Digital Power Factor Correction. Recent approaches with and without current sensor

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Introduction

- **PFC**
  - AC-to-DC converters meet power quality standards

![Diagram of PFC circuit](image)

### IEC 61000-3-2

<table>
<thead>
<tr>
<th>n</th>
<th>Class A (A rms)</th>
<th>Class B (A rms)</th>
<th>Class C (% fun.)</th>
<th>Class A (mA/W)</th>
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<td>6</td>
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<td>8&lt;(n\leq40)</td>
<td>1.84/n</td>
<td>2.76/n</td>
<td>-</td>
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</tr>
</tbody>
</table>

IEC 61000-3-2 standards for power quality.
Introduction

- Passive Valley-fill circuit
  - Low power applications

Introduction

- **DCM / CCM-DCM**
  - No active current control
  - One control loop required

\[
\begin{align*}
\langle i_g \rangle_T &= v_g \frac{D^2 T}{2L} \\
\langle i_g \rangle_T &\approx v_g \frac{dT}{2L}
\end{align*}
\]
Introduction

- **CCM**
  - Higher power
  - Lower component stress
  - Lower conduction emission

  - Analog controller

  - Digital controller
Introduction

- Benefits of digital circuit capabilities broadly discussed
  - Interaction with other circuits (e.g. multi-phase)
  - Implementation of advanced control techniques (adaptive, non-linear, predictive, multi-mode, improved stability)
- Programmability, configuration
- System identification, autotuning
- Lower number of external passive components, including sensors
- Higher performance (more variables, complexity, protection, wider input and output ranges, dynamic response)
- Reduced sensitivity to ripple, noise, temperature, aging, process variation …
Examples of digital controllers

Higher performance


Reduction of measurements


Includes a load estimator

Includes a notch filter

All harmonics below 40dB after offset compensation
Examples of digital controllers


- Sampling instants for the switch current in the middle of the switch-on time (CCM is assumed)
- Sampling instants for the switch voltage during off-time and synchronized with the estimated peak input voltage
- PI current controller derived from a given stability condition
- Input voltage estimated with the integral part of the current controller

Digitalized control law

\[ d[n] = 1 - u \cdot i_L[n] \]

- Sampling instants for the switch current in the middle of the switch-on time \((d[n-1] > 0.5)\) or switch off time \((d[n-1] < 0.5)\)
- Modification for stability at light loads \((d_{max} \text{ limited})\)
- \(\Sigma\Delta\) modulator to improve the resolution of \(u\)

Examples of digital controllers

**Higher performance**


- Feedforward maintain resistance at high frequency
- Output voltage controller affects first harmonic resistance
- Constant higher harmonic resistance

\[ d_{ff} = 1 - \frac{V_{in}}{V_o} \]

**Circuit simplification**

- Duty-cycles calculated in advance
- Switching frequency not dependent on DSP speed
- Algorithm to minimize THDi

Examples of digital controllers

System identification

Digital controllers with current sensor

- **Current acquisition**
  - Fast and high resolution ADC
    - Expensive
    - Filter required
  - One sample $i[k] = <i>$
    - Poor noise immunity
    - Accurate timing

- **Circuit simplification**
  - Only $dT$ is required to compute $<i>$
  - Advantages of concurrency

Digital controllers with current sensor

- One sample per switching period
- Noise caused by sampling delay
- Alternative edge sampling
  - Rising edge sampling (large duty ratio)
  - Falling edge sampling (small duty ratio)

*Improve noise immunity*

Digital controllers with current sensor

Circuit simplification

- No ADC device
- No low pass filter required
- Also computes input voltage using synchronizing signal
- No synchronization to measure output voltage

Digital controllers with current sensor

Circuit simplification

- No ADC device
- Adapted for CCM and DCM

Digital controllers without current sensor

- **Motivation**

![Diagram of Digital Controllers without Current Sensor]

- **Current sensor**
  - High frequency ADC
  - Inverter circuit to adequate the signal

\[ P_{R_s} = I_{in}^2 R_s \]
Digital controllers without current sensor

- CCM PFC is used high power solutions (200 – 1000 W)

- The designer has to achieve a voltage threshold drop in the current sensor at nominal power

  \[ R_s \approx 0.2 \, \Omega \]

- Table with the estimated power losses in the current sensor for different output loads and American/European grid

<table>
<thead>
<tr>
<th>( V_{in} (V) )</th>
<th>( V_o (V) )</th>
<th>( P_o (W) )</th>
<th>( I_{in} (A) )</th>
<th>( P_{Rs} (W) )</th>
</tr>
</thead>
<tbody>
<tr>
<td>120</td>
<td>400</td>
<td>320</td>
<td>2.68</td>
<td>1.44</td>
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<td>400</td>
<td>480</td>
<td>4.03</td>
<td>3.24</td>
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<td>230</td>
<td>400</td>
<td>960</td>
<td>4.19</td>
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<tr>
<td>230</td>
<td>400</td>
<td>1120</td>
<td>4.89</td>
<td>4.78</td>
</tr>
</tbody>
</table>
Digital controllers without current sensor


Similarly

\[ I'(\theta) = \frac{\sqrt{2}}{k_d} \sin(\omega t) - \sqrt{2} \left( \frac{\omega L}{R_d} \right) k_d \cos(\omega t) \]

\[ k_d = \frac{V}{V_{g,\text{rms}}} \text{ (limited by the controllable region) } \]

\[ I_{\text{rms}} = \frac{V^2}{R_d V_{g,\text{rms}}} \]

Digital controllers without current sensor

Also, similar approach using
\[
d = 1 - \frac{V_{g, \text{peak}}}{V} \sin(\omega t - \theta), \text{with } \theta \text{ as the control variable}
\]

\[
L \frac{di}{dt} = V_{g, \text{peak}} \sin(\omega t) - (1-d)V
\]

\[
L \frac{di}{dt} = V_{g, \text{peak}} \sin(\omega t) - V_{g, \text{peak}} \sin(\omega t - \theta)
\]

with \(\theta\) small, \(L \frac{di}{dt} \approx V_{g, \text{peak}} \theta \cos(\omega t)\), and \(i = \frac{V_{g, \text{peak}} \theta}{\omega L} \sin(\omega t)\)


Taking \(V_F\) and \(r_L\) into account

Artificial \(V_g\) using LUT to gain immunity under distorted input voltage


International Power Conversion and Drive Colloquium 2013 (IPCDC 2013)  Moscow, 1/24/2017
Digital controllers without current sensor

- Replace the sensor by an observer defined by an Extended Kalman Filter
  - Feedback to compensate estimation errors
  - EKF 1 obtain $v_g$ (sinusoidal waveform is assumed)
    - Analog and quantization noise are reduced
  - EKF2 defines the plant (observer) and is adjusted by comparing with $v_{out}$
    - Amplitude and ripple phase are used
  - Deadbeat control

---

Digital controllers without current sensor

- **Pre-calculated duty-cycle**
  - Duty-cycles stored in a digital memory and recalled by a counter. Synch with $v_g$
  - Good results with just 4 bits or more
  - Uncontrolled (no $v_g$ nor $v_o$ sensing) and controlled version (only $v_o$ sensing)


- Predictive algorithm with $\Delta v_g$ feedforward compensation to account for harmonics
- Modified to take into account $\Delta v_o$, $r_L$, $r_{on}$

$$d[k] = \frac{V_{o,ref} - v_g[k]}{V_{o,ref}} + \frac{(i_{ref}[k+1] - i_{ref}[k])L}{T_s} \quad d_{update}[k] = d[k] + \Delta d[k]$$

$$i_{ref}[k] = \frac{\Delta v_g}{V_{o,ref}}$$

Digital controllers without current sensor

- **Pre-calculated duty-cycle**
  - Experimental acquisition of gate-drive signal sequences for different load conditions
  - Operation in programming (with sensor) and programmed (without sensor) modes


- Same technique as in previous slide to calculate the pre-stored duty cycles
- Implementation in FPGA including ΣΔ ADCs

Digital controllers without current sensor

- Pre-calculated duty-cycle
  - Pre-calculated duty-cycles divided into three components, $d_a$, $d_b$, $d_c$
  - Single ADC for $v_o$ extracting mean value and ripple, $v_g$ comparator for synchronization
  - $1-d_a$, $1-d_i$ and $d_c$ values stored in a memory (for nominal $v_g$ and load)

$$\begin{align*}
    d_a[k] &= \frac{V_{o,av} - v_g[k]}{V_{o,av}} \\
    d_c[k] &= \frac{(i_{ref}[k+1]-i_{ref}[k])L}{T_i V_{o,av}} \\
    d_i[k] &= \frac{v_o[k]-v_g[k]}{v_o[k]} \\
    d_b[k] &= d_i[k] - d_a[k]
\end{align*}$$

A. Sánchez, A. de Castro, V.M. López, F.J. Azcondo, J. Garrido
UC proposal

- Current sensorless proposals work better with
  - Large $L$
  - Low $f_{sw}$
  - Limited $v_g$ range (frequency and amplitude)
  - Limited load range

- Affected by
  - $v_g$ distortion at different degree
  - Parasitic elements
    - $R_{on}, R_L, v_d$
    - Delays
      - ADC
      - Drivers

![Commercial analog ICs for CCM PFC controllers &
Sensorless PFC controller based on current rebuilding concept]

![Graph showing RMS input voltage vs. power]
UC proposal

- Universal solution if CCM
  - Observer inherited from the HIL concept
  - Non-linear controller
  - Non-dependence on $L$
  - Broad load range
  - Universal $v_g$ (conventional grid or avionics)
  - Full compensation of the estimation errors
  - Resistor emulator and low THDi solutions with distorted $v_g$
  - No reconfiguration or tuning is required
  - FPGA implementation
Design framework

- Schematic of the development hardware

Test bench
1. Behavioral model of the power converter
   1. Inductor model
2. PFC controller and ADC (for synthesis)
   1. ADC for \( v_g \)
   2. ADC for \( v_o \)
   3. Current estimator (observer)
4. NLC for the PFC
5. Voltage controller
3. Behavioral model analog part of the ADC

Precedent “hardware in the loop”
Current estimation

- Input current estimated with $v_g$ and $v_o$.
- $Clk$ determines the calculation speed.

\[ i_L(k+1) = i_L(k) + \frac{v_{in}}{L} \Delta t \]
\[ i_L(k+1) = i_L(k) + \frac{v_{in} - v_o}{L} \Delta t \]

Non-linear current control

- Control no lineal (NLC) Peak-current nonlinear carrier control.
- $V_m$ se obtiene del lazo de tensión.
- El resultado es corrección del factor de potencia en CCM.

$V_m - V_m \frac{t}{T_s} = r_s i_L$

Boost

$V_m (1-d) = r_s i_{Lpk}$

$V_o = \frac{|v_g|}{1-d}$

$i_{Lpk} \approx |v_g|$

$V_m - V_m \frac{t}{T_s} = r_s i_L \frac{t}{T_s}$

SEPIC

$V_m \frac{(1-d)}{d} = r_s i_{Lpk}$

$V_o = \frac{|v_g|}{1-d}$

$i_{Lpk} \approx |v_g|$

Different references by Prof. S. Cuk on OCC, by Profs. R. Erickson, D. Maksimovic & R. Zane on NLC, by K. Smedley on CCM - PFC and implementations by Prof. G. Spiazzi and J. Sebastián (VCCR), entre otros.
Non-linear current control

- NLC applied to averaged estimated input current
Non-linear current control

- DCM is also estimated.
- Minimum estimated current forced to zero.
- Mode_{Dig} shows that DCM is estimated.

\[
\langle i_{est} \rangle_T = \frac{i_{estp}k}{2} \left(1 - \frac{t_d}{T}\right)
\]
Simulation of the algorithm (ideal)

- Left with DCM. Right always in CCM.
- Algorithm not modified for DCM.
- ModelSim simulation (*student edition*).
Estimation errors

- **Ideal case**
Estimation errors

\[ \Delta t_{\text{off-on}} \]

\[ \Delta t_{\text{on-off}} \]

Time error
Estimation errors

\[
\Delta t_{ON}[j] = t_{ON}[j] - t_{ON}^*[j] = \Delta t_{on-off} - \Delta t_{off-on}
\]

\[
i_g^{error}[n] = \sum_{j=1}^{j=n} \frac{v_o[j]}{L} \Delta t_{ON}[j] \approx \frac{V_o}{L} \frac{\Delta t_{on}}{T} t
\]

Accumulation of the time error
Estimation errors

\[ \Delta t_{\text{on-off}} = 40 \text{ ns}, \Delta t_{\text{off-on}} = 80 \text{ ns} \]

\[ \Delta t_{\text{on-off}} = 80 \text{ ns}, \Delta t_{\text{off-on}} = 40 \text{ ns} \]

Effect of the time error
### Estimation errors

**Effect of the voltage measurement error**

\[
i_{\text{on-off}}^\text{error} = \sum_{j=1}^{n} \left( \frac{T v_{g}^\text{error}[j]}{L} \right)
\]

\[
\Delta t_{\text{on}} = 10 \text{ ns} \Rightarrow i_{g}^\text{error}[n_{u}] = 2.92 \text{ A}
\]

<table>
<thead>
<tr>
<th>Nbits</th>
<th>(i_{g}^\text{error}[n_{u}])</th>
<th>(q^{o}(\text{V/\text{bit}}))</th>
</tr>
</thead>
<tbody>
<tr>
<td>10</td>
<td>(\pm 2.94 \text{ A})</td>
<td>0.4617</td>
</tr>
<tr>
<td>11</td>
<td>(\pm 1.47 \text{ A})</td>
<td>0.2307</td>
</tr>
<tr>
<td>12</td>
<td>(\pm 0.73 \text{ A})</td>
<td>0.1153</td>
</tr>
<tr>
<td>13</td>
<td>(\pm 0.36 \text{ A})</td>
<td>0.0578</td>
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</tbody>
</table>
Estimation errors

- Comparison of voltage and time resolution

Equivalent current estimation error generated by $T_{clk} = 10$ ns and due to $N_{bits} = 10$

Minimum current estimation error due to the resolution of $\pm 0.5$ LSB
Estimation errors

- Quantization effects
Estimation errors (V)

- Effect of the error in the voltage measurement
Estimation errors

\[ i_g[j] = \frac{v_g}{L} d[j]T + \frac{v_g - v_o}{L} (1 - d[j])T \]

- Error of the inductance value

\[ i_{est}[j] = \frac{v_g}{L_{est}} d[j]T + \frac{v_g^* - v_o^*}{L_{est}} (1 - d[j])T \]

\[ i_g[j] = i_{est}[j] \frac{L_{est}}{L} \]
Estimation errors

- Error of the inductance value

\[ q = q_s = q_e \]
\[ L = 1 \text{ mH} \]
\[ L_{	ext{req}} = 1.8 \text{ mH} \]
\[ I_{	ext{ref}} / I_e = 0.555 \]
\[ PF = 1 \]

\[ q = q_s = q_e \]
\[ L = 1.5 \times 0.147 I_e \text{ mH} \]
\[ L_{	ext{req}} = 1 \text{ mH} \]
\[ PF = 0.994 \]
\[ TDHI = 8.4 \% \]
Contribution of $v_o$ ripple

\[ V_m \frac{V_{gpk} \sin(\omega t)}{V_o - \Delta V_o \sin(2\omega t)} = r_s i_{gpk} \]

$\Delta V_o/V_o = 10\%$

$\Delta V_o/V_o = 30\%$

Current observer does not account for $\Delta v_o$
Parasitic elements
Parasitic elements

\[ qV_{\text{dig}} = V_\beta \]
Summary of the effect of the errors

- **$V \cdot s$ error**

  Estimated $V \cdot s <$ actual $V \cdot s$ across the inductor

  Estimated $V \cdot s >$ actual $V \cdot s$ across the inductor
Error compensation

- Feedforward and feedback
Error compensation

- Extrapolation

\[ \Delta v_g[j] = \frac{v_g[j] - v_g[j-1]}{f_{ADCclk}} \]

\[ \Delta v_g[j] = v_g[j + 1] - v_g[j] - v_g[j - 1] \]

\[ v_g[j] = v_g[j - 1] \]
Error compensation

- Ideal case: delay and compensation are coincident
Error compensation

\[ L = 1 \text{ mH (3A)} \]
\[ V_o = 400 \text{ V} \]
\[ f_s = 73.2 \text{ kHz} \]
\[ P_o = 640 \text{ W} \]
Error compensation

- Delay compensation

- Initial

- Continuous
Feedback compensation

- $\text{Mode}_{\text{Dig}}$ = “1” if the algorithm calculates $i_{in} \leq 0$.
- $\text{Mode}_{\text{Real}}$ = “1” if DCM is reached during the switching period.

$f_s = 73.2 \text{ kHz}$
$P_o = 160 \text{ W}$
$V_g = 230 \text{ V}_{\text{rms}} (50 \text{ Hz})$
Feedback compensation

DCM detection

![Diagram of feedback compensation and DCM detection](image-url)
Feedback compensation

\[ \Delta t_{on} = 0 \text{ ns} \]
\[ T_{DCM}^{in} = T_{DCM}^{inreb} \]

\[ \Delta t_{on} = -10 \text{ ns} \]
\[ T_{DCM}^{in} > T_{DCM}^{inreb} \]

\[ \Delta t_{on} = 10 \text{ ns} \]
\[ T_{DCM}^{in} < T_{DCM}^{inreb} \]

\[ e_{DCM} = T_{DCM}^{in} - T_{DCM}^{inreb} \]
Feedback compensation

\[ e_{DCM} = T_{DCM}^{in} - T_{DCM}^{inreb} \]
Feedback compensation
Feedback compensation

\[ e_{DCM} = T^{reb}_{DCM} - T^{in}_{DCM} \]

Plant to control

\[ G_T = \frac{\partial T^g_{DCM}}{\partial v_{dig}} \approx -\frac{q}{\Gamma} \]

around the operation point \( qV_{dig} = V_\beta \)

\[ \Gamma = \frac{V_o \pi^2 f_w^2 L}{R} \left( 2 \frac{D_{\text{max}}}{M_g^2} + \frac{K}{M_g^2} \right) \]

\[ I_g = \frac{V_o^2}{RV_g} \]

\[ M_g = \frac{V_{g,\text{peak}}}{V_o} \]

\[ K = \frac{2LF_{\text{sw}}}{R} \]

Operation as a resistor emulator
Dynamic response
Low THDi
Experimental results

\[ V_o = 400 \text{ V}_{\text{dc}}, L_1 = 1 \text{mH} \quad (R_{L1} = 0.25 \Omega) \]
\[ V_g = 230 \text{ V}_{\text{rms}} \text{(50Hz)} \]

\[ f_{\text{sw}} = 144 \text{ kHz} \]
Experimental results

<table>
<thead>
<tr>
<th>Results at 230 Vrms - 400 Hz sinusoidal grid</th>
</tr>
</thead>
<tbody>
<tr>
<td>$V_g$ (V)</td>
</tr>
<tr>
<td>----------</td>
</tr>
<tr>
<td>224.8</td>
</tr>
<tr>
<td>225.7</td>
</tr>
<tr>
<td>226.7</td>
</tr>
<tr>
<td>225.7</td>
</tr>
<tr>
<td>225.7</td>
</tr>
</tbody>
</table>
Experimental results

Inductors used in the experimental results.
Left: \( L = 1 \, mH \) (RM12-3C90 core with \( R_L = 0.25 \, \Omega \)).
Right: \( L_2 = 1.5 \, mH \) (soft saturation Kool m\( \mu \) core with \( R_{L2} = 0.35 \, \Omega \)).

\( V_o = 400 \, V_{dc} \) and \( L_2 = 1.5mH \) (\( R_{L2} = 0.35 \, \Omega \)).
(a) \( V_g = 230 \, V_{rms} \) (50Hz), \( P_g = 970W \).
(b) \( V_g = 85 \, V_{rms} \) (60Hz), \( P_g = 320W \).
Experimental results

<table>
<thead>
<tr>
<th>$V_g$</th>
<th>$P_g$</th>
<th>$PF$</th>
<th>$THD_i$</th>
</tr>
</thead>
<tbody>
<tr>
<td>250</td>
<td>460 W</td>
<td>0.975</td>
<td>9.0 %</td>
</tr>
<tr>
<td></td>
<td>645 W</td>
<td>0.991</td>
<td>8.5 %</td>
</tr>
<tr>
<td></td>
<td>800 W</td>
<td>0.993</td>
<td>9.5 %</td>
</tr>
<tr>
<td></td>
<td>970 W</td>
<td>0.993</td>
<td>10.5 %</td>
</tr>
<tr>
<td>230</td>
<td>460 W</td>
<td>0.984</td>
<td>8.1 %</td>
</tr>
<tr>
<td></td>
<td>640 W</td>
<td>0.988</td>
<td>9.1 %</td>
</tr>
<tr>
<td></td>
<td>800 W</td>
<td>0.992</td>
<td>9.8 %</td>
</tr>
<tr>
<td></td>
<td>970 W</td>
<td>0.993</td>
<td>10.5 %</td>
</tr>
<tr>
<td>180</td>
<td>323 W</td>
<td>0.980</td>
<td>5.4 %</td>
</tr>
<tr>
<td></td>
<td>485 W</td>
<td>0.989</td>
<td>7.1 %</td>
</tr>
<tr>
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<td>650 W</td>
<td>0.992</td>
<td>8.6 %</td>
</tr>
<tr>
<td></td>
<td>820 W</td>
<td>0.991</td>
<td>10.5 %</td>
</tr>
<tr>
<td>120</td>
<td>497 W</td>
<td>0.996</td>
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<td>323 W</td>
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<td>9.8 %</td>
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<tr>
<td></td>
<td>497 W</td>
<td>0.996</td>
<td>9.8 %</td>
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<tr>
<td>85</td>
<td>161 W</td>
<td>0.989</td>
<td>5.0 %</td>
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<tr>
<td></td>
<td>336 W</td>
<td>0.993</td>
<td>9.0 %</td>
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Experimental results

Resistance emulator under different power levels and input voltage distortions.
\[ V_g = 230 \, V, f_{sw} = 96 \, kHz. \]
With \( THDv = 5 \% \) at (a) \( P_g = 966.7 \, W \) and (b) \( P_g = 800.6 \, W. \)
With \( THDv = 12 \% \) at (c) \( P_g = 965.1 \, W \) and (d) \( P_g = 800.9 \, W. \)

Sinusoidal behavior under different power levels and input voltage distortions.
\[ V_g = 230 \, V, f_{sw} = 96 \, kHz. \]
With \( THDv = 5 \% \) at (a) \( P_g = 964.9 \, W \) and (b) \( P_g = 799.5 \, W. \)
With \( THDv = 12 \% \) at (c) \( P_g = 963.1 \, W \) and (d) \( P_g = 800.4 \, W. \)
Experimental results

<table>
<thead>
<tr>
<th>Voltage ($V_g$) (V)</th>
<th>Power ($P_g$) (W)</th>
<th>PF</th>
<th>$THD_v$ (%)</th>
<th>$THD_i$ (%)</th>
<th>$I_1$ (A)</th>
<th>$I_3$ (A)</th>
<th>$I_5$ (A)</th>
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Future works

- To extend the Sensorless PFC controller with current rebuilding to other topologies
  - SEPIC
  - Bridgeless and bidirectional Boost PFC rectifiers
  - Three phase bidirectional converters
- Investigate the use of time high resolution modulator
- Incorporate notch filter action in the voltage loop
- Investigate other parameters to achieve the feedback compensation of estimation errors
- Increase the speed of the feedback compensation
- Investigate the use of autotuning circuits
- Improve the PF in DCM
Conclusions

- Active error compensation is required to achieve universal current sensorless controllers.
- Extrapolation of voltage data and time feedforward compensation can achieve a satisfactory power factor under limited and static input voltage and load conditions.
- Continuous feedforward compensation is required because the switching delays depend on the devices and temperature.
- Feedforward time compensation can also compensate for voltage acquisition errors.
- Feedforward time compensation is fast but coarse and cannot reach the best power factor.
- It is easy to increase the resolution of the compensation for the difference between the real and estimated $V \cdot s$ using the voltage variable instead of the time variable.
- The error between the estimation and detection of the DCM is an adequate variable to achieve the compensation of $V \cdot s$ estimation errors.
- Feedback compensation achieves satisfactory power factor at different loads and input voltage.
- Combination of feedforward and feedback compensation achieve universal PFC controller without losing much performance in transients.
- The proposal requires no modification for different PFC design operating in CCM.
Thank you!