Flying Capacitor Resonant Pole Inverter with Direct Inductor Current Feedback

Sjef J. Settels* Non-member, Jorge L. Duarte† Non-member
Jeroen van Duivenbode‡ Non-member

(Manuscript received July 17, 2018, revised Oct. 18, 2018)

Industrial applications, such as semiconductor manufacturing equipment, require power amplifiers that provide high power with high precision and bandwidth. The Flying Capacitor Resonant Pole Inverter (FC RPI) provides a multilevel configuration with high switching frequencies and Zero-Voltage Switching (ZVS) across the entire operating range. However, the charge-based modulation scheme that is applied to ensure ZVS depends heavily on correct measurement of the zero-crossings of the filter inductor current. The delay incorporated in the measurement chain results in significant distortion of the output current, which deteriorates the performance of the end application. This research proposes to apply direct current feedback of the per-period average filter inductor current, measured using a high bandwidth Anisotropic Magneto-Resistive (AMR) sensor, to correct the distortion introduced in the output current. The simulation results of the complete converter and control configuration indicate a significant improvement in performance: 9 dB increase in Spurious Free Dynamic Range (SFDR) and 16 dB decrease in Total Harmonic Distortion (THD).

Keywords: current control, multilevel converters, resonant inverters, zero voltage switching

1. Introduction

Equipment used in semiconductor manufacturing applies high precision fast-moving stages for accurate positioning of wafers to fulfill tasks such as exposure, inspection or diced (11–15). The required positioning accuracy is in the nanometer range which results in stringent requirements on the output current accuracy and bandwidth of the power amplifiers driving the various types of actuators involved. Therefore, mitigating the errors produced by the amplifier results in improved performance of the end application and is the subject of this research (16–18).

Current power amplifiers used in semiconductor manufacturing equipment include a resonant pole inverter topology with variable hysteresis control (16). The incorporated soft-switching behavior results in a switching frequency range of 10…100 kHz. In order to meet the requirements for future generation semiconductor manufacturing equipment, a switching frequency of ≥100 kHz across the entire operating range is proposed. Due to the stringent requirements on output current accuracy for high precision mechatronic systems, the required switching frequency of the power amplifier is several orders higher than applied in common inverter applications, such as electrical propulsion and grid connected converters. This imposes significant challenges for the design and control of the power amplifier.

Previous research conducted in this area has shown that the Flying Capacitor Resonant Pole Inverter (FC RPI) is a suitable multilevel topology with increased switching frequency and Zero-Voltage Switching (ZVS) across the entire operating range (17–19), see Fig. 1. This enables the use of fast switching devices with a voltage rating lower than the bus voltage by dividing the voltage stress over multiple switches. The addition of two switches and a flying capacitor \( C_f \) to a standard half bridge results in a 3-level converter (19).

To ensure ZVS across the entire operating range, a charge-based modulation scheme, as proposed in (11) and implemented for a 3-level FC RPI in (9), is applied. However, this modulation scheme depends heavily on the correct measuring of the zero-crossings of the filter inductor current to synchronize the calculated switching times with the actual filter
inductor current. The delay incorporated in the measurement chain results in significant distortion of the output current which deteriorates the performance of the end application.

This research proposes to apply direct current feedback of the per-period average filter inductor current. An Anisotropic Magneto-Resistive (AMR) sensor with a bandwidth of around 2 MHz is used as a measurement device. The measured filter inductor current is fed to a controller in order to correct the introduced distortion of the output current. A simulation framework has been constructed incorporating the control systems as well as the converter implementation. The performance of the resulting system is verified and compared to the original system.

In Sect. 2, a brief introduction is provided regarding the Flying Capacitor Resonant Pole Inverter topology and the implemented charge-based ZVS modulation scheme. The applied elaborate control strategy with direct filter inductor current feedback is presented in Sect. 3, and the resulting simulation results of the combined control and converter framework are given in Sect. 4. The drawn conclusions are finally presented in Sect. 5.

2. Flying Capacitor Resonant Pole Inverter

2.1 Trapezoidal Filter Current A 3-level Flying Capacitor Resonant Pole Inverter (FC RPI) contains a single flying capacitor \( C_x \), see Fig. 1, resulting in three possible voltage levels of the switch-node \( v_{sn} \) (i.e. \( V_{dc} \) and \( -V_{dc} \)), and limits the voltage stress on each switch \( S_i \) to \( V_{dc} \). When actively applying this additional voltage level, a trapezoidal filter current shape is obtained instead of a triangular shape, which results in decreased rms current through the filter inductor.

A schematic representation of the filter inductor current \( i_{L_f} \) for 3-level modulation is shown in Fig. 3. For the steep gradients \( \alpha \) and \( \beta \), voltage levels \( V_{dc} \) and \( -V_{dc} \) are selected for \( v_{sn} \) respectively. For the middle part of the trapezoidal shape with gradient \( \gamma \), the voltage level depends on the voltage across the flying capacitor \( v_{cx} \), and whether the flying capacitor has to be charged \( (v_{sn} = V_{dc} - v_{cx}) \) or discharged \( (v_{sn} = -V_{dc} + v_{cx}) \). This enables the voltage across the flying capacitor \( v_{cx} \) to be actively regulated to \( V_{dc} \), assuring the availability of the \( \approx 0 \) V voltage level of the switch-node \( v_{sn} \). The corresponding equations for the slopes are:

\[
\alpha = \frac{V_{dc} - v_{out}}{L_f} \quad \beta = \frac{-V_{dc} - v_{out}}{L_f} \quad \gamma = \frac{v_{sn} - v_{out}}{L_f}
\]  

Assuming \( v_{cx} \) is balanced properly around \( V_{dc} \) with relatively low variations, the approximation is made that \( v_{sn} \approx 0 \) V during intervals \([t_2, t_5]\) and \([t_6, t_7]\). For the specific case drawn in Fig. 3, with the desired average current \( i_{amp} > 0 \) A and \( v_{out} = 0 \) V, this results in a flat middle part \( \gamma \). Next to \( i_{L_f} \) and corresponding \( v_{sn} \), the flying capacitor voltage \( v_{cx} \) and current \( i_{cx} \) are shown together with the drive signals for switches \( S_1 \)–\( 4 \), illustrating the functioning of the modulation strategy.

The duration of each switching state is adjusted to obtain the desired average filter current \( i_{amp} \), and can be adapted to minimize the rms value of the current through the filter inductor, to regulate the switching frequency, and to guarantee soft-switching at each switching instant. The latter requires the proper calculation of the corresponding time intervals and current values, in order for \( v_{sn} \) to completely commutate to the appropriate voltage level.

Fig. 2. Outline of high precision wafer positioning system (Courtesy of ASML)

Fig. 3. Schematic representation of \( i_{L_f}, i_{cx}, v_{cx}, \) and \( v_{sn} \) waveforms for a single switching period. The corresponding drive signals for switches \( S_1 \)–\( 4 \) are shown at the top of the figure. The desired average current for this exemplary schematic representation is \( i_{amp} > 0 \) A, and \( v_{out} = 0 \) V.
2.2 Charge-based Zero-Voltage Switching

The nonlinear output capacitance $C_{ov,\text{out}}$ of a MOSFET is used as a resonant component for the circulating filter current $i_L$ and to achieve ZVS in a switching leg \cite{10}. The charge model of $C_{ov,\text{out}}(v_{ds})$ enables the use of piece-wise linear approximation of the trapezoidal shape of $i_L$, of which a schematic representation is shown in Fig. 4 for $i_{\text{amp}} > 0$ A and $v_{\text{out}} < 0$ V. The charge-based analysis presented in \cite{v} was applied to a 3-level FC RPI in \cite{2} with the latter approach being used in this research.

To minimize the rms value of $i_L$, and therefore maximize efficiency, the surface of the negative part of $i_L$ (indicated with $t_9$), needed to achieve ZVS, is to be minimized. An additional advantage is that this maximizes switching frequency. As a first step, a critical area underneath $i_L$ is defined which has to contain at least the commutation charge $Q_c$ of the output capacitance $C_{ov,\text{out}}(v_{ds})$ of a single MOSFET, assuming $v_{ds} = V_{\text{dc}}$. The critical area is indicated with $Q_c$ in Fig. 4. From this area and the slope of $i_L$ for that section $\alpha$, the corresponding minimum current $i_{L,\text{amp}}$ needed to ensure ZVS is calculated according to:

$$i_{L,\text{amp}} = \pm \frac{\sqrt{2} \cdot Q_c \cdot \alpha}{\pi} \cdot \sin(\pm \frac{\pi}{2} \cdot \frac{t}{T}) \cdot T \cdot f$$

(4)

In order to achieve ZVS for each switching instant, the charge present in the current for each of the commutation intervals, being $[t_1, t_2]$, $[t_5, t_4]$ and $[t_7, t_8]$, should be at least $2Q_c$. This is indicated in Fig. 4 with areas in different shades of gray, each representing an area corresponding to $Q_c$. The indicated switching-state timing intervals correspond to Fig. 5. Note that $t_8 = t_9$ since $[t_8, t_9] = 0$ for this modulation state ($i_{\text{amp}} > 0$ A, $v_{\text{out}} < 0$ V). Furthermore, interval $[t_5, t_4]$ is set to a fixed value for regulating $v_{\text{cx}}$, but can also be used as a tuning parameter for the switching frequency \cite{v}.

Given $\alpha$, $\beta$, $\gamma$, the required commutation charge $Q_c$, and $i_{L,\text{amp}}$, all time values $t_{4a} - t_9$ and their respective current values for the negative part of $i_L$ can be calculated. This gives a total charge area $A_N$ of $i_L$ in the time interval $[t_{4a}, t_9]$, defined as $N$. Furthermore, the per switching period average value of $i_L$ has to be equal to the setpoint current, $i_{\text{amp}}$. This gives for the total charge area $A_P$ of the positive part of $i_L$ in the time interval $[t_0, t_{3a}]$, defined as $p$:

$$A_P = A_N + i_{\text{amp}} (t_N + t_P) \cdot \sin(\pm \frac{\pi}{2} \cdot \frac{t}{T}) \cdot T \cdot f$$

(5)

However, the equations for both time and current values of $i_L$ for the charge area of the positive part are still unbounded. To oppose this in a way to have simple calculations, $i_{\text{amp}}$, as indicated in Fig. 4, is proposed to be made dependent of $i_{\text{amp}}$ and the minimum commutation current $i_{L,\text{amp}}$ for the same slope $\alpha$, and set to:

$$i_{L,\text{amp}} = -i_{L,\text{amp}} + i_{L,\text{amp}} \cdot \sin(\pm \frac{\pi}{2} \cdot \frac{t}{T}) \cdot T \cdot f$$

(6)

From the given slopes of $i_L$, defined current value $i_{L,\text{amp}}$, commutation charge $Q_c$, area $A_N$ and Eq. (5), the remaining time and current values for the positive part of $i_L$ can be calculated. This results in a complete description of $i_L$ and timing of the switching intervals for a single switching period, which can be implemented in a controller.

For a converter with a trapezoidal filter current operating in all four quadrants, corresponding piece-wise linear approximations of $i_L$ can be made. The critical area of $i_L$ with a minimum required commutation charge $Q_c$ is located differently for each modulation state, depending on the signs of $i_{\text{amp}}$ and $v_{\text{out}}$. However, the reasoning and calculations are analogous to the case discussed above. Simulation results for this modulation strategy presented in \cite{12} indicate that ZVS can be achieved for the entire operation range of the power amplifier.

Zero-crossing detection of $i_L$ is used to ensure the calculated timing model remains in phase with the actual current. This means the functioning of the modulation principle is heavily dependent on an accurate and low-latency zero-crossing detection. Significant jitter and delay occurring on the zero-detection signal arriving at the controller will have a significant impact on the performance of the converter.

3. Direct Filter Inductor Current Feedback

The charge-based ZVS modulation scheme with trapezoidal filter current as detailed in Sect. 2, relies on the accurate measurement of the zero-crossings of the filter inductor current to synchronize the switching time calculations with the actual current. An accurate measurement of the filter inductor current with a high bandwidth is therefore required. An Anisotropic Magneto-Resistive (AMR) sensor. The available datasheet of the modeled sensor, being a Sensitec CMS3050 AMR current sensor, provides measured transfer characteristics \cite{13}. From the given transfer characteristics, a simplified model of the sensor is derived, resulting in a second order Butterworth low-pass filter with a cut-off frequency of 2 MHz. A desired (steady state) average filter inductor current of $i_{\text{amp}} = 5$ A is applied as the setpoint current.

The introduced delay on the filter inductor current waveform is shown in Fig. 5. The figure shows the actual filter
inductor current $i_{Lf}$ as simulated in black, and the filtered output of the sensor in gray. The delay introduced by the sensor in this simulation is in the order of 100 ns which is significant with respect to the switching times calculated by the charge-based ZVS modulation scheme.

When the filtered $i_{Lf}$ signal is used by the controller of the converter to determine the zero-crossings of the filter inductor current, an overshoot of the current occurs, which is shown in Fig. 6. The simulated waveform of $i_{Lf}$ with ideal zero-crossing detection is shown in black, and the waveform of $i_{Lf}$ with filtered zero-crossing detection is shown in gray.

The resulting per-period average filter inductor current values are 5.000 A for the ideal and 4.779 A for the filtered zero-crossing detection. This indicates a significant deviation from the required average current $i_{avg}$, which according to previous research presented in (5) would result in a significant position error of a high precision moving stage. This will be investigated further with more extensive simulations when a sinusoidal input signal is applied (see Sect. 4).

To correct for the introduced distortion of the output current, this research proposes to implement a control strategy that directly takes the measured filter inductor current from the AMR sensor as an input. A schematic outline of the complete resulting control configuration is shown in Fig. 7. It consists of two control loops comprising filter inductor current and output current control.

### 3.1 Filter Inductor Current Control

The outline of the filter inductor current control is shown in Fig. 7 in the respective dashed box. The filter inductor current $i_{Lf}$, as resulting from the simulated FC RPI, is filtered with the modeled transfer of the AMR sensor $H_{sensor}$, which consists of a second order Butterworth low-pass filter with a cut-off frequency $\omega_c = 2 \cdot \pi \cdot 2$ Mrad/s. The resulting transfer function is given by:

$$H_{sensor} = \frac{\omega_s^2}{s^2 + 2\zeta_s \omega_s s + \omega_s^2},$$  \hspace{1cm} (7)

where $\zeta_s = \frac{1}{2} \sqrt{2}$ is the damping factor of the filter.

The resulting filtered $i_{Lf}$ is then fed into the $i_{avg}$ calc block, which determines the per-period average value of the measured filter inductor current $i_{avg}$. The per-period average current $i_{avg}$ is subtracted from the average setpoint current $i_{set}$, generating an error current that is fed to a controller $C_{iLf}$. A simple feed forward of the setpoint current completes the filter inductor current control outline.

In order to obtain error correction and noise shaping, the characteristics of an integrator are required for the controller $C_{iLf}$. However, since a significant phase delay exists due to the calculation and measurement delays in the $i_{avg}$ calc block and $H_{sensor}$ respectively, the addition of a zero to the controller around the 0 dB crossing of the controller transfer $C_{iLf}$ is required. This ensures sufficient phase margin to obtain a stable control loop. A complex pole is furthermore added to obtain significant roll-off at the intended switching frequency range of $\geq 100$ kHz. The resulting Bode plot of the filter inductor current feedback controller transfer $C_{iLf}$ is shown in Fig. 8.

![Fig. 5. Simulated waveforms of $i_{Lf}$ without sensor filter (black), and with sensor filter (gray)](image)

![Fig. 6. Simulated waveforms of $i_{Lf}$ with ideal zero-crossing detection (black), and with filtered $i_{Lf}$ measured zero-crossing detection (gray). The per-period average $i_{Lf}$ is 5.000 A and 4.779 A respectively](image)

![Fig. 7. Schematic outline of the complete control configuration, with the respective parts indicated with dashed boxes](image)
The resulting transfer of $C_{ILF}$ is given by

$$C_{ILF}(s) = \frac{\omega_p^2 s + \omega_p^2 \omega_z}{s^3 + \sqrt{2}\omega_p s^2 + \omega_p^2 s}, \quad \quad \quad \quad \quad (8)$$

where $\omega_p$ is the frequency of the complex roll-off pole at 40 kHz, and $\omega_z$ is the frequency of the zero at around 3 kHz to obtain sufficient phase margin around the 0 dB crossing of the controller transfer. The resulting 0 dB crossing of $C_{ILF}$ is around 10 kHz, an order of magnitude lower than the intended switching frequency range of $\geq 100$ kHz, indicating the open-loop bandwidth of the filter inductor current control loop.

### 3.2 Output Current Control

The proposed converter configuration is intended to drive an inductive load (actuator) in a motion control system. Therefore an output current control loop is added to the filter inductor current control loop. Output disturbances, e.g. due to EMF, are to be compensated by the controller.

The outline of the output current control loop is shown in Fig. 7 in the respective dashed box. It consists of an anti-aliasing filter $H_{AA}$, current controller $C_e$, and a direct setpoint current feed forward $i_{fset}$. The anti-aliasing filter is a second order Butterworth low-pass filter with the following transfer function:

$$H_{AA}(s) = \frac{\omega_a^2}{s^2 + 2\zeta \omega_a s + \omega_a^2}, \quad \quad \quad \quad \quad (9)$$

where $\omega_a$ is the cut-off frequency at around $2 \cdot \pi \cdot 26 \text{ krad/s}$, and $\zeta = \frac{1}{2} \sqrt{2}$ the damping factor of the filter.

The switching stage as discussed in the previous sections, generates an average filter inductor current $i_{avg}$ according to a setpoint current $i_{fset}^*$. This operation principle can be approximated by a current source when the switching frequency is significantly higher than the intended frequency range of the setpoint current, and the filter inductor current control is functioning properly. This results in a current-controlled amplifier and the objective of the current controller $C_e$ is to regulate the current through the inductive load.

Since a current-controlled current amplifier is to be obtained, the controller $C_e$ is of the PID type (5) (10). Using a generic model for the plant $P_e$, of which a schematic representation is shown in Fig. 9, the open-loop transfer of the resulting output current control loop can be tuned to a desired bandwidth. The transfer functions of the corresponding blocks are:

$$P_e(s) = \frac{i_{out}}{i_{amp}} = \frac{1}{C_l L_o s^2 + C_l R_o s + 1}, \quad \quad \quad \quad \quad (10)$$

$$C_e(s) = \frac{K_D s^2 + K_p s + K_i}{s}, \quad \quad \quad \quad \quad \quad (11)$$

where $L_o$ and $R_o$ are the load inductance and resistance respectively, and $K_D$, $K_p$ and $K_i$ are gain coefficients of the PID controller.

Applying the system parameters as defined in Table 1 and a desired open-loop bandwidth of 1 kHz, being an order of magnitude lower than the bandwidth of the filter inductor current control loop, the transfer for the current controller $C_e$ and the resulting open-loop transfer $H_{OL}$ can be obtained. The resulting gain coefficients of $C_e$ are $K_D = 0.217 \cdot 10^{-3}$, $K_p = 2.41$, and $K_i = 8.81 \cdot 10^3$. Bode plots of both transfers are displayed in Fig. 10, from which can be concluded that sufficient phase margin is achieved resulting in a stable control loop.

### 4. Simulation Results

The 3-level FC RPI configuration as shown in Fig. 1 was implemented using the PLECS blockset for Matlab Simulink. The charge-based ZVS modulation scheme with trapezoidal filter inductor current and the control configuration as shown in Fig. 7 was implemented in Matlab Simulink. The simulation parameters are given in Table 1. The load used in the

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
<th>Unit</th>
</tr>
</thead>
<tbody>
<tr>
<td>$V_{dc}$</td>
<td>250</td>
<td>V</td>
</tr>
<tr>
<td>$C_1$</td>
<td>100</td>
<td>μF</td>
</tr>
<tr>
<td>$L_i$</td>
<td>20</td>
<td>μH</td>
</tr>
<tr>
<td>$C_{f1}$</td>
<td>5</td>
<td>μF</td>
</tr>
<tr>
<td>$R_o$</td>
<td>4</td>
<td>Ω</td>
</tr>
<tr>
<td>$L_o$</td>
<td>5.5</td>
<td>mH</td>
</tr>
<tr>
<td>$i_{fset}$</td>
<td>110</td>
<td>Hz</td>
</tr>
<tr>
<td>$i_{fset}$</td>
<td>10</td>
<td>A</td>
</tr>
</tbody>
</table>
simulation model is a series connection of a resistor $R_o$ and inductor $L_o$, emulating an actuator. In order to focus on the accuracy of the generated output current, the EMF resulting from actual movement of the actuator is neglected in this research.

### 4.1 Filter Inductor Current Comparison

When applying a desired (steady state) average filter inductor current of $i_{avg} = 5\, \text{A}$ to the system, the waveforms of the filter inductor current as shown in Fig. 11 are obtained for the different configurations. The waveform for ideal sensor characteristics is plotted in black, the non-ideal configuration in dark gray, and the compensated filter inductor current with the control loop enabled in light gray, where the latter two nearly overlap. The output current control loop is disabled for these simulations.

As can be seen in the figure, the filter inductor current controller adjusts the average setpoint current $i_{avg}$ in order to compensate the error originating from the sensor characteristics. The resulting corresponding average currents $i_{avg}$ are shown in Table 2, from which can be concluded that the (steady state) error caused by the non-ideal characteristics of the current sensor is significantly reduced by the added control loop.

When compared to the ideal sensor characteristics, the switching frequency is slightly increased for the compensated configuration. If required, this effect can be mitigated by reducing the length of the middle part of the trapezoidal current shape.

#### 4.2 Spectral Analysis

For high precision motion control, non-linearity of the generated output current of the amplifier has a significant impact on position accuracy. Non-linearity is investigated by applying spectral analysis to a simulated sinusoidal load current. The input frequency is set to $f_{in} = 110\, \text{Hz}$ with an amplitude of $|i_{set}| = 10\, \text{A}$, and the simulation is run for 11 periods to obtain coherent sampling throughout the entire simulation, and ample frequency points.

The same three configurations as in the previous section are simulated. The results for the Spurious Free Dynamic Range (SFDR) and Total Harmonic Distortion (THD) calculations are summarized in Table 2. The FFT plots for the three configurations are shown in Fig. 12, Fig. 13 and Fig. 14 respectively. The sample frequency is $f_{sys} = 100\, \text{MHz}$ but the frequency range of the FFT plots is confined to $10\, \text{Hz}$…$500\, \text{kHz}$ to focus on the range of interest. The output current control loop is enabled for these simulations.

From the simulation results can be concluded that including the sensor characteristics, which cannot be avoided in an actual converter, has a significant impact on performance. However, including the proposed direct $i_{L}$ control scheme results in a significant improvement: 9 dB increased SFDR and 16 dB decreased THD.

#### 4.3 Position Accuracy

To evaluate the resulting improvement of the position error in a high precision mechatronic system by applying the proposed direct inductor current feedback, a simplified mechatronic model is used based on the system described in \cite{settels1995}. A linear motor is assumed to be moving at a constant velocity with a commutation frequency of $110\, \text{Hz}$. The resulting setpoint current for the amplifier is $110\, \text{Hz}$ with an amplitude of $|i_{set}| = 10\, \text{A}$, and the simulation is run for 11 periods to obtain coherent sampling throughout the entire simulation, and ample frequency points.

The same three configurations as in the previous section are simulated. The results for the Spurious Free Dynamic Range (SFDR) and Total Harmonic Distortion (THD) calculations are summarized in Table 2. The FFT plots for the three configurations are shown in Fig. 12, Fig. 13 and Fig. 14 respectively. The sample frequency is $f_{sys} = 100\, \text{MHz}$ but the frequency range of the FFT plots is confined to $10\, \text{Hz}$…$500\, \text{kHz}$ to focus on the range of interest. The output current control loop is enabled for these simulations.

From the simulation results can be concluded that including the sensor characteristics, which cannot be avoided in an actual converter, has a significant impact on performance. However, including the proposed direct $i_{L}$ control scheme results in a significant improvement: 9 dB increased SFDR and 16 dB decreased THD.

### Table 2. Simulation results

<table>
<thead>
<tr>
<th>Configuration</th>
<th>$i_{avg}, \text{[A]}$</th>
<th>SFDR [dB]</th>
<th>THD [dB]</th>
<th>Position error [nm]</th>
</tr>
</thead>
<tbody>
<tr>
<td>Ideal sensor</td>
<td>5.000</td>
<td>62.7</td>
<td>−61.0</td>
<td>2.58</td>
</tr>
<tr>
<td>Non-ideal sensor</td>
<td>4.779</td>
<td>47.8</td>
<td>−46.1</td>
<td>6.67</td>
</tr>
<tr>
<td>Non-ideal sensor+control</td>
<td>4.998</td>
<td>66.6</td>
<td>−61.8</td>
<td>2.04</td>
</tr>
</tbody>
</table>

\[ P_{\text{mech.error}} = \frac{x_{err}}{t_{error}} = \frac{k_m}{m s^2 + cs} \]  
where $x_{error}$ is the resulting position error, $k_m = 50\, \text{N/A}$ the motor constant, $m = 75\, \text{kg}$ the mass, and $c = 200\, \text{Ns/m}$ the damping coefficient of the moving stage.

For imaging (lithographic) applications, high frequency position errors result in blurring of the projected image.
Reducing contrast. Low frequency components of the position error are to be compensated by a position control loop. The FFT plots of the previous section have shown that the harmonics of the 110 Hz setpoint current are dominant in determining the SFDR and THD of the generated output current. To evaluate the resulting position error from the generated harmonics, the Fourier coefficients of $x_{\text{error}}$ are calculated and a root-sum-square operation is applied to the derived values. The resulting position errors $x_{\text{error, harm}}$ for the three configurations are shown in Table 2.

The increased distortion levels induced by including the characteristics of the filter inductor current sensor result in a factor 2.6 increase in position error. By applying the proposed direct filter inductor current feedback, an improvement with a factor of 3.3 can be achieved, resulting in even better performance than the configuration with ideal filter inductor current measurement. This results in a significant improvement in position accuracy for equipment operating at nanometer accuracy.

5. Conclusion

Direct filter inductor current control was applied to a Flying Capacitor Resonant Pole Inverter with charge-based Zero-Voltage Switching. An output current control loop was added to the resulting configuration regulating the current through the load in order to compensate for disturbances. The complete converter framework was simulated and its performance verified using time domain and frequency domain analysis. A simplified mechatronic model was used to estimate the achievable position accuracy increase with respect to position accuracy.

Results obtained from the simulation framework indicated that a significant performance increase can be achieved with the proposed control strategy by adding the direct filter inductor current control loop. From the performed dc analysis can be concluded that the error originating from adding the filter inductor current sensor characteristics can be significantly mitigated. The subsequent ac analysis showed that a significant reduction in introduced harmonics can be achieved: 9 dB increased SFDR and 16 dB decreased THD. Following the ac analysis, a simplified mechatronic model was applied to estimate the obtained increase in position accuracy for the end application. An improvement with a factor 3.3 in position accuracy was achieved, resulting in a significant improvement for equipment operating at nanometer accuracy.

Future work will include hardware verification and increased voltage and current specifications.

References

Flying Capacitor Resonant Pole Inverter Inductor Current Feedback (Sjef J. Settels et al.)


Sjef J. Settels (Non-member) received the M.Sc. degree in electrical engineering from the Eindhoven University of Technology, Eindhoven, the Netherlands, in 2012 with a focus on power electronics and signal processing. Following his graduation, he started working at ASML, a semiconductor lithography company, after which he rejoined the university where he is currently pursuing a PhD degree at the Electromechanics and Power Electronics group. His current research interests include multilevel topologies for high-power high-accuracy power electronics.

Jorge L. Duarte (Non-member) received the M.Sc. degree from the Federal University of Rio de Janeiro, Rio de Janeiro, Brazil, in 1980, and the Dr. Ing. degree from the Institute National Polytechnique de Lorraine, Nancy, France, in 1985. In 1989, he was appointed as a Research Engineer with Philips Lighting Central Development Laboratory. Since 1990, he has been a Member of the academic staff at the Electromechanics and Power Electronics Group, Eindhoven University of Technology, Eindhoven, The Netherlands. Since October 2000, he has been a Consultant Engineer on a regular basis at high-tech industries around Eindhoven. His teaching and research interests include modeling, simulation, and design optimization of power electronic systems.

Jeroen van Duivenbode (Non-member) received the M.Sc. degree on the topics of power electronics and avionics from Delft University, Delft, the Netherlands, in 1987. Since then, he has developed satellite electronics during employments in Toulouse, France, and Horten, Norway, up to 1998, when he joined semiconductor lithography company ASML in Veldhoven, The Netherlands. Since 2012, he has been a part time Research Fellow within the Department of Electrical Engineering, Eindhoven University of Technology, Eindhoven, The Netherlands, and was appointed as a “TU/e Fellow” in 2014. He holds a patent on vacuum high-voltage connectors and has published on the subject of power device failures in terrestrial applications due to cosmic rays. His current interests include the development of highly accurate and reliable power electronics.