Simulation, Implementation and Validation of an On-Board Battery Charging Device for Electric Vehicles using Solely Drivetrain Components

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Abstract - Subject of this paper is the simulation, implementation and concept validation of a battery charging device for electrical vehicles, using only the on-board components of its drive chain without the need of implementing any additional electrical hardware. A description of the concept is followed by a mathematical derivation and simulation of the system behavior. An electrical drive chain including motor, power inverter and control unit were implemented on a laboratory test bench to plot the characteristic curve of the system and validate the simulation results. The closing part of this paper includes the critical interpretation of the test results, as well as an outlook on further steps of increasing the complexity of the concept.

Keywords - battery charging, electric vehicle, traction components

I. INTRODUCTION

Most battery-charging concepts in the field of electromobility are using additional external or on-board devices for rectifying and controlling charging current. These devices cause additional costs and weight, take away trunk space of the vehicle and ultimately have negative effects on their range and provided comfort. These devices usually consist of a power rectifier with integrated filter hardware followed by a dc/dc power converter to boost the input voltage and control the charging current. Now, looking at the drive chain of an electric vehicle, we find almost the same topology: The coils of the electric machine are connected to the bridges of a power inverter that usually is used to commutate the dc battery current to the three phases of the motor. Connecting a dc power source to one of the three phases, it should now be possible to drive the transistors of the power inverter in such a way, that the inductance of the motor coils and the inverters capacitor form a dc/dc power converter that can be used to charge the vehicle’s battery.

The concept of integral battery chargers was reported in [1] [2] [3]. This paper takes on a basic and more theoretical and analytical approach to the concept of the integral battery charger itself. By determining different simulation models and obtaining measurement data in a realistic physical environment, it shall help obtaining a deeper understanding of the system and the concept, and the relations of theory and practice. The results of this paper can and shall be used as a basis for further development and research.

II. LABORATORY TESTBENCH OF AN ELECTRICAL DRIVE CHAIN

To evaluate the charging concept, an electrical drivechain was implemented on a mobile rolling table shown in figure 1. The main components include a Semikron SKAI 45 A2 GD12-WDI power inverter 5, a rapid prototyping controller box from dSpace (Microautobox II) 3 and an Emrax 228 synchronous permanent magnet motor as electric machine 2. A review of different designs of electric engines and their characteristics in regard of integrated chargers can be found in [4]. Secondary components include a cooling system for the power inverter 6, as well as a simple printed circuit board (PCB) for communication signal adaptation between the inverter and control system. The input source, as well as the battery are represented by two Regatron TopCon power units. They deliver up to 32kW of power at a maximum voltage of 440 V and can be operated as source or sink. The model parameters resolve from the datasheets of the used components and are shown in table I.

![Electric drive chain implemented on a rolling table for testing](image-url)
These parameters are also used for simulation of the system to achieve results with good reference to the real behavior of the system.

### III. MATHEMATICAL MODELING AND SIMULATION

This section will elaborate a simplified model description of the underlying system. The mathematical methods used to derive the equations for system behaviour are based on Slodoban M. Cuk’s Doctoral Thesis ‘Modeling, analysis and design of switching converters’ [5], in which he delivered comprehensive mathematical modeling methods for the design of dc/dc switching converters. For this particular case, a state space model of the system is taken as basis to find the transfer function of the system for the output charging current related to the duty cycle of the inverter switches.

Analysing the behavior of switching converters, there can be differed between three different states of current flow. Images in table II display the corresponding replacement diagrams. In the first state, while the transistor closes the path between the inductance and ground, the input source drives a linear current, given a constant input voltage, and electrical Energy is stored in the magnetic field of the coil. The output current of the system is now only delivered by the Energy stored in the capacitor. These considerations lead to the following state-space for Condition I:

\[
\begin{align*}
\dot{x}_I &= \begin{bmatrix} i_L \\ u_C \end{bmatrix} = \begin{bmatrix} 0 & 0 \\ 0 & -1/R_m \end{bmatrix} \begin{bmatrix} i_L \\ u_C \end{bmatrix} + \begin{bmatrix} 1/0 \\ 0 \end{bmatrix} \begin{bmatrix} i_L \\ u_C \end{bmatrix} + \begin{bmatrix} 0 \\ u_Bat \end{bmatrix} \\
A_1 \\ B_1
\end{align*}
\]

\[
\begin{align*}
\dot{y}_I &= \begin{bmatrix} i_B \\ u_B \end{bmatrix} = \begin{bmatrix} 0 \\ 1/R_m \end{bmatrix} \begin{bmatrix} i_L \\ u_C \end{bmatrix} + \begin{bmatrix} 0 \\ 0 \end{bmatrix} \begin{bmatrix} i_L \\ u_C \end{bmatrix} + \begin{bmatrix} 0 \\ u_Bat \end{bmatrix} \\
C_1 \\ D_1
\end{align*}
\]

where \( \dot{x}_I \) is the input vector, \( \dot{y}_I \) is the output vector and \( A \) to \( D \) are the corresponding system matrices. In the second state the transistor is no longer driven by the control source, therefore input and output circuit are now connected. Because the Battery voltage is higher than the input voltage, polarity of the inductance changes, which causes it to drive a current to the output to keep its magnetic field alive. The change of output and capacitor voltage is now proportional to the current driven into the battery. The state space for Condition II follows

\[
\begin{align*}
\dot{x}_{II} &= \begin{bmatrix} \dot{t}_L \\ u_C \end{bmatrix} = \begin{bmatrix} 0 \\ -1/R_m \end{bmatrix} \begin{bmatrix} \dot{t}_L \\ u_C \end{bmatrix} + \begin{bmatrix} 1 \\ 0 \end{bmatrix} \begin{bmatrix} \dot{t}_L \\ u_C \end{bmatrix} + \begin{bmatrix} 0 \\ u_Bat \end{bmatrix} \\
A_{II} \\ B_{II}
\end{align*}
\]

\[
\begin{align*}
\dot{y}_{II} &= \begin{bmatrix} \dot{t}_L \\ u_B \end{bmatrix} = \begin{bmatrix} 0 \\ 1/R_m \end{bmatrix} \begin{bmatrix} \dot{t}_L \\ u_C \end{bmatrix} + \begin{bmatrix} 0 \\ 0 \end{bmatrix} \begin{bmatrix} \dot{t}_L \\ u_C \end{bmatrix} + \begin{bmatrix} 0 \\ u_Bat \end{bmatrix} \\
C_{II} \\ D_{II}
\end{align*}
\]

Now for further analysis, there is to differ between two operation areas of a switching converter. Depending on the duty-cycle of the transistor’s gate-signal, there will come a point during state II where all of the energy stored in the inductance was transferred to the output, which will cause a current gap in the input circle. This condition leads to state III. Its state space can be found easily by setting the first component of the state vector zero and resuming the output vector of state I:
\[
\begin{align*}
\vec{e}_{III} &= \begin{bmatrix} 0 & 0 & 1 \\ 0 & 0 & \frac{1}{R_B \cdot C} \end{bmatrix} \begin{bmatrix} i_L \\ u_C \end{bmatrix} + \begin{bmatrix} 0 & 0 & 1 \\ 0 & 0 & \frac{1}{R_B \cdot C} \end{bmatrix} \begin{bmatrix} u_V \\ u_{Bat} \end{bmatrix} \\
\vec{y}_{III} &= \begin{bmatrix} 0 & \frac{1}{R_B} & 1 \\ 0 & 0 & 0 \end{bmatrix} \begin{bmatrix} i_L \\ u_C \end{bmatrix} + \begin{bmatrix} 0 & 0 & 1 \\ 0 & 0 & \frac{1}{R_B \cdot C} \end{bmatrix} \begin{bmatrix} u_V \\ u_{Bat} \end{bmatrix}
\end{align*}
\]

(5)

The time functions of the input current in the individual states can now be derived from the state space models. According to the conditions in Table II, they follow
\begin{align*}
i_L &= \frac{1}{L} \cdot u_V \cdot t + i_L(t = 0) = \frac{1}{L} \cdot u_V \cdot t \\
i_e &= \frac{u_V - u_{Bat}}{L} \cdot t + i_L(t = t_1) = \frac{u_V - u_{Bat}}{L} \cdot t + \frac{u_V}{L} \cdot t_1,
\end{align*}

(7)

(8)

Due to the input current gap in State III we can conclude that the amount of charge delivered to the inductance in state I and the amount of charge delivered to the output by the inductance in State II must add to zero. This leads to a condition for the times of state transitions leading to
\[
\frac{1}{L} u_V \cdot t_1 = \frac{u_{Bat} - u_V}{L} \cdot t_2 \\
t_2 = \frac{u_V}{u_{Bat} - u_V} \cdot t_1
\]

(9)

Setting the transition times in relation to the periodic time of the driving signal of the transistor duty cycles can be introduced to
\begin{align*}
d_1 &= \frac{t_1}{T} \\
d_2 &= \frac{t_2 - t_1}{T}
\end{align*}

(10)

(11)

where \( T \) is the periodic time. Assuming that the amount of charge to and from the capacitor during one period is equal, the time mean value of the output current equals the integral of the inductive current over one signal period and resolves to
\[
\bar{i}_B \approx i_L = \frac{1}{T} \int_{t_1}^{t_2} i_L \, dt = \frac{u_V}{2L \cdot T} \cdot t_1 (t_2 - t_1)
\]

(12)

\[
= \frac{u_V^2}{2L \cdot (u_{Bat} - u_V)} \cdot d_1.
\]

From here the transition into the second operation area of the switching converter can be calculated. As soon as the duty cycle \( d_1 \)) increases to a critical value, there will no longer be a current gap in the input current, for state II doesn’t leave enough time for the inductance to transmit all its stored energy to the output. Accordingly state III will not be entered anymore. That is, when time \( t_2 \) equals the periodic time \( T \). Assuming an input voltage of \( V_i = 250V \) and a typical battery voltage of \( V_{Bat} = 350V \), the critical duty cycle calculates to
\[
d_{1, crit} = \frac{u_{Bat} - u_V}{u_{Bat}} \approx 0.3
\]

(13)

For duty cycles above this value, only state spaces for state I and II are describing the system. By using the method of state-space-averaging, introduced by Slobodan Cuk and R.D. Middlebrook [5], the two models can be converted into one state-space description by adding the separate input and output vectors, weighted with their according duty-cycles. The resulting state space model adds up to
\[
\bar{x} = \begin{bmatrix} \bar{i}_L \\ \bar{u}_C \end{bmatrix} = \begin{bmatrix} u_V - u_{Bat} \\ \frac{E_R - E_R}{E_R} \end{bmatrix} \cdot (d_1 - 1) + \frac{d_1 \cdot u_V}{u_{Bat}} \\
\bar{y} = \begin{bmatrix} \bar{i}_B \\ \bar{u}_B \end{bmatrix} = \begin{bmatrix} \frac{E_R - u_{Bat}}{E_R} \cdot d_1 + \frac{u_V - u_{Bat}}{u_{Bat}} \cdot (1 - d_1) \\ u_{Bat} \end{bmatrix}
\]

(14)

(15)

with \( \bar{x} \) and \( \bar{y} \) representing the mean value of the systems state space model. Considering only the steady state of the system, where all derivatives equal zero, a linear equation system evolves, allowing a simple calculation of the output currents mean value for the operation area to
\begin{align*}
\bar{i}_B &\approx \bar{i}_L = \frac{u_V^2}{2L \cdot (u_{Bat} - u_V)}.
\end{align*}

(16)

The diagram shown in figure 2 displays a Matlab plot of the two output current functions relative to the duty cycle of the transistor driving signal. The intersection point marks the transition of the two operating areas and fits to the critical duty-cycle calculated in eq. (13). The simulation parameters can be found in Table I. In the plot, the yellow line represents the behavior of the current-gapping operating area following eq. (12), while the green line refers to eq. (16), where there’s no longer a current gap during the off-time of the transistor. The intersection of the two curves matches the critical duty-cycle calculated in eq. (15) with good conformity.

\begin{figure}[h]
\centering
\includegraphics[width=0.5\textwidth]{output_current_duty_cycle.png}
\caption{Output current over duty cycle}
\end{figure}

To gain Simulation results on a higher complexity-level, a physical reference model of the laboratory testbench was
implemented in Matlab Simulink using Simscape Toolbox. The resulting current-duty cycle diagram is displayed in figure 3. Comparing the plots, it shows a decent correlation between the two simulation models.

![Graph showing output current vs duty cycle](image)

Figure 3. Simulation result for the two different operating areas

Although several additional parasitic elements were considered in the physical model, the output current shows the same behavior in both models. In both figures 2 and 3 the current crosses the 50 A mark at approx. 37% duty-cycle and the critical duty-cycle lies slightly above 30%. The physical reference model used for the simulation in Simscape is shown in figure 4.

![Diagram of physical reference model in Matlab Simscape](image)

Figure 4. Physical reference model in Matlab Simscape

IV. TEST RESULTS AND MODEL VALIDATION

To verify the simulation results, the testbench was used to investigate the real physical behavior of the system. Two power units were connected to both, the input and output of the circuit, where one was used as the power source and the other as the emulated high-voltage battery and current sink. During the test run, the duty cycle of the power inverters IGB-transistor was increased in steps of 1% until the voltage limit of 440 V was reached at the sink. The corresponding test results are shown in figure 5. The significant curve progression of the two simulation models also shows in the measurement results at the real system. The critical duty cycle matches the calculation of eq. (13) and the simulation result of the Simscape model with good correlation. The second significant point, which is the 50 A mark, is reached at 35% duty-cycle, also matching the simulation results. The measurement had to be stopped at a duty-cycle of 39% due to the voltage limit of 440 V being reached at the current sink power unit.

V. TRANSFER FUNCTION AND CONTROLLER DESIGN

In this section a controller is designed to control the charging current and output voltage by manipulating the duty-cycle of the inverters switching transistor. The following design only applies to an operating area between 0% and the critical value, above which system behavior changes magnificently. The first step now will be to find the transfer function of the system. This can be obtained through linearization and Laplace-transformation of eq. (12), the equation describes the average current delivered by the inductivity over one switching period. This relation can also be expressed through the second Taylor-polynomial \((n=2)\), calculated around its equilibrium point \(a\):

\[
T_f(x,\alpha) = \sum_{k=0}^{n} \frac{f^{(k)}(\alpha)}{n!} (x - \alpha)^n
\]  

(17)

with \(\bar{\nu}\) being a function of three variables the expression extends to the partial derivatives of the function for each variable:

\[
\bar{\nu} = f(u, \bar{u}_{\text{bat}}, \bar{d}), \quad \frac{\partial f}{\partial u} (u - \bar{u}) + \frac{\partial f}{\partial u_{\text{bat}}} (u_{\text{bat}} - \bar{u}_{\text{bat}}) + \frac{\partial f}{\partial d} (d - \bar{d})
\]  

(18)

where \(\bar{u}, \bar{u}_{\text{bat}}, \bar{d}\) are the equilibrium points of the state variables. The equilibrium point \(\bar{u}\) of the charging current itself is represented by the first order term of eq. (19).
Considering only the currents small signal change around its equilibrium point, the expression simplifies to:

\[
\tau_L = \frac{\partial f}{\partial \bar{u}_V} \bar{u}_V + \frac{\partial f}{\partial \bar{u}_{Bat}} \bar{u}_{Bat} + \frac{\partial f}{\partial \dot{d}_1} \dot{d}_1
\]  

(19)

with \( \tau_B, \bar{u}_V, \bar{u}_{Bat} \) and \( \dot{d}_1 \) representing the small signal variations around their according equilibrium points. Applying the scheme presented in eq. (19) to eq. (12) resolves in the linearized function of charging current:

\[
\tau_B = \frac{T}{2L} \left( \begin{array}{c}
2 \bar{u}_V \bar{u}_{Bat} - \bar{u}_V^2 \\
(\bar{u}_{Bat} - \bar{u}_V)^2 \end{array} \right) \left( \begin{array}{c}
\dot{d}_1 \\
\dot{u}_{Bat} - \bar{u}_V \end{array} \right)
\]  

(20)

Assuming the supply voltage is constant, we can state \( \bar{u}_V = 0 \) and therefore obtain the linearized battery current, described by the small signal variations of the duty-cycle and the change of battery voltage.

\[
\tau_L = \frac{T}{2L} \left( \begin{array}{c}
\dot{u}_{Bat} \dot{d}_1 \\
(\bar{u}_{Bat} - \bar{u}_V)^2 \\
\end{array} \right) \left( \begin{array}{c}
\dot{d}_1 \\
\dot{u}_{Bat} - \bar{u}_V \\
\end{array} \right)
\]

(21)

Calculating the charging current in section 3.3, we applied the simplification, that the current drawn and delivered by the capacitor over one switching period is adding to zero and therefore equals the average current delivered by the inductance. For the following considerations however, we want to take a closer look at the system's output circuit shown in figure (6). The output circuit is forming a first order filter.

![Diagram of output filter circuit](image)

Solving the differential equation and transforming it into Laplace domain leads to the following relation between the state variables:

\[
U_B(s) = G_{o}(s)I_C(s) + G_{o}(s)U_{Bat}(s) = \frac{R_B}{R_B + G \cdot s + 1} I_C(s) + \frac{1}{R_B G \cdot s + 1} U_{Bat}(s)
\]  

(22)

where \( G_{o}(s) \) is the output transfer function, while \( G_{o}(s) \) can be interpreted as the disturbance transfer function.

### Table III

<table>
<thead>
<tr>
<th>Table III List of updated system parameters</th>
</tr>
</thead>
<tbody>
<tr>
<td>updated system parameters</td>
</tr>
<tr>
<td>input Voltage ( U^i_c ) ( 35 ) ( V )</td>
</tr>
<tr>
<td>battery voltage ( U_{Bat} ) ( 49 ) ( V )</td>
</tr>
<tr>
<td>delay time ( T_d ) ( 2,5 \cdot 10^{-6} ) ( s )</td>
</tr>
</tbody>
</table>

To now obtain the system transfer function, the following considerations apply:

- all disturbances are considered zero
- only the small signal dynamics of the system are of interest, i.e. \( \dot{u}_B(s) \)

Now, by taking these considerations into credit, we can constitute \( \dot{u}_B(s) = \dot{u}_{Bat}(s) \) and obtain the system transfer function by applying eq. (21):

\[
G_S = \frac{\dot{u}_{Bat}(s)}{\dot{d}_1(s)} = \frac{k_2}{T_2 s + 1}
\]

(23)

\[
k_2 = \frac{2 \cdot (\dot{u}_{Bat} - \bar{u}_V) \cdot R_B T_d \bar{d}_1 \dot{u}_V}{2L(\bar{u}_{Bat} - \bar{u}_V)^2 + 2R_B T_d \bar{d}_1 \dot{u}_V}
\]

\[
T_2 = \frac{2R_B C L \cdot (\dot{u}_{Bat} - \bar{u}_V)^2}{2L(\bar{u}_{Bat} - \bar{u}_V)^2 + 2R_B T_d \bar{d}_1 \dot{u}_V}
\]

A comparison between the physical reference model shown in figure 4 and the analytical model obtained in this section again shows good correlation between the two approaches up to a duty-cycle of \( d_{crit} \). The parameters applied to the model remain those shown in table I, except for \( U_c \) and \( U_{Bat} \). In addition, the dead time of the system was considered. This time delay is caused by a safety function of the switching converter’s driver unit. Because of the high-power switches inertia, they are still conducting for a short amount of time after turning off the gate voltage. Therefore, to prevent short circuits, the actual driving signal is always delayed to the control signal. The new parameters are summarized in table III. Figure 7 shows the simulation results for a step-signal stimulation for both models.

Based on the analytic model, a controller is created that is valid for the operating area limited to duty-cycles between \( 0 - d_{crit} \). In a first step, the delay time \( T_d \) must be considered in the system transfer function calculated in eq. (21). Because the systems Period time exceeds the delay time significantly by a factor of 40, the delay can be modeled as a FOTD-System. The resulting system transfer function then resolves to:

\[
G_S = \frac{\dot{u}_{Bat}(s)}{\dot{d}_1(s)} = \frac{k_2}{(T_2 s + 1)(T_2 s + 1)}
\]  

(24)

The controller was designed as a PI-controller to prevent an steady-state error and decrease noise sensitivity of the system. The controller tuning was performed by applying a
tuning method introduced by German engineer Udo Kühn in an article for the journal "Automatisierungstechnische Praxis" in 1995 [6]. This method only uses two system parameters to calculate the controller parameters, the first one being the system gain $k_0$. The second parameter resolves to the sum of all time constants of the system describing FOTD-model. In this case, the controller parameters $T_H$ and $k_R$ are calculated as follows:

$$ k_R = \frac{1}{2k_0}, \quad T_H = \frac{T_2 + T_d}{2} \quad (25) $$

The resulting open loop transfer function now allows a stability check using the Nyquist-criterion. Figure 8 displays the Nyquist plot of the open loop for different duty-cycles, showing that the controlled circuit is stable in its defined operation area.

After developing a first controller design, a simulation model was implemented to the Simulink reference model. As input signal, a random trajectory was designed simulating different operating situations inside the defined operating area, as well as exceeding the duty-cycle limit. The simulation results are displayed in figure 9.

The plot shows that the system output reaches the setpoint very fast without significant overshoots or steady-state errors. Reaching an output current of 2A, the system begins to oscillate. That is by exceeding the critical duty-cycle and in this context leaving the defined operating area.

VI. SIMPLIFICATIONS AND FURTHER CONCEPT IMPROVEMENT

The system behavior reveals difficulty in regard of controller design. The discontinuous transition of the system behavior demands different controller models for the two operation areas. Operating the charging system in the first operating area leads to voltage peaks and a high current at the beginning of each switching cycle, due to the discharged inductivity in the input circuit, which leads to relatively bad efficiency, stressing of the remaining parts of the system and bad semi-behavior. The latter could already be experienced during the experiments at the test bench. Due to electromagnetic interferences, communication between the control unit and host computer crashed on a regular basis and will require further actions to make the system more robust. The efficiency curve can be calculated from the current and voltage measurements at the input and output of the circuit.
As figure 10 reveals, the average efficiency differs about 10 - 15% between the areas below and above the critical duty cycle of 30%.

A way more complex problem exists in the concept of the electric machine used to provide the inductivity for the boost-converter. Driving current through only one phase of the machine will lead to a unidirectional electromagnetic field in the stator coils. This will eventually result in a short moment of torque at the start of a charging procedure, if the permanent magnet in the rotor of the machine is not in alignment with the field of the current-carrying coil in the stator. The further development of the charging concept therefore includes the validation of a 3-phase-charging method that is designed to avoid any torque generation in the electric machine. This can be achieved by remaining the angle of the phase currents space phasor constant. A detailed description of this strategy can be found in [7]. The resulting space phasor of the phase currents resolves to:

\[ i_{res} \cdot e^{j\Phi} = i_{L1} + i_{L2} \cdot e^{j\frac{2\pi}{3}} + i_{L3} \cdot e^{-j\frac{2\pi}{3}} \]  

(26)

The angle \( \Phi \) of the resulting space phasor is now to be held constant, while the angle itself can take on any value. A control method for single-phase On-Board chargers using torque elimination is described in [8].

VII. CONCLUSION

This paper is presenting the simulation and validation of an on-board charging concept for electric vehicles. It provides the mathematical modeling and simulation of the system behavior in a typical hardware environment of an electric drive chain, represented by the chosen model parameters. To validate the simulation results, a test-bench was implemented to run measurements and proof the concept.

Test results showed good correlation between the mathematically predicted behavior and the physical system. The test results and occurring problems during the measurements where then analyzed and reflected to evaluate further steps of concept improvement. This paper delivers a foundation for developers and researchers of charging concepts in the field of electro-mobility.

REFERENCES