-	O.C.	$-V_s$	N/A	A	t_γ
	M3'	-	-	$-V_s+V_o$	$-V_o$	B	t_β
2	M4	+	O.C.	$+V_s$	N/A	A	$t_{\gamma 1}$
	M5	+	-	$+V_s+V_o$	$-V_o$	B	t_α
	M6	+	O.C.	$+V_s$	N/A	A	$t_{\gamma 2}$
	M4'	-	O.C.	$-V_s$	N/A	A	$t_{\gamma 1}$
	M5'	-	+	$-V_s+V_o$	$+V_o$	B	t_α
	M6'	-	O.C.	$-V_s$	N/A	A	$t_{\gamma 2}$

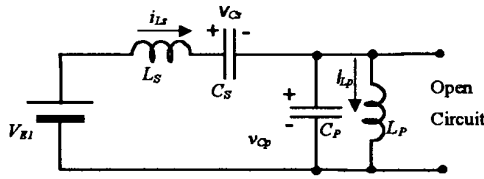


Fig. 8. Equivalent circuit for Type-A modes.

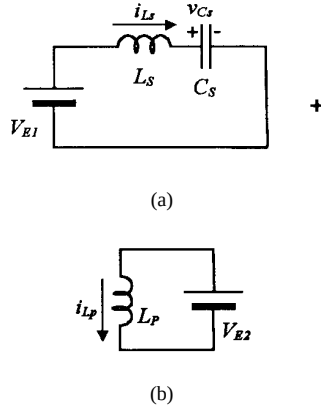


Fig. 9. Simplified decoupled subcircuits (a) and (b) for Type-B modes.

since the current in the third element i_{Lp} varies linearly, as it is in parallel with a fixed voltage source. Thus, the equivalent circuit for Type-B modes can be reduced to the two simple decoupled circuits shown in Fig. 9(a) and (b).

Trajectories 1 and 2 have a total of six modes each, as outlined in Table II. Column 3 in Table II shows the polarity of the input voltage for the particular mode. Column 4 indicates the polarity of the output voltage or if an open-circuit (O.C.) condition is present. Columns 5 and 6 define the equivalent voltage sources for the simplified equivalent circuits. Column 7 lists the mode type. Column 8 shows the time duration of each mode.

Initially, the operation of the two distinct generic resonant modes is examined, and a general solution is proposed. The resultant equations are then assigned initial and final values for each mode in the trajectory, and the solution set is

reduced to two equations in two unknowns. These equations are in transcendental form and the final solution is obtained numerically.

For this analysis, the following assumptions are made.

- 1) The output voltage is ripple free.
- 2) The switches are gated in a complementary fashion for exactly 50% of the period.
- 3) All switches and components are ideal.

A. Modal Type-A General Solution Equations

For Type-A modes, the load is decoupled from the resonant tanks as in Fig. 8. The following differential equations in matrix form give the mathematical description of the circuit model:

$$\frac{d}{dt} \begin{Bmatrix} v_{Cs}(t) \\ i_{Ls}(t) \\ v_{Cp}(t) \\ i_{Lp}(t) \end{Bmatrix} = \begin{Bmatrix} 0 & \frac{1}{C_s} & 0 & 0 \\ -\frac{1}{L_s} & 0 & -\frac{1}{L_s} & 0 \\ 0 & \frac{1}{C_p} & 0 & -\frac{1}{C_p} \\ 0 & 0 & \frac{1}{L_p} & 0 \end{Bmatrix} \begin{Bmatrix} v_{Cs}(t) \\ i_{Ls}(t) \\ v_{Cp}(t) \\ i_{Lp}(t) \end{Bmatrix} + \begin{Bmatrix} 0 \\ \frac{1}{L_s} \\ 0 \\ 0 \end{Bmatrix} V_{E1} \quad (7)$$

or

$$\frac{d\mathbf{x}(t)}{dt} = \mathbf{A_A}\mathbf{x}(t) + \mathbf{B_A}U_A \quad (8)$$

where $\mathbf{x}(t) = [v_{Cs}, i_{Ls}, v_{Cp}, i_{Lp}]^T$ is the vector representing the four coupled state variables $v_{Cs}, i_{Ls}, v_{Cp}, i_{Lp}$, as defined in (7), $\mathbf{A_A}$ is the characteristic matrix, $\mathbf{B_A}$ is the input matrix, and U_A is a scalar representing the source voltage.

The above equation represents a fourth-order system which can be solved in a systematic way by using the elegant state variable transformation matrix approach described in [7]. This technique allows the replacement of the four coupled state variables $\mathbf{x}(t)$ by two mutually decoupled state variable pairs (v_1, i_1) and (v_2, i_2) defined by the transformation $\mathbf{T}$ such that

$$\begin{Bmatrix} v_1(t) \\ i_1(t) \\ v_2(t) \\ i_2(t) \end{Bmatrix} = \begin{Bmatrix} 1 & 0 & K_1 & 0 \\ 0 & 1 & 0 & K_3 \\ 1 & 0 & K_2 & 0 \\ 0 & 1 & 0 & K_4 \end{Bmatrix} \begin{Bmatrix} v_{Cs}(t) \\ i_{Ls}(t) \\ v_{Cp}(t) \\ i_{Lp}(t) \end{Bmatrix} \quad (9)$$

or

$$\mathbf{x}_L(t) = \mathbf{T}\mathbf{x}(t) \quad (10)$$

where $\mathbf{x}_L(t) = [v_1, i_1, v_2, i_2]^T$ and K_1, K_2, K_3 , and K_4 are real constants.

A new differential equation in terms of the decoupled state variables can be defined by combining (8) and (10)

$$\frac{d\mathbf{x}_L(t)}{dt} = \mathbf{T}\mathbf{A_A}\mathbf{T}^{-1}\mathbf{x}_L(t) + \mathbf{T}\mathbf{B_A}U_A. \quad (11)$$

To achieve mutual decoupling of the selected state variable pairs, it can be shown that the parameters, K_1, K_2, K_3 , and K_4 , have the following values:

$$K_1 = \frac{1}{2} \left[1 + \frac{L_S}{L_P} - \frac{C_P}{C_S} + \sqrt{\left(1 + \frac{L_S}{L_P} - \frac{C_P}{C_S} \right)^2 + 4 \frac{C_P}{C_S}} \right] \quad (12)$$

$$K_2 = \frac{1}{2} \left[1 + \frac{L_S}{L_P} - \frac{C_P}{C_S} - \sqrt{\left(1 + \frac{L_S}{L_P} - \frac{C_P}{C_S} \right)^2 + 4 \frac{C_P}{C_S}} \right] \quad (13)$$

$$K_3 = \frac{L_P}{L_S} (1 - K_1) \quad (14)$$

$$K_4 = \frac{L_P}{L_S} (1 - K_2). \quad (15)$$

Additionally, there exist characteristic impedances Z_{01} and Z_{02} and radian resonant frequencies ω_{01} and ω_{02} corresponding to the resonant transitions of the decoupled pairs which are given by

$$Z_{0j} = \sqrt{\frac{L_S}{C_S} \left(1 + \frac{C_S}{C_P} K_j \right)} \quad (16)$$

$$\omega_{0j} = \sqrt{\frac{1}{L_S C_S} \left(1 + \frac{C_S}{C_P} K_j \right)} \quad (17)$$

for $j = 1, 2$.

Using the above equations, it can be shown that the general solution to (11) for the state variables is

$$\mathbf{x}_L(t) = \mathbf{S}_{A1}(t) \mathbf{x}_L(t_i) + \mathbf{S}_{S2}(t) U_A \quad (18)$$

where t_i is the initial time instant of the mode. The complete form of (18) is shown in (A1) in the Appendix.

From the general solution in terms of the decoupled state variables, the corresponding solution using the actual component state variable values can be found using the inverse of the transformation matrix $\mathbf{T}$

$$\mathbf{x}(t) = \mathbf{T}^{-1} \mathbf{x}_L(t). \quad (19)$$

The above general solution for Type-A modes involves four equations containing nine variables: the four initial conditions, the four final conditions, and the time interval.

B. Modal Type-B General Solution Equations

Type-B modes differ from Type-A modes, as only three passive elements have to be considered, since v_{C_P} is clamped to the output voltage. The equivalent circuit, illustrated in

Fig. 9, can be described in state variable format as shown in complete form

$$\frac{d}{dt} \begin{Bmatrix} v_{C_S}(t) \\ i_{L_S}(t) \\ v_{C_P}(t) \\ i_{L_P}(t) \end{Bmatrix} = \begin{Bmatrix} 0 & \frac{1}{C_S} & 0 & 0 \\ -\frac{1}{L_S} & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 \end{Bmatrix} \begin{Bmatrix} v_{C_S}(t) \\ i_{L_S}(t) \\ v_{C_P}(t) \\ i_{L_P}(t) \end{Bmatrix} + \begin{Bmatrix} 0 & 0 \\ \frac{1}{L_S} & 0 \\ 0 & 0 \\ 0 & \frac{1}{L_P} \end{Bmatrix} \begin{Bmatrix} V_{E1} \\ V_{E2} \end{Bmatrix} \quad (20)$$

or

$$\frac{d\mathbf{x}(t)}{dt} = \mathbf{A}_B \mathbf{x}(t) + \mathbf{B}_B U_B. \quad (21)$$

The generic solution to these state variable equations can be shown to be of the form

$$\mathbf{x}(t) = \mathbf{S}_{B1}(t) \mathbf{x}(t_i) + \mathbf{S}_{B2}(t) U_B. \quad (22)$$

The complete solution form is given in the Appendix as (A2). The above general solution to Type-B modes involves three equations containing seven variables (assuming a defined output voltage): the three initial conditions, the three final conditions and the time interval.

C. Trajectory Solutions

Trajectory 1 has 11 unknowns $V_{C_S0}, I_{L_S0}, I_{L_P0}, V_{C_S1}, I_{L_S1}, V_{C_S2}, I_{L_S2}, I_{L_P2}, t_\gamma, t_\beta$, and t_α , assuming that the input and output voltages, the switching frequency f and the four component values of the series and parallel tanks are defined. The subscripts 0, 1, and 2 designate the instantaneous value of the variable at instant t_0, t_1 , and t_2 , respectively. Additionally, in this case, there are ten mode transition equations, four Type-A and six Type-B, together with the frequency equation defined as

$$f = \frac{1}{T_s} = \frac{1}{2(t_\alpha + t_\beta + t_\gamma)}. \quad (23)$$

Thus, there are 11 equations and 11 unknowns and a unique solution exists for given component values, input voltage, output voltage, and switching frequency f . The equations have been reduced to two nonlinear equations in terms of two variables. Subsequently, the unique solution can easily be determined numerically, and waveforms as shown in Figs. 4 and 5 can be generated. Once the mode durations and the critical voltages and currents as listed above are known, then all the converter characteristics can be derived.

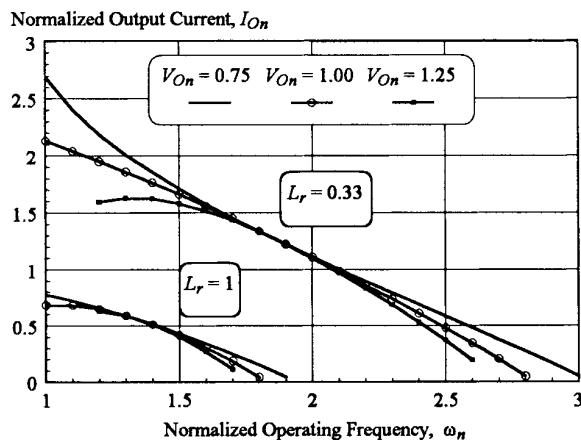


Fig. 10. Normalized output current I_{On} as a function of normalized operating frequency ω_n for $\omega_{OS} = 0.5\omega_{OP}$, calculated using modal analysis.

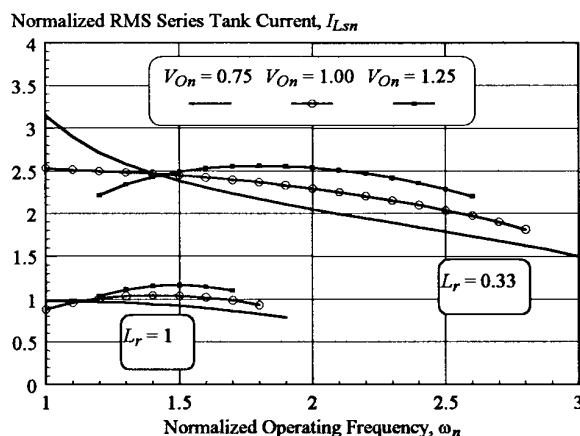


Fig. 11. Normalized rms series tank current I_{Lsn} as a function of normalized operating frequency ω_n for $\omega_{OS} = 0.5\omega_{OP}$, calculated using modal analysis.

Trajectory 2 can be shown to have 12 unknowns and 12 equations and can be solved using a similar, but more complex, method.

D. Converter Characteristics

Critical converter characteristics for two values of L_r , specifically, 0.33 and 1.0, and for a fixed series tank natural resonant frequency $\omega_{OS} = 0.5\omega_{OP}$ are shown in Figs. 10–12.

The basic current-source behavior characteristic of the converter predicted by FMA is again seen in Fig. 10. The plots indicate that, as a first approximation, the output current is constant over a range of voltages, confirming that the topology acts as a frequency-controlled current source over a significant part of the frequency range, thus rendering it ideally suitable to the inductive charging application. However, unlike the FMA characteristics shown in Fig. 7, the modal analysis predicts significantly higher output currents at high battery voltages, and at operating frequencies below the current source frequency.

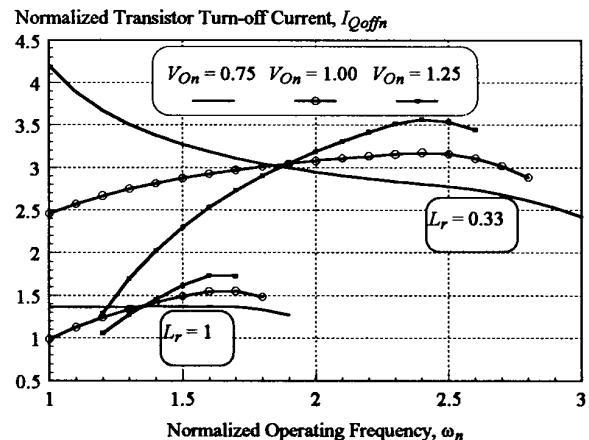


Fig. 12. Normalized transistor turn-off current I_{Qoffn} versus normalized operating frequency ω_n for $\omega_{OS} = 0.5\omega_{OP}$, calculated using modal analysis.

For the voltages shown in Fig. 10, the output power increases essentially monotonically as the frequency decreases over the frequency range. This linear large-signal transfer function characteristic of the topology greatly simplifies the design of the control loop. Additionally, the frequency range from zero to full load is limited to a reasonable value. For $L_r = 1.0$, the frequency range is just over 2:1, while for $L_r = 0.33$, the frequency range is increased to approximately 3:1.

As shown in Fig. 10, the output current and, consequently, the output power can effectively be reduced to zero by increasing the frequency. This is a critical characteristic for steady-state trickle charging of batteries.

The variation of the rms series tank current as a function of frequency for a range of normalized output voltages is shown in Fig. 11. The plots illustrate that, for the lower voltage levels, the series tank current decreases as the frequency increases and the output current decreases. However, at high voltage levels, the current in the series tank can actually increase as frequency increases, peaking in the mid-frequency range near the current-source frequency and then decreasing. This behavior can result in reduced efficiencies at lower loads for the higher voltages.

A critical parameter to enable soft switching within the converter is the turn-off current in the switch I_{Qoff} . The corresponding normalized curves are plotted in Fig. 12. The curves show that, over the required frequency range, there is a significant current in the switch at turn-off. This current magnitude is critical as sufficient current must be present to completely charge and discharge the parasitic capacitances of the commutating switches thus enabling the soft voltage transition.

VI. EXPERIMENTAL VALIDATION

A MOSFET-based prototype converter has been built and experimentally tested. Typical experimental converter waveforms for Trajectory 1 are shown in Fig. 13. These waveforms correlate closely with those predicted by theory and have the general form shown in Fig. 4. The parasitic oscillations in v_{AB}

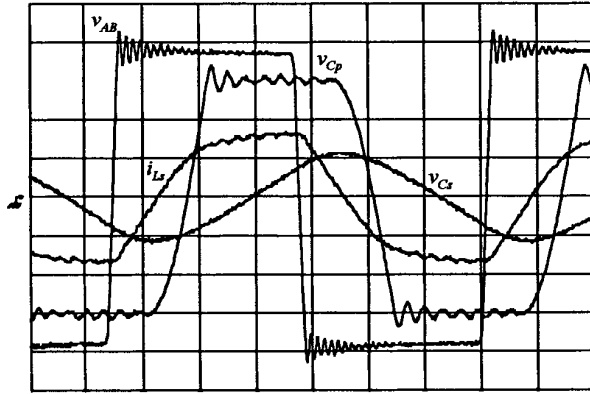


Fig. 13. Experimental waveforms for v_{AB} , v_{CS} , v_{CP} , and i_{LS} .

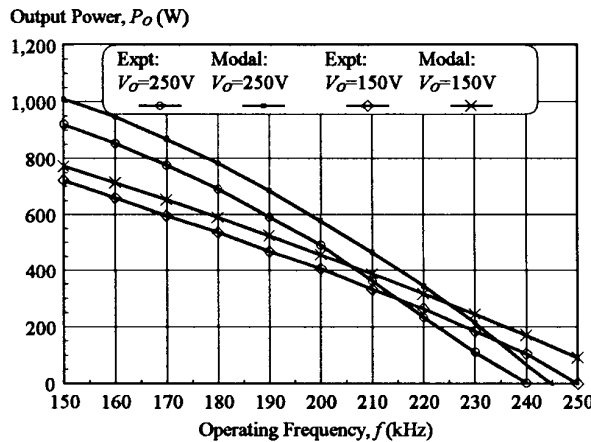


Fig. 14. Comparison of output power P_O as a function of operating frequency f for $L_r = 0.70$, $\omega_{OS} = 0.25\omega_{OP}$, and $V_S = 200$ V for experimental and modal analytical data.

and v_{CP} are independent of each other and are a function of the capacitances and parasitic inductances of the MOSFET-based H-bridge and the output rectifiers.

A comparison of experimental and analytical output power curves are shown in Figs. 14 and 15 for modal and FMA analyses, respectively. As expected, the experimental results in Fig. 14 correlate closely with the plot generated by the modal analysis. The error between the experimental data and the modal data can be accounted for by the converter inefficiency. Converter efficiency is discussed in greater detail later in this section.

Significant discrepancies can exist under some operating conditions between the results of the FMA and the experimental results, particularly for the high output voltage condition, as shown in Fig. 15. The explanation for this discrepancy is that the FMA does not account for the tristate nature of the load voltage at the rectifier input terminals. Hence, while the FMA is basically a very powerful design technique, care should be taken in its application.

A comparison of experimental and modal data for two easily measurable converter currents is shown in Fig. 16. The plots

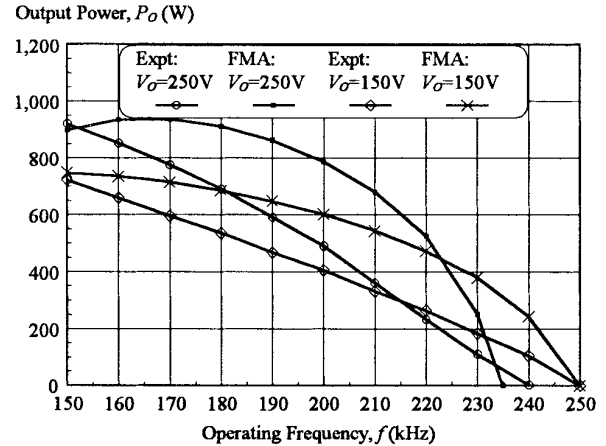


Fig. 15. Comparison of output power P_O as a function of operating frequency f for $L_r = 0.70$, $\omega_{OS} = 0.25\omega_{OP}$, and $V_S = 200$ V for experimental and modal analytical data.

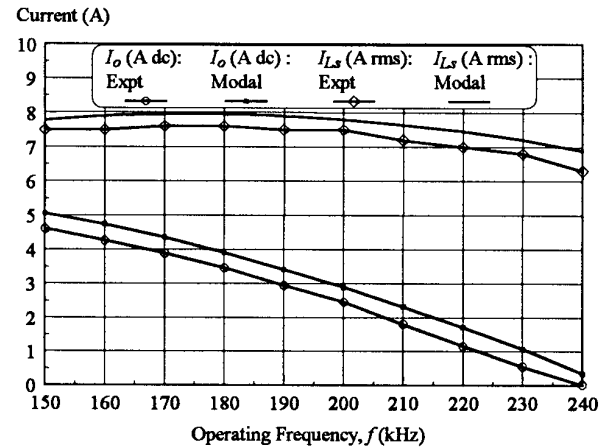


Fig. 16. Comparisons of rms series tank current I_{LS} and dc output current I_O as a function of operating frequency f for $L_r = 0.70$, $\omega_{OS} = 0.25\omega_{OP}$, $V_O = 200$ V, and $V_S = 200$ V for experimental and modal analytical data.

for both the rms series tank current I_{LS} and dc output current I_O agree well over the range, with a relatively consistent error between analysis and experimental data.

The errors between the predicted modal results and the measured experimental results are investigated in detail as follows. The modal equations are solved to generate data curves for the critical currents within the converter. These curves are shown in Figs. 17 and 18 for the following currents:

I_O	dc output current;
I_{LS}	rms series tank current;
I_{Cb}	rms bus capacitor current;
I_Q	rms MOSFET current;
I_{Qoff}	instantaneous turn-off current in MOSFET;
I_{Dq}	average current of MOSFET antiparallel diode;
I_{LP}	rms current in parallel or magnetizing inductance;
I_{CP}	rms current in parallel capacitor;

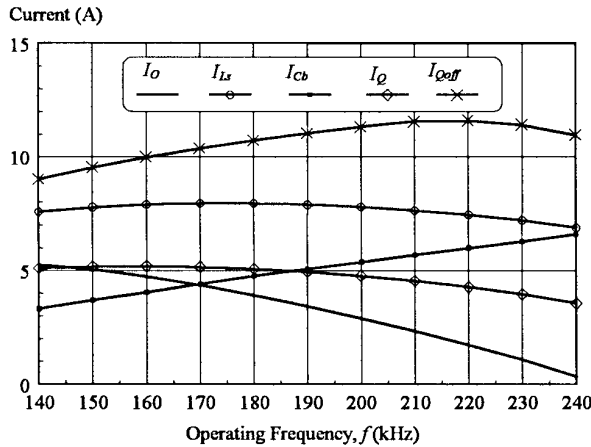


Fig. 17. Modal data sets for critical converter currents I_O , I_{Ls} , I_{Cb} , I_Q , and I_{Qoff} as a function of operating frequency f for $L_r = 0.70$, $\omega_{OS} = 0.25\omega_{OP}$, $V_O = 200$ V, and $V_S = 200$ V.

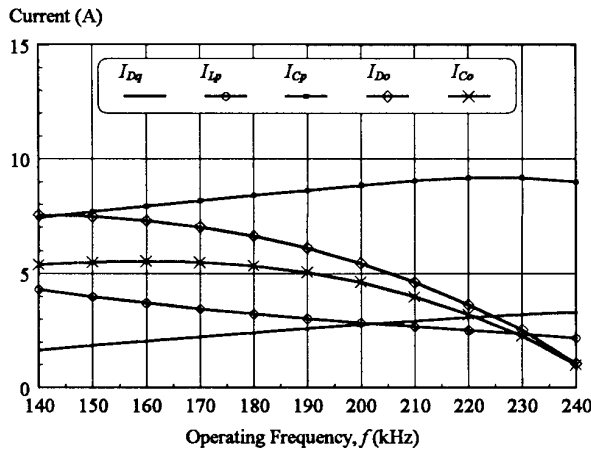


Fig. 18. Modal data sets for critical converter currents I_{Dq} , I_{Lp} , I_{Cp} , I_{Do} , and I_{Co} as a function of operating frequency f for $L_r = 0.70$, $\omega_{OS} = 0.25\omega_{OP}$, $V_O = 200$ V, and $V_S = 200$ V.

I_{Do} rms current in output rectifier diode;
 I_{Co} rms output capacitor current.

Data from current curve sets are combined with information from the typical component specifications to calculate theoretical power losses for the converter. Comparisons of calculated power loss and efficiency with experimental power loss and efficiency are shown in Fig. 19 and the curves show good correlation. The curves additionally show that most of the error between analytical and experimental results previously presented can be explained by converter inefficiency. Note that the calculated component power losses do not consider harmonic effects or second-order loss mechanisms, such as diode forward recovery.

VII. CONCLUSIONS

The inductive coupling application and its topological driving requirements have been investigated. The interface con-

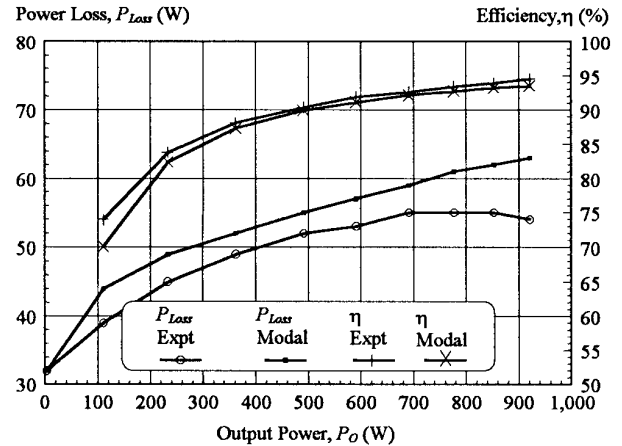


Fig. 19. Comparisons of converter power loss and efficiency as a function of output power P_O for $L_r = 0.70$, $\omega_{OS} = 0.25\omega_{OP}$, $V_O = 200$ V, and $V_S = 200$ V for experimental and modal analytical data.

tains a discrete parallel capacitor which facilitates the use of the SP-LCLC resonant topology.

The resultant four-element multiresonant topology is analyzed using the phasor-based FMA and the time-based modal approaches. The novel FMA approach presented here for battery loads provides a very simple but powerful insight into converter operation. The more complicated time-domain modal analysis is developed for the case of capacitive output filters. The describing equations are derived and solved numerically, and all circuit operating characteristics are accurately predicted. A prototype circuit was built and tested and excellent agreement was found between the modal and experimental data and a limited agreement was found between the FMA and experimental results.

Overall, the four-element topology has many desirable characteristics for use in inductive coupling battery charging, most notably, the following:

- 1) unity transformer turns ratio;
- 2) buck/boost voltage gain;
- 3) current-source capability;
- 4) monotonic power transfer curve over a wide load range;
- 5) throttling capability down to no load;
- 6) high-frequency operation;
- 7) complementary lower load/higher frequency operation;
- 8) narrow control frequency range;
- 9) zero-voltage switching;
- 10) high efficiency;
- 11) inherent short-circuit protection;
- 12) soft recovery of output rectifiers;
- 13) secondary dv/dt control;
- 14) continuous conduction mode quasi-sinusoidal current wave shaping.

The topology works well for Level 1 (120 V, 15 A) and Level 2 (230 V, 40 A) charging, as specified in the SAE J-1773 standard. The values of the series inductance and capacitance can be selected to optimize the power curves over

$$\begin{Bmatrix} v_1(t) \\ i_1(t) \\ v_2(t) \\ i_2(t) \end{Bmatrix} = \begin{Bmatrix} \cos \omega_{01}(t-t_i) & Z_{01} \sin \omega_{01}(t-t_i) & 0 & 0 \\ -\frac{1}{Z_{01}} \sin \omega_{01}(t-t_i) & \cos \omega_{01}(t-t_i) & 0 & 0 \\ 0 & 0 & \cos \omega_{02}(t-t_i) & Z_{02} \sin \omega_{02}(t-t_i) \\ 0 & 0 & -\frac{1}{Z_{02}} \sin \omega_{02}(t-t_i) & \cos \omega_{02}(t-t_i) \end{Bmatrix} \begin{Bmatrix} v_1(t_i) \\ i_1(t_i) \\ v_2(t_i) \\ i_2(t_i) \end{Bmatrix} + \begin{Bmatrix} 1 - \cos \omega_{01}(t-t_i) \\ \frac{1}{Z_{01}} \sin \omega_{01}(t-t_i) \\ 1 - \cos \omega_{02}(t-t_i) \\ \frac{1}{Z_{02}} \sin \omega_{02}(t-t_i) \end{Bmatrix} V_{E1} \quad (A1)$$

the required frequency range. The topology can also be used for Level 3 charging, but other approaches may prove more suitable, given the voltage and current magnitudes involved [4].

APPENDIX

See (A1), at the top of the page, and

$$\begin{Bmatrix} v_{Cs}(t) \\ i_{Ls}(t) \\ v_{Cp}(t) \\ i_{Lp}(t) \end{Bmatrix} = \begin{Bmatrix} \cos \omega_0(t-t_i) & Z_0 \sin \omega_0(t-t_i) & 0 & 0 \\ -\frac{1}{Z_0} \sin \omega_0(t-t_i) & \cos \omega_0(t-t_i) & 0 & 0 \\ 0 & 0 & 1 & 0 \\ 0 & 0 & 0 & 1 \end{Bmatrix} \begin{Bmatrix} v_{Cs}(t_i) \\ i_{Ls}(t_i) \\ v_{Cp}(t_i) \\ i_{Lp}(t_i) \end{Bmatrix} + \begin{Bmatrix} 1 - \cos \omega_0(t-t_i) & 0 \\ \frac{1}{Z_0} \sin \omega_0(t-t_i) & 0 \\ 0 & 0 \\ 0 & \frac{1}{L_P}(t-t_i) \end{Bmatrix} \begin{Bmatrix} V_{E1} \\ V_{E2} \end{Bmatrix} \quad (A2)$$

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