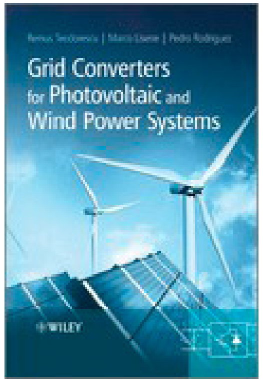

Grid Converters for Photovoltaic and Wind Power Systems

by R. Teodorescu, M. Liserre and P. Rodriguez

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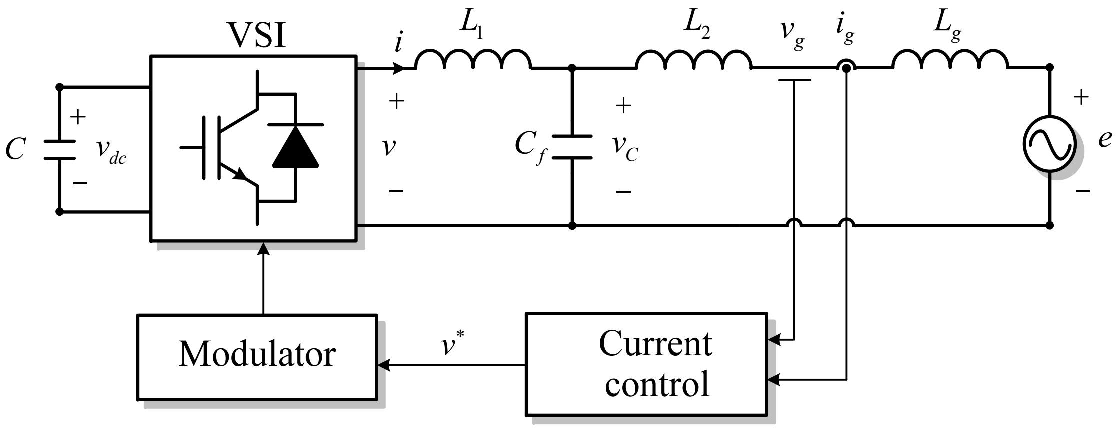
Chapter 12

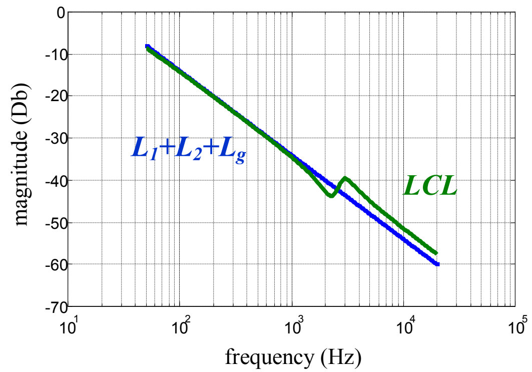
Grid Current Control

Outline

- Introduction
- Harmonic requirements
- Current control overview
- Linear current control with separated modulation
  - Use of averaging
  - PI-based control
  - Dead-beat control
  - Resonant control
  - Harmonic compensation
- Modulation techniques
  - Bipolar and unipolar modulation
  - Three-phase modulation (continuous and discontinuous)
  - Multilevel modulation
  - Interleaved converters
- Operating limits of the grid converter

Introduction





- The influence of the capacitor of the filter will be neglected since it is only dealing with the switching ripple frequencies. In fact at frequencies lower than half of the resonance frequency the LCL-filter inverter model and the L-filter inverter models have the same frequency characteristic
- PI-based current control implemented in a synchronous frame is commonly used in three-phase converters
- In single-phase converters the PI controller capability to track a sinusoidal reference is limited and Proportional Resonant (PR) can offer better performances
- Modulation has an influence on the design of the converter (dc voltage value), losses and EMC problems including leakage current

Harmonic requirements: PV-systems

- In Europe there is the standard IEC 61727
- In US there is the recommendation IEEE 929
- The recommendation IEEE 1547 is valid for all distributed resources technologies with aggregate capacity of 10 MVA or less at the point of common coupling interconnected with electrical power systems at typical primary and/or secondary distribution voltages
- All of them impose the following conditions regarding grid current harmonic content

ODD HARMONICS	***DISTORTION LIMIT***
3rd through 9th	***less than 4.0%***
11th through 15th	***less than 2.0%***
17th through 21st	***less than 1.5%***
23rd through 33rd	***less than 0.6%***

- The total THD of the grid current should not be higher than 5%

Harmonic requirements: WT-systems

- In Europe the standard 61400-21 recommends to apply the standard 61000-3-6 valid for polluting loads requiring the current THD smaller than 6-8% depending on the type of network

harmonic	limit
5^{th}	5-6 %
7^{th}	3-4 %
11^{th}	1.5-3 %
13^{th}	1-2.5 %

- In case of several WT systems

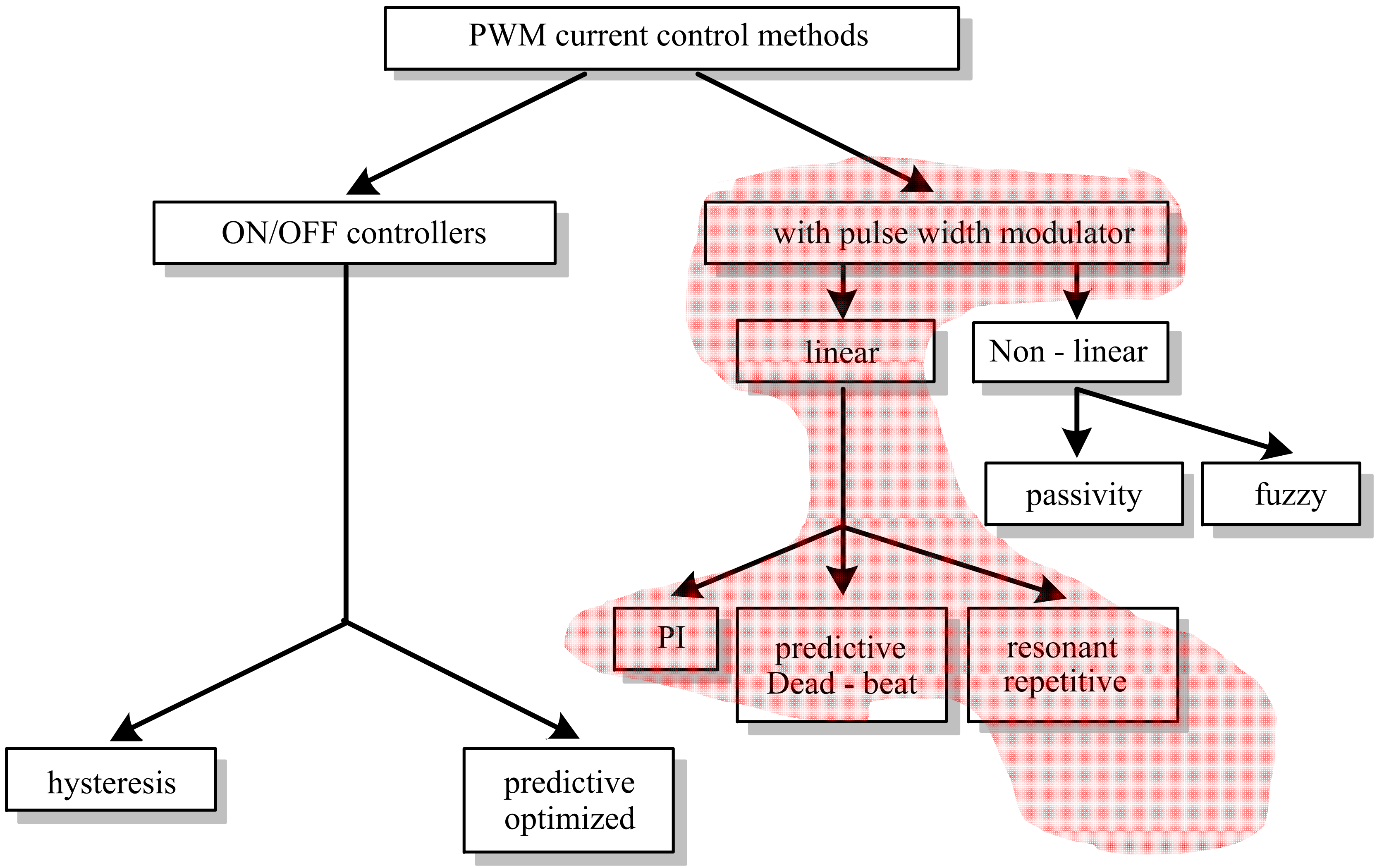
υ_i – ratio of the transformer at the ith wind turbine

$$I_{h\Sigma} = \sqrt[\beta]{\sum_{i=1}^{N} \left(\frac{I_{hi}}{\upsilon_i} \right)^{\beta}}$$

Harmonic order	Exponent β
$h < 5$	1.0
$5 \le h \le 10$	1.4
$h > 10$	2.0

- In WT systems asynchronous and synchronous generators directly connected to the grid have no limitations respect to current harmonics

Current Control overview



Linear current control with separated modulation: use of averaging

- Continuous switching vector which components are the duty cycle of each converter leg

$$\overline{d}(t)=\frac{2}{3}\left(d_a(t)+\alpha d_b(t)+\alpha^2 d_c(t)\right)$$

- Average model

$$\begin{cases}\dfrac{di_d(t)}{dt}-\omega i_q(t)=\dfrac{1}{L}\left[-Ri_d(t)-e_d(t)+v_d(t)\right]\\[2ex]\dfrac{di_q(t)}{dt}+\omega i_d(t)=\dfrac{1}{L}\left[-Ri_q(t)-e_q(t)+v_q(t)\right]\end{cases}$$

$$\begin{cases}v_d(t)=d_d(t)v_{dc}(t)\\ v_q(t)=d_q(t)v_{dc}(t)\end{cases}$$

- Linearized model $v_{dc}(t)=V_{dc}$

$$\frac{d}{dt}\begin{bmatrix}i_d(t)\\ i_q(t)\end{bmatrix}=\begin{bmatrix}-\dfrac{R}{L} & \omega\\ -\omega & -\dfrac{R}{L}\end{bmatrix}\begin{bmatrix}i_d(t)\\ i_q(t)\end{bmatrix}+\begin{bmatrix}\dfrac{V_{dc}}{L} & 0\\ 0 & \dfrac{V_{dc}}{L}\end{bmatrix}\begin{bmatrix}d_d(t)\\ d_q(t)\end{bmatrix}+\begin{bmatrix}-\dfrac{1}{L} & 0\\ 0 & -\dfrac{1}{L}\end{bmatrix}\begin{bmatrix}e_d(t)\\ e_q(t)\end{bmatrix}$$

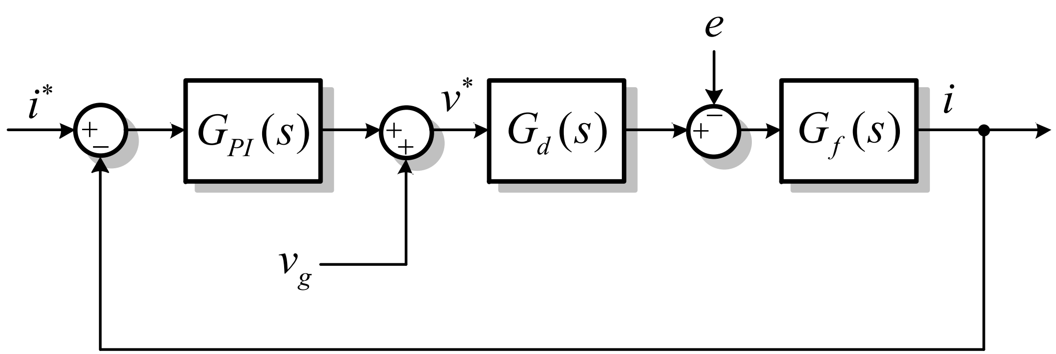
Linear current control with separated modulation: PI current control

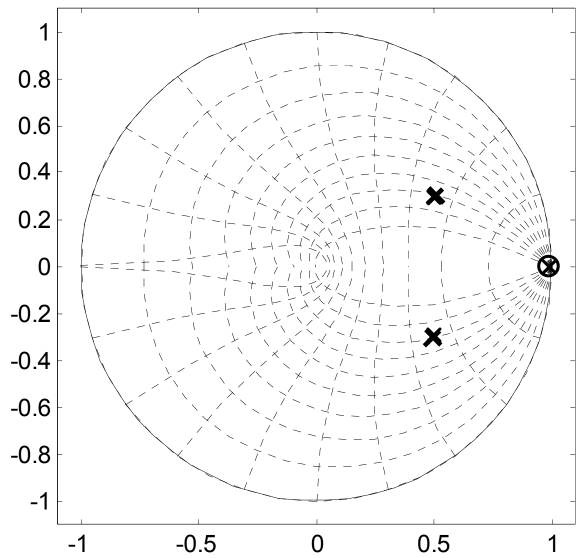
- Typically PI controllers are used for the current loop in grid inverters
- Technical optimum design (damping 0.707 overshoot 5%)

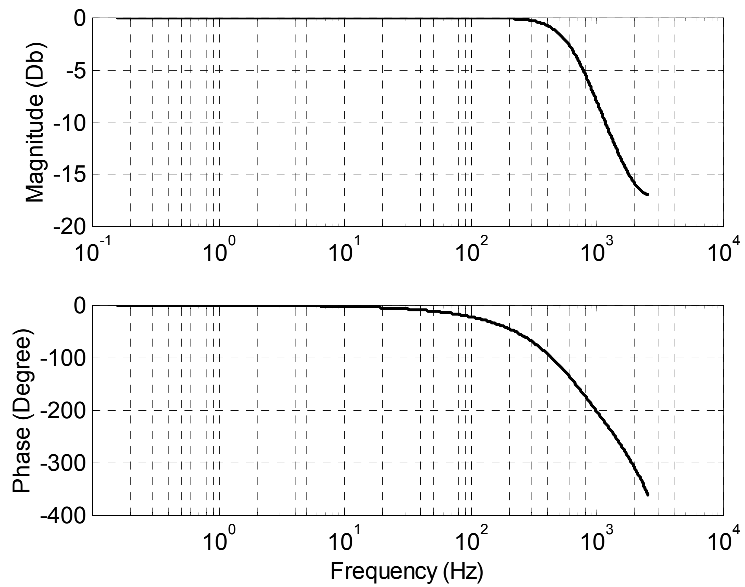
$$G_{PI}(s) = k_P + \frac{k_I}{s}$$

$$G_d(s) = \frac{1}{1 + 1.5T_s s}$$

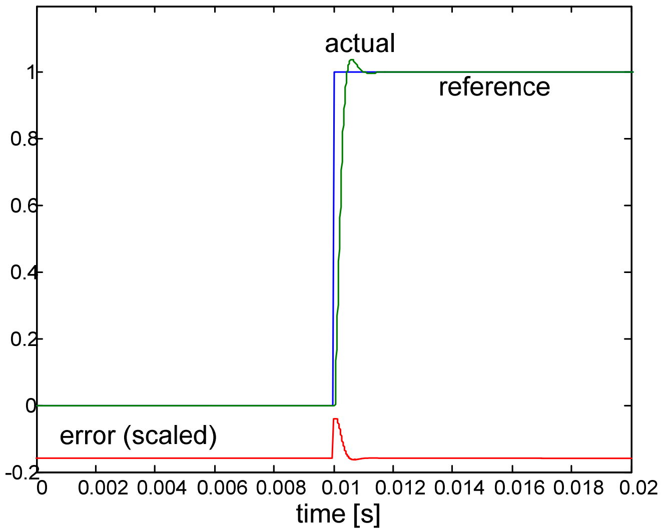
$$G_f(s) = \frac{i(s)}{v(s)} = \frac{1}{R + Ls}$$





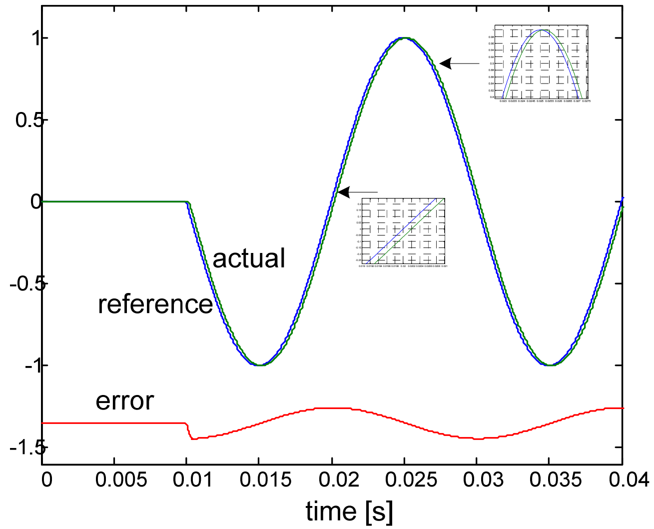


Shortcomings of PI controller



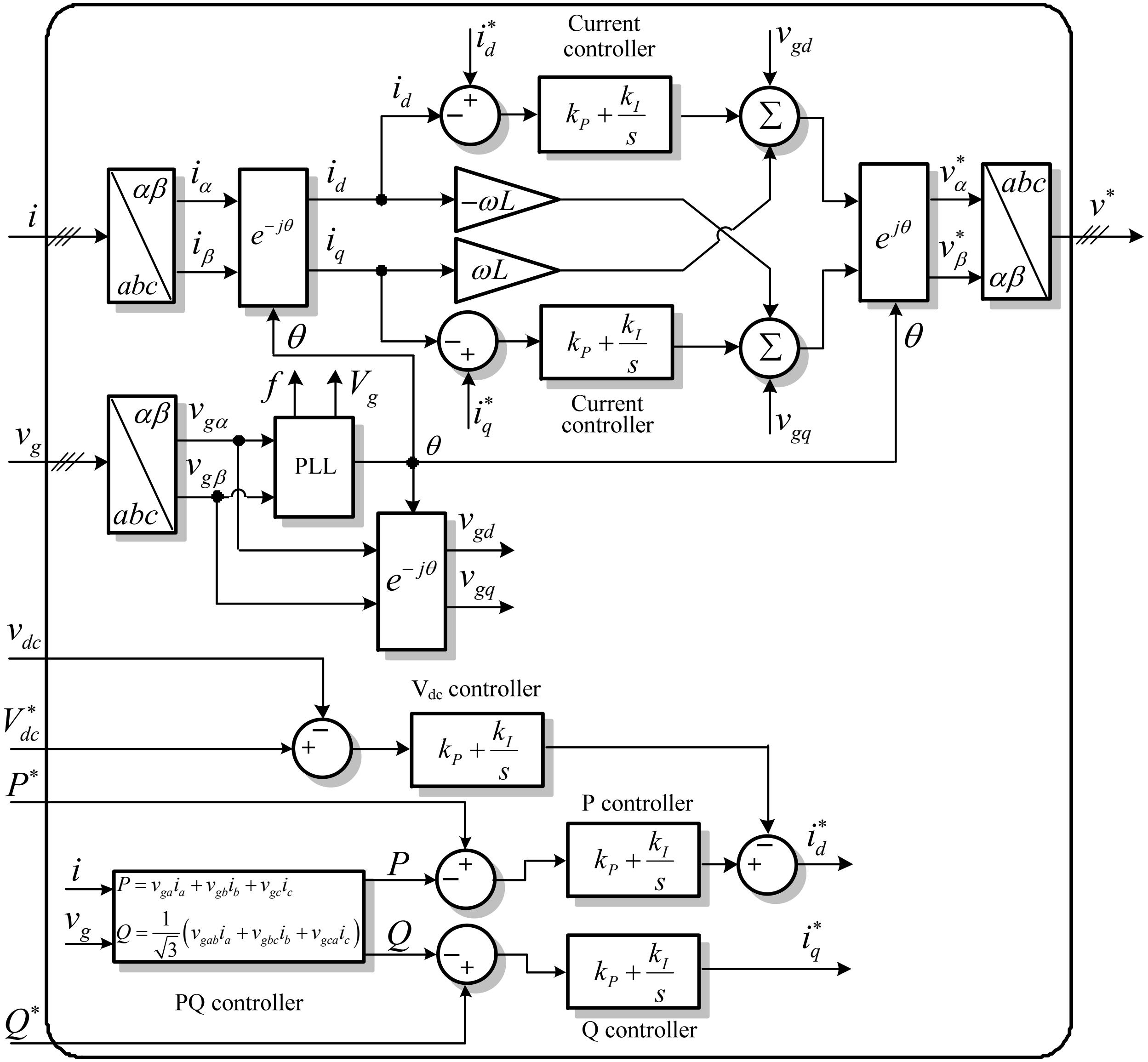
steady-state magnitude and phase error

limited disturbance rejection capability

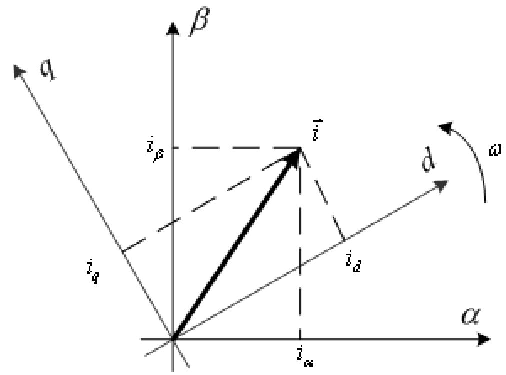


- When the current controlled inverter is connected to the grid, the phase error results in a power factor decrement and the limited disturbance rejection capability leads to the need of grid feed-forward compensation
- However the imperfect compensation action of the feed-forward control due to the background distortion results in high harmonic distortion of the current and consequently non-compliance with international power quality standards

Use of a PI controller in a rotating frame



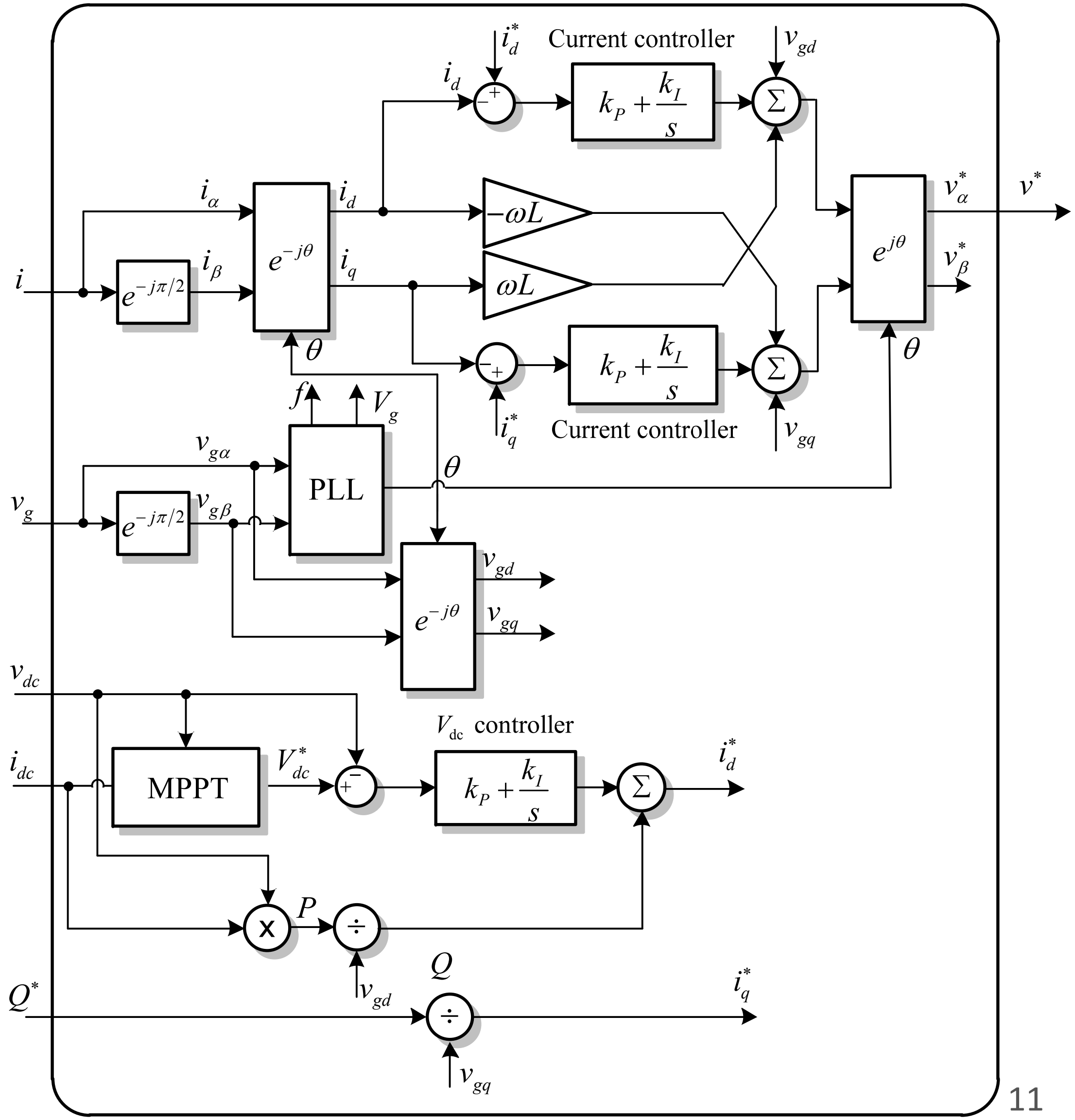
In order to overcome the limit of the PI in dealing with sinusoidal reference and harmonic disturbances, the PI control is implemented in a rotating frame



$$G_{PI}(s)_{dq} = \begin{bmatrix} k_P + \dfrac{k_I}{s} & 0 \\ 0 & k_P + \dfrac{k_I}{s} \end{bmatrix}$$

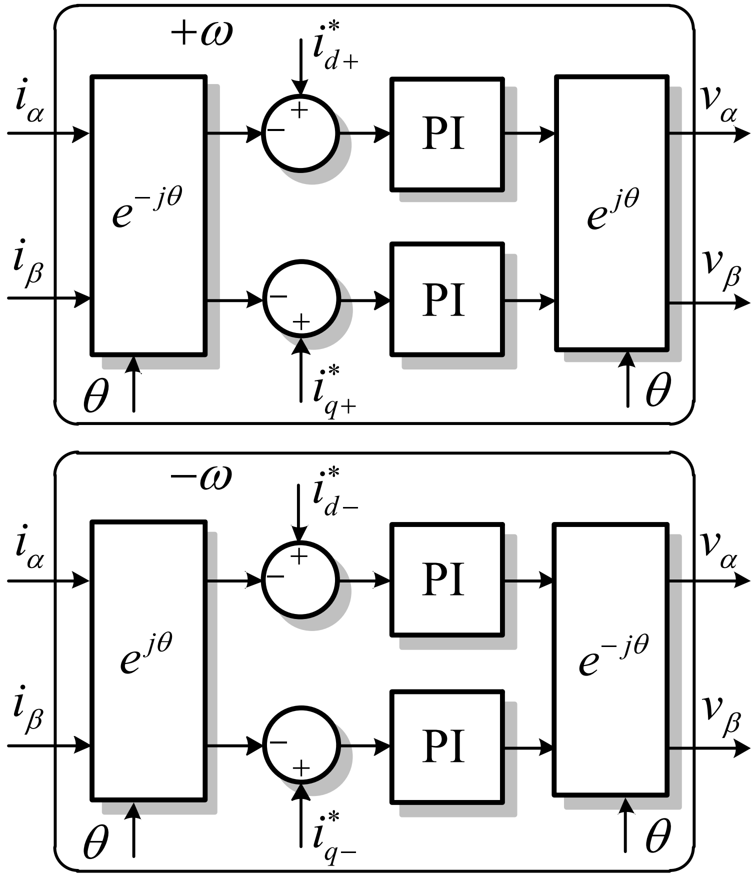
Use of a PI controllers in a rotating frame in single-phase systems

- An independent Q control is achieved
- A phase delay block create the virtual quadrature component that allows to emulate a two-phase system
- The v_β component of the command voltage is ignored for the calculation of the duty-cycle



Use of a PI controllers in two rotating frames

- Under unbalanced conditions in order to compensate the harmonics generated by the inverse sequence present in the grid voltage both the positive- and negative-sequence reference frames are required
- Obviously using this approach, double computational effort must be devoted

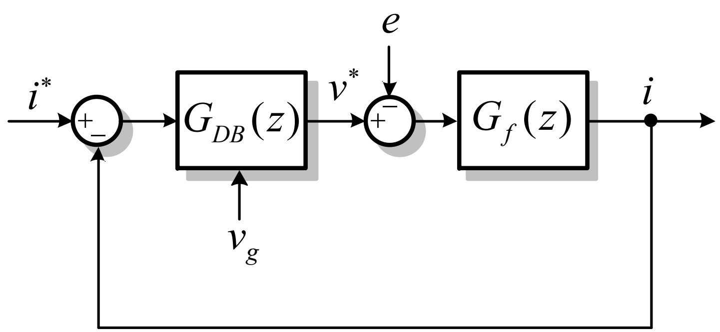


Linear current control with separated modulation: Dead-beat controller

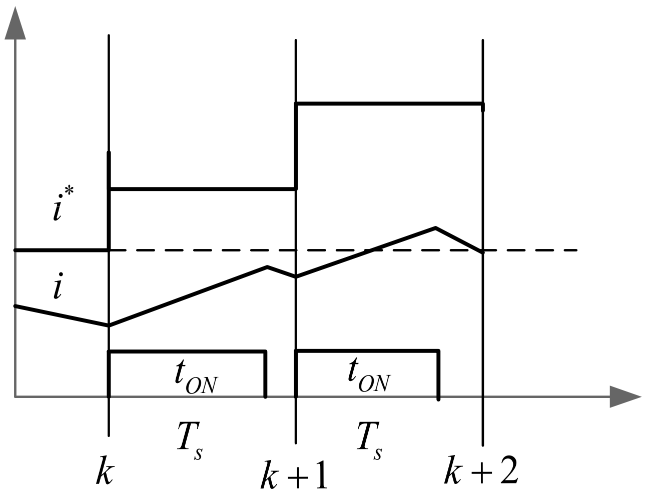
- The dead-beat controller belongs to the family of the predictive controllers
- They are based on a common principle: to foresee the evolution of the controlled quantity (the current) and on the basis of this prediction:
  - To choose the state of the converter (ON-OFF predictive) or
  - The average voltage produced by the converter (predictive with pulse width modulator)
- The starting point is to calculate its derivative to predict the effect of the control action
- The controller is developed on the basis of the model of the filter and of the grid, which is used to predict the system dynamic behavior: the controller is inherently sensitive to model and parameter mismatches

Dead-beat controller

- The information on the model is used to decide the switching state of the converter with the aim to minimize the possible commutations (ON-OFF predictive) or the average voltage that the converter has to produce in order to null it



- In case it is imposed that the error at the end of the next sampling period is zero the controller is defined as “dead-beat”. It can be demonstrated that it is the fastest current controller allowing nulling the error after two sampling periods

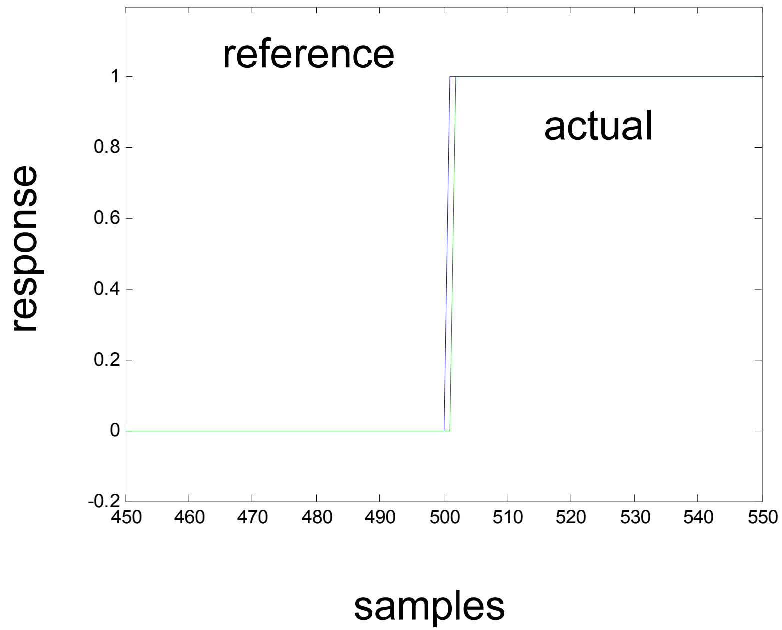


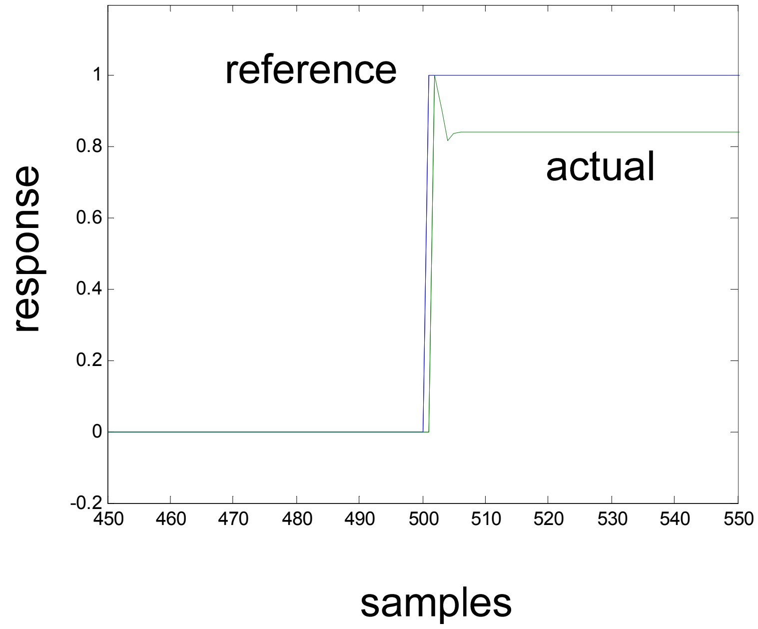
Dead-beat controller

$$v(k+1) = v(k-1) + \frac{1}{b}\Delta i(k) - \frac{a}{b}\Delta i(k-1) + e(k+1) - e(k-1)$$

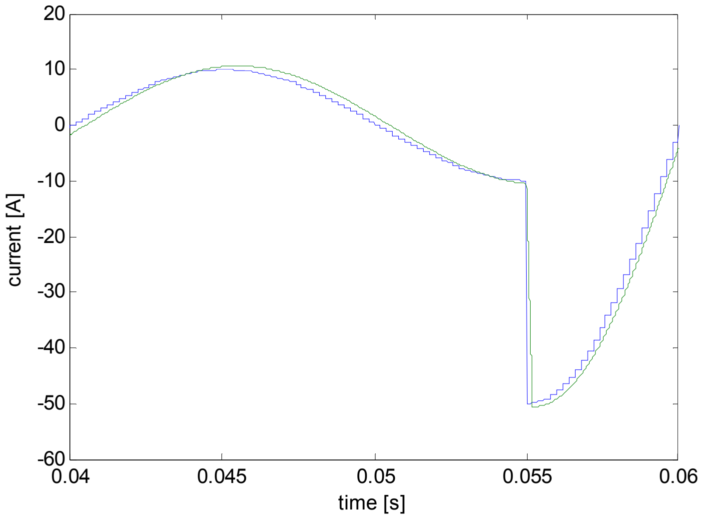
$$v(k+1) = -v(k) + \frac{1}{b}\Delta i(k) + e(k+1) + e(k)$$

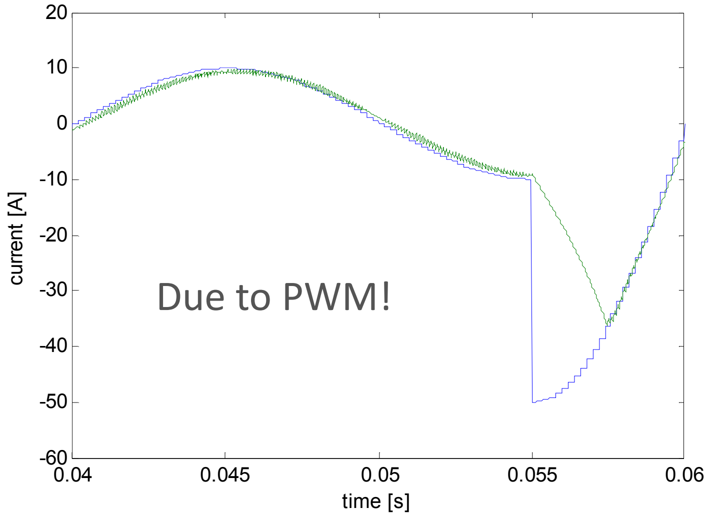
Neglecting R!

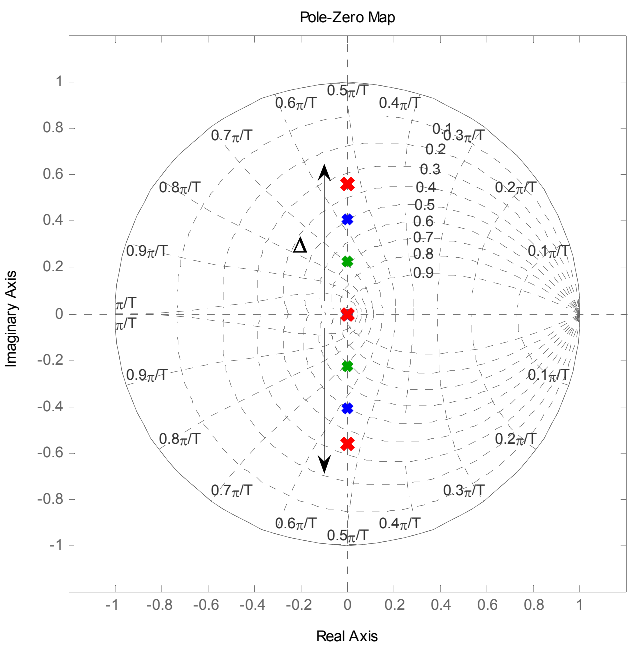


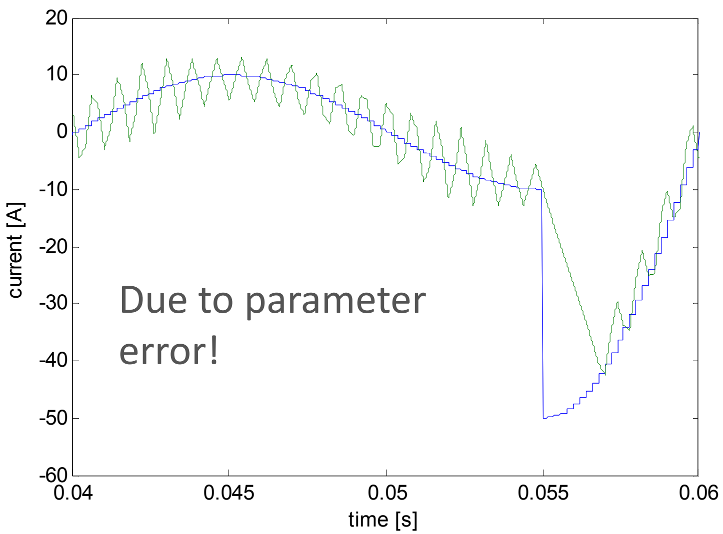


Dead-beat controller: limits







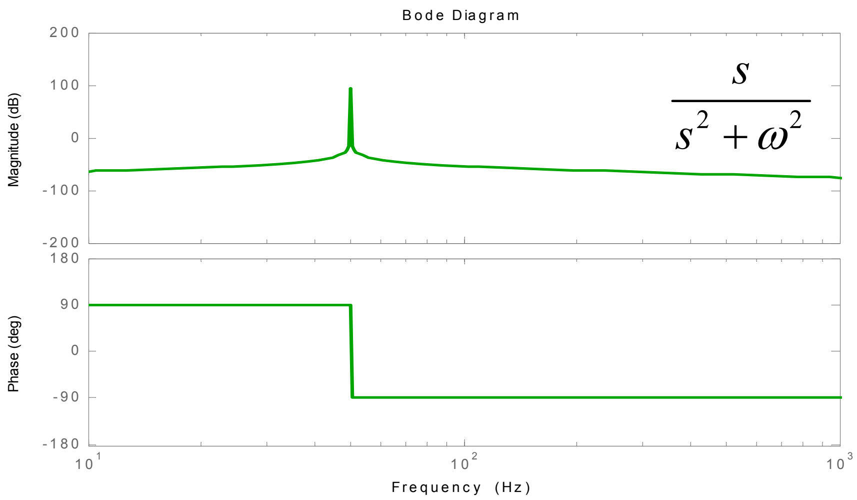


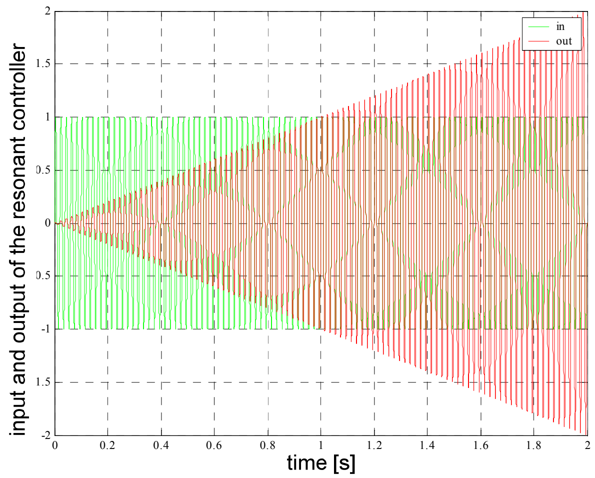
Linear current control with separated modulation: Resonant controller

- Resonant control is based on the use of Generalized Integrator (GI)
- A double integrator achieves infinite gain at a certain frequency, called resonance frequency, and almost no attenuation outside this frequency

$$\text{GI} \qquad \frac{s}{s^2+\omega^2}$$

- The GI will lead to zero stationary error and improved and selective disturbance rejection as compared with PI controller





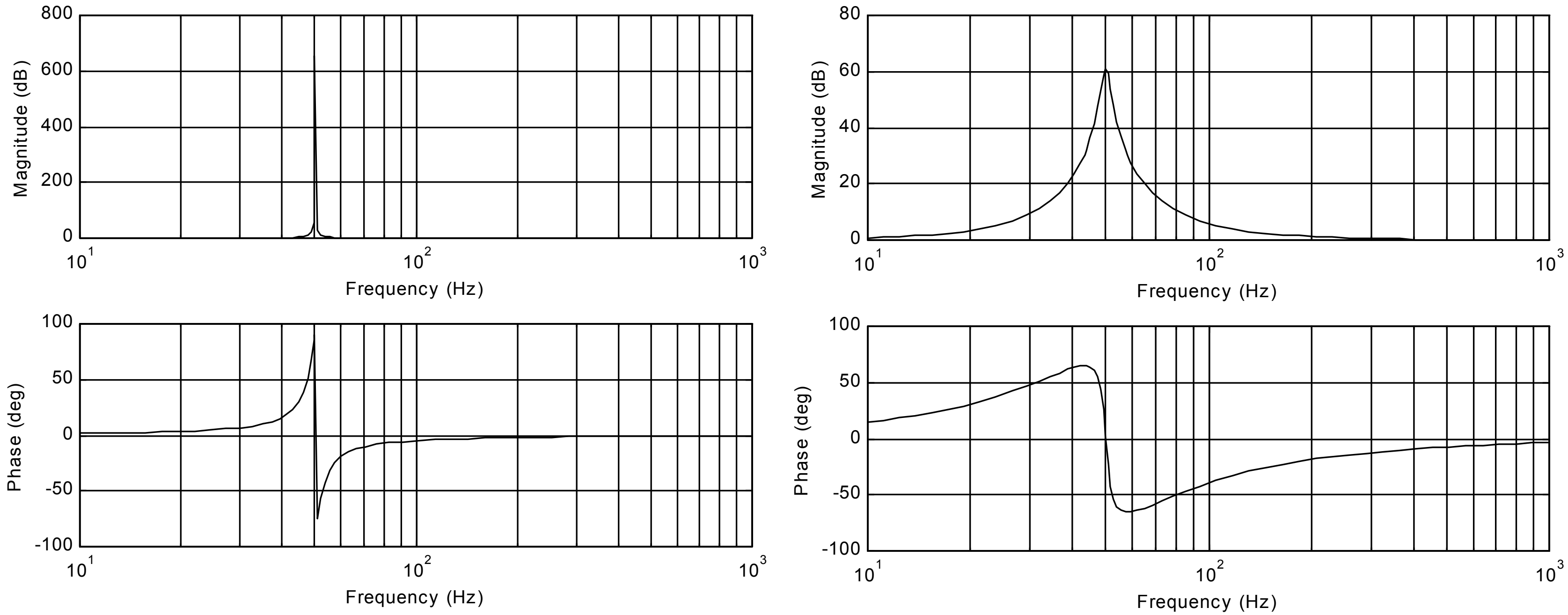
Resonant control

- The resonant controller can be obtained via a frequency shift

$$G_{AC}(s) = G_{DC}(s - j\omega) + G_{DC}(s + j\omega)$$

$$G_{DC}(s) = \frac{k_I}{s} \longrightarrow G_{AC}(s) = \frac{2k_I s}{s^2 + \omega^2}$$

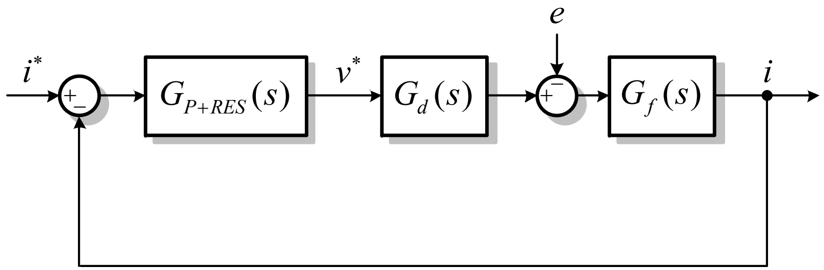
$$G_{DC}(s) = \frac{k_I}{\left(1 + (s/\omega_c)\right)} \longrightarrow G_{AC}(s) \approx \frac{2k_I \omega_c s}{s^2 + 2\omega_c s + \omega^2}$$



Bode plots of ideal and non-ideal PR with $k_P = 1$, $k_I = 20$, $\omega = 314$ rad/s, $\omega_c = 10$ rad/s

Resonant control

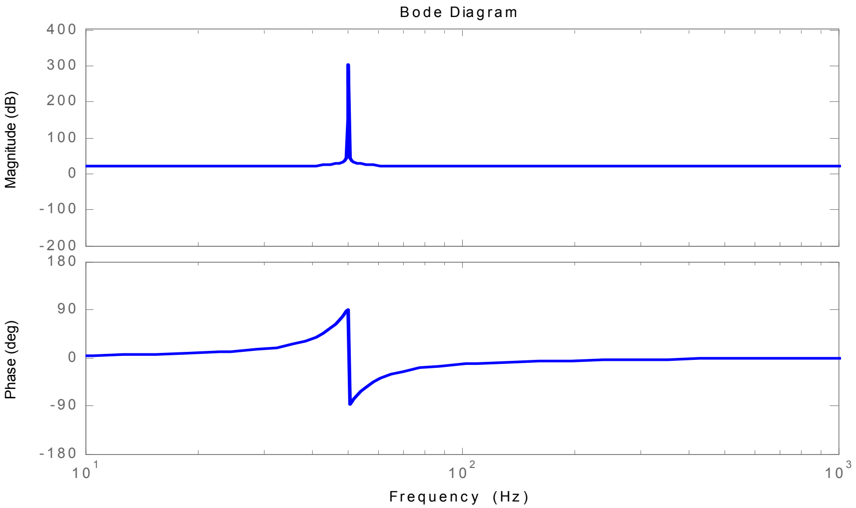
- The stability of the system should be taken into consideration

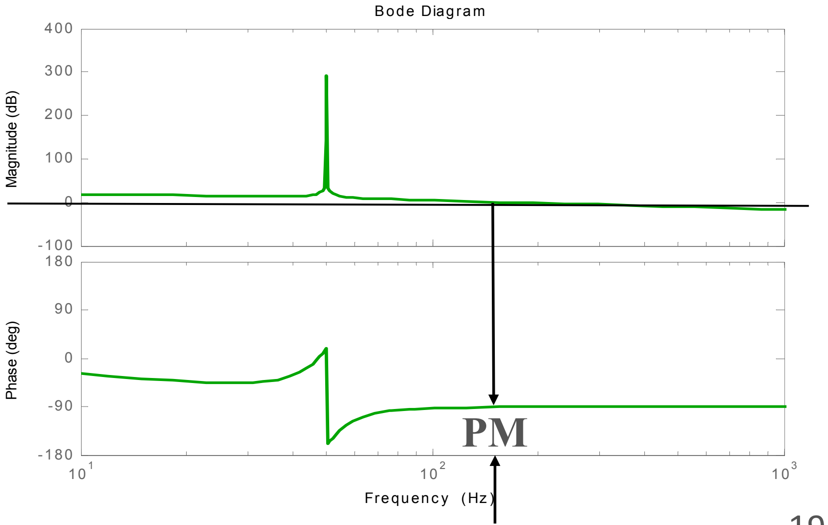


- The phase margin (PM) decreases as the resonant frequency approach to the crossover frequency

$$k_P + k_I \frac{s}{s^2+\omega^2}$$

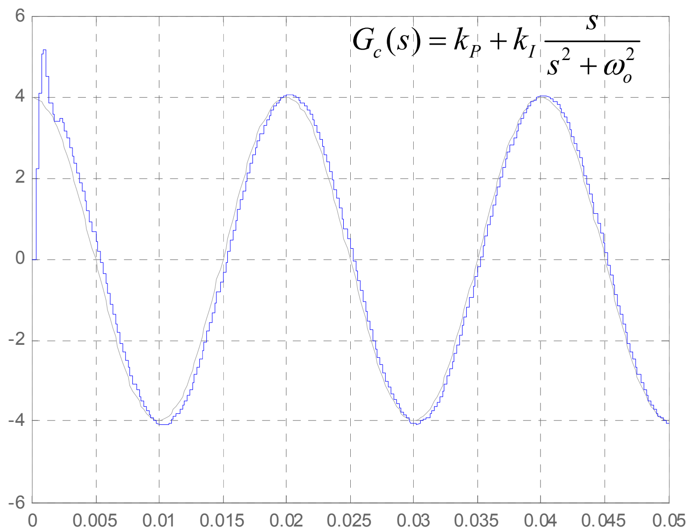
$$\left(k_P + k_I \frac{s}{s^2+\omega^2}\right)\left(\frac{1}{R+Ls}\right)$$



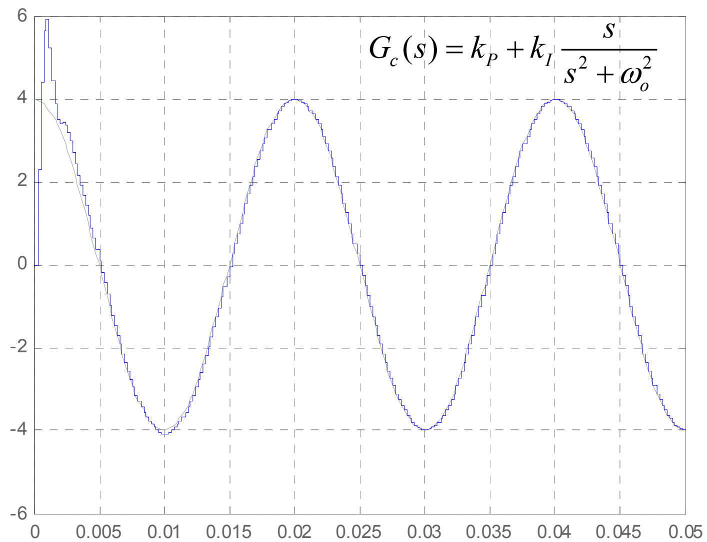


Tuning of resonant control

- The gain k_P is founded by ensuring the desired bandwidth using either rlocus Matlab function or SISOTOOL
- The integral constant k_I acts to eliminate the steady-state phase error



$k_I = 100$



$k_I = 500$

- A higher k_I will "catch" the reference faster but with higher overshoot
- Another aspect is that k_I determines the bandwidth centered at the resonance frequency, in this case the grid frequency, where the attenuation is positive. Usually, the grid frequency is stiff and is only allowed to vary in a narrow range, typically ± 1%

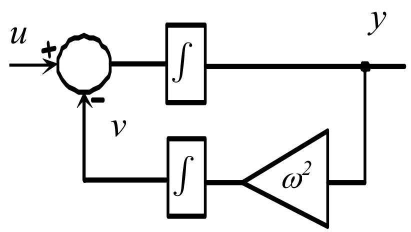
Discretization of generalized integrators

GI integrator decomposed in two simple integrators

$$\frac{y(s)}{u(s)} = \frac{s}{s^2+\omega^2} \Leftrightarrow \begin{cases} y(s) = \frac{1}{s}\left[u(s) - v(s)\right] \\ v(s) = \frac{1}{s} \cdot \omega^2 \cdot y(s) \end{cases}$$

Forward integrator for direct path and backward for feedback path

$$\begin{cases} y_k = y_{k-1} + T_s \cdot (u_{k-1} - v_{k-1}) \\ v_k = v_{k-1} + T_s \cdot \omega^2 \cdot y_k \end{cases}$$



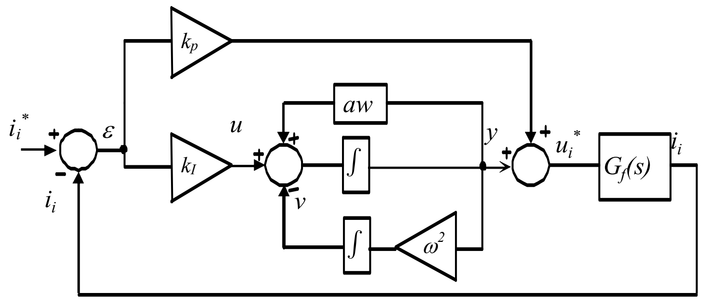
The inverter voltage reference

$$u_i^*(s) = \varepsilon(s) \cdot \left(k_P + \frac{k_I \cdot s}{s^2+\omega^2}\right)$$

Difference equations

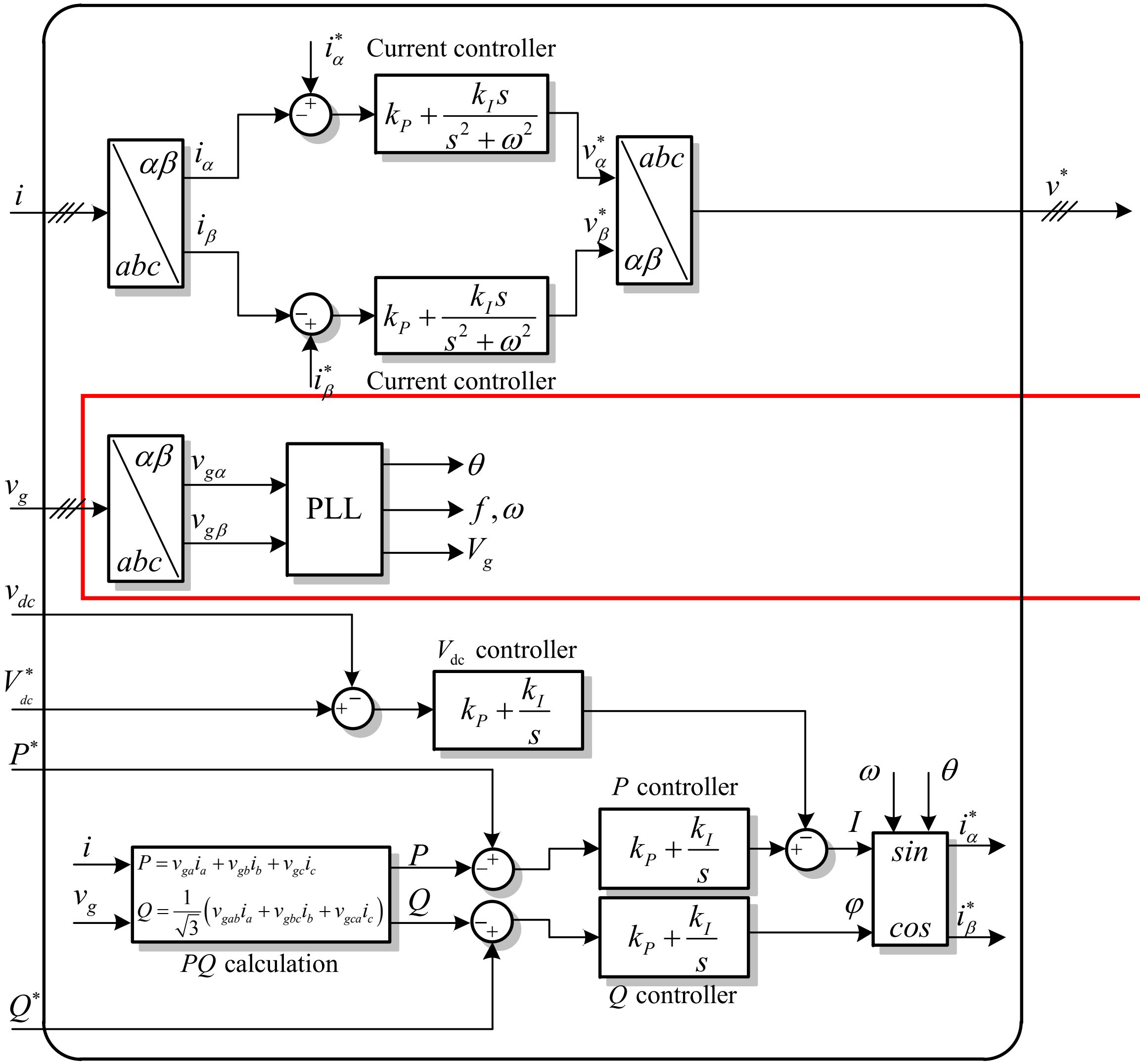
$$\begin{cases} y_k = y_{k-1} + T_s \cdot k_I \cdot \varepsilon_{k-1} - T_s \cdot v_{k-1} \\ u_{i,k}^* = k_P \cdot \varepsilon_k + y_k \\ v_k = v_{k-1} + T_s \cdot \omega^2 \cdot y_k \\ \varepsilon_{k-1} = \varepsilon_k \\ y_{k-1} = y_k \\ v_{k-1} = v_k \end{cases}$$

Control diagram of PR implementation



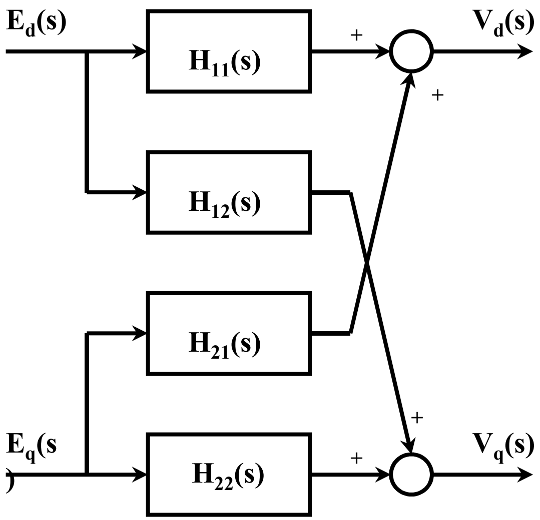
$$aw = \begin{cases} y_{\max} - y & , y > y_{\max} \\ -y_{\max} - y & , y < -y_{\max} \end{cases}$$

Use of P+resonant controller in stationary frame



- PLL is still indispensable for reference generation

From PI in a rotating-frame to P+res for each phase

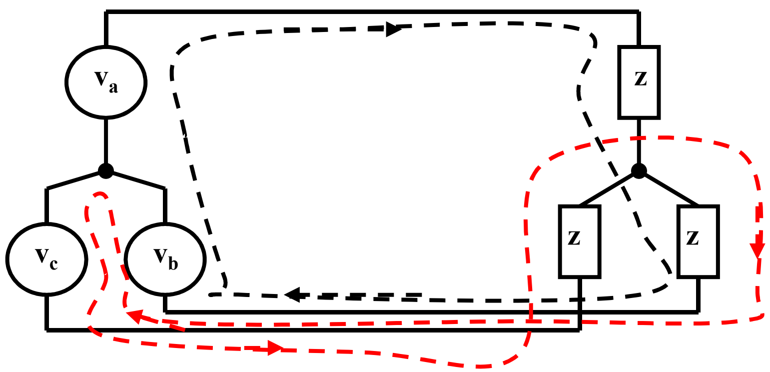


- In the hypothesis
- $\mathrm{H}_{11}(\mathrm{s}) = \mathrm{H}_{22}(\mathrm{s}) = k_P + \dfrac{k_I}{s}$
- $\mathrm{H}_{12}(\mathrm{s}) = \mathrm{H}_{21}(\mathrm{s}) = 0$

$$\begin{cases} v_d(t) = h_{11}(t) * e_d(t) \\ v_q(t) = h_{22}(t) * e_q(t) \end{cases}$$

$$G_c^{(d,q)} = \begin{bmatrix} k_P + \dfrac{k_I}{s} & 0 \\ 0 & k_P + \dfrac{k_I}{s} \end{bmatrix} \Leftrightarrow G_c^{(a,b,c)}(s) = \frac{2}{3} \cdot \begin{bmatrix} k_P + \dfrac{k_I s}{s^2+\omega_0^2} & -\dfrac{k_P}{2} - \dfrac{k_I s + \sqrt{3} k_I \omega_0}{2\cdot(s^2+\omega_0^2)} & -\dfrac{k_P}{2} - \dfrac{k_I s - \sqrt{3} k_I \omega_0}{2\cdot(s^2+\omega_0^2)} \\ -\dfrac{k_P}{2} - \dfrac{k_I s - \sqrt{3} k_I \omega_0}{2\cdot(s^2+\omega_0^2)} & k_P + \dfrac{k_I s}{s^2+\omega_0^2} & -\dfrac{k_P}{2} - \dfrac{k_I s + \sqrt{3} k_I \omega_0}{2\cdot(s^2+\omega_0^2)} \\ -\dfrac{k_P}{2} - \dfrac{k_I s + \sqrt{3} k_I \omega_0}{2\cdot(s^2+\omega_0^2)} & -\dfrac{k_P}{2} - \dfrac{k_I s - \sqrt{3} k_I \omega_0}{2\cdot(s^2+\omega_0^2)} & k_P + \dfrac{k_I s}{s^2+\omega_0^2} \end{bmatrix}$$

Linear controllers: from PI in a rotating-frame to P+res for each phase



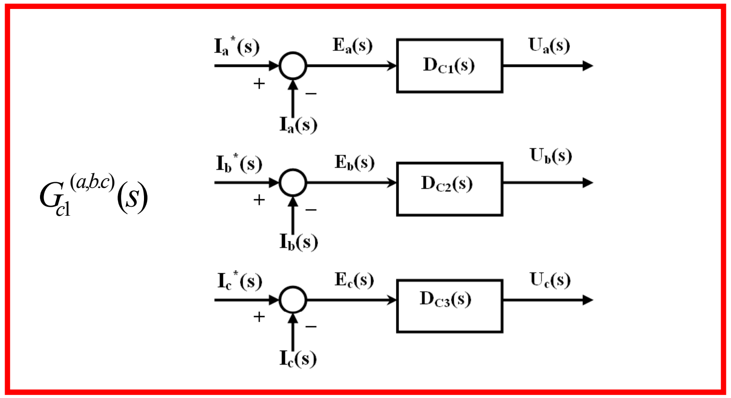
$$\frac{d}{dt}\begin{bmatrix} i_a(t) \\ i_c(t) \end{bmatrix} = \frac{1}{L}\begin{bmatrix} -R & 0 \\ 0 & R \end{bmatrix}\cdot\begin{bmatrix} i_a(t) \\ i_c(t) \end{bmatrix} + \frac{1}{3}\begin{bmatrix} 2 & -1 & -1 \\ 1 & 1 & 2 \end{bmatrix}\cdot\begin{bmatrix} v_a(t) \\ v_b(t) \\ v_c(t) \end{bmatrix}$$

$$v_a(t) + v_b(t) + v_c(t) = 0$$

$$\frac{d}{dt}\begin{bmatrix} i_a(t) \\ i_c(t) \end{bmatrix} = \frac{1}{L}\begin{bmatrix} -R & 0 \\ 0 & R \end{bmatrix}\cdot\begin{bmatrix} i_a(t) \\ i_c(t) \end{bmatrix} + \begin{bmatrix} 1 & 0 \\ 0 & -1 \end{bmatrix}\cdot\begin{bmatrix} v_a(t) \\ v_c(t) \end{bmatrix}$$

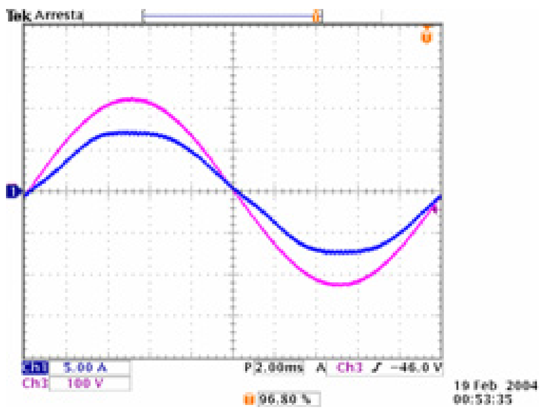
Each current is determined only by its voltage!

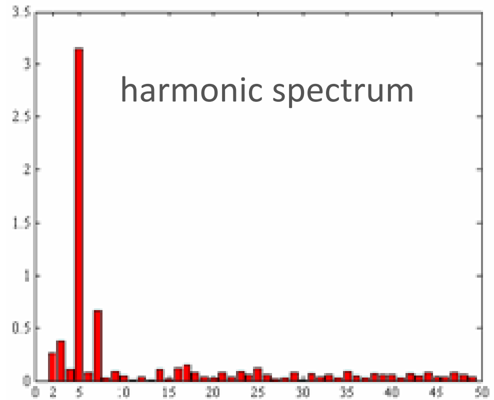
$$G_c^{(a,b,c)}(s) = \frac{2}{3}\cdot\begin{bmatrix} k_P + \frac{k_I s}{s^2+\omega_0^2} & -\frac{k_P}{2} - \frac{k_I s + \sqrt{3}k_I\omega_0}{2\cdot(s^2+\omega_0^2)} & -\frac{k_P}{2} - \frac{k_I s - \sqrt{3}k_I\omega_0}{2\cdot(s^2+\omega_0^2)} \\ -\frac{k_P}{2} - \frac{k_I s - \sqrt{3}k_I\omega_0}{2\cdot(s^2+\omega_0^2)} & k_P + \frac{k_I s}{s^2+\omega_0^2} & -\frac{k_P}{2} - \frac{k_I s + \sqrt{3}k_I\omega_0}{2\cdot(s^2+\omega_0^2)} \\ -\frac{k_P}{2} - \frac{k_I s + \sqrt{3}k_I\omega_0}{2\cdot(s^2+\omega_0^2)} & -\frac{k_P}{2} - \frac{k_I s - \sqrt{3}k_I\omega_0}{2\cdot(s^2+\omega_0^2)} & k_P + \frac{k_I s}{s^2+\omega_0^2} \end{bmatrix} \Leftrightarrow$$

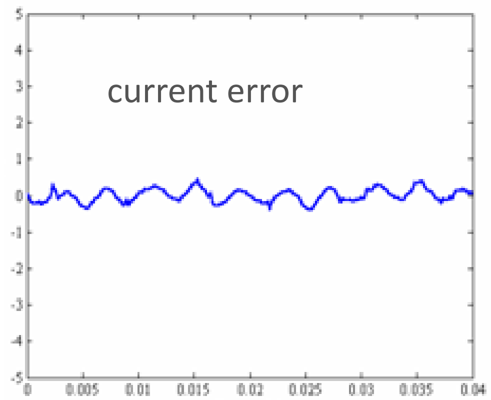


Linear controllers: results (ideal grid conditions)

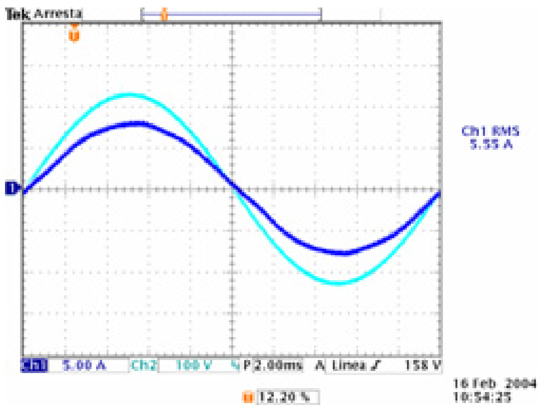
PI controller in a rotating frame

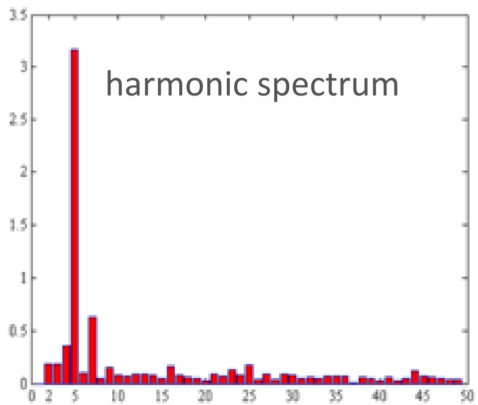


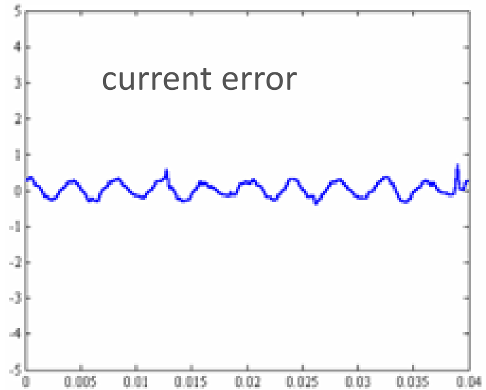




P+resonant controller for each phase

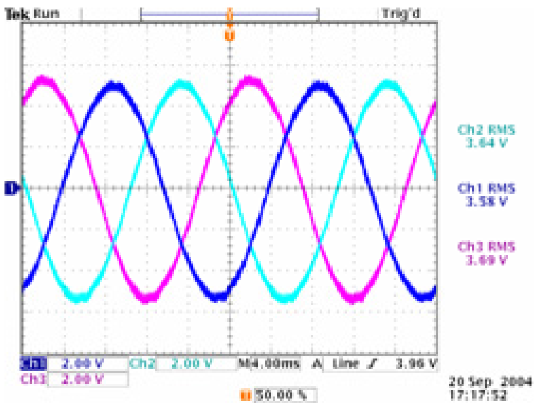




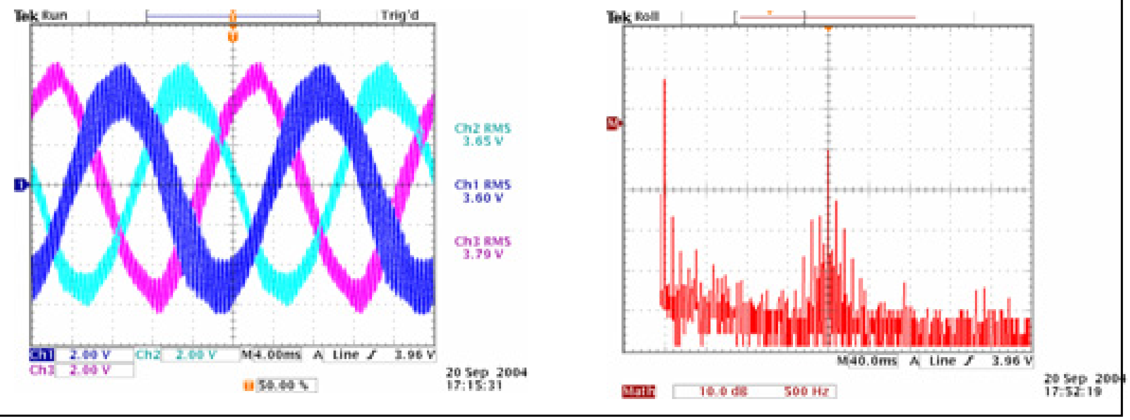


Linear controllers: results (equivalence of PI in dq and P+res in $\alpha\beta$)

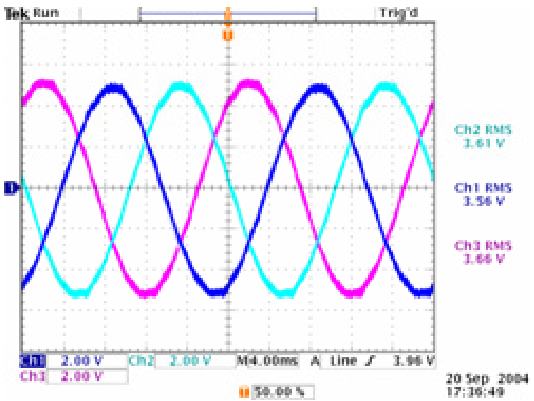
PI controller in a rotating frame



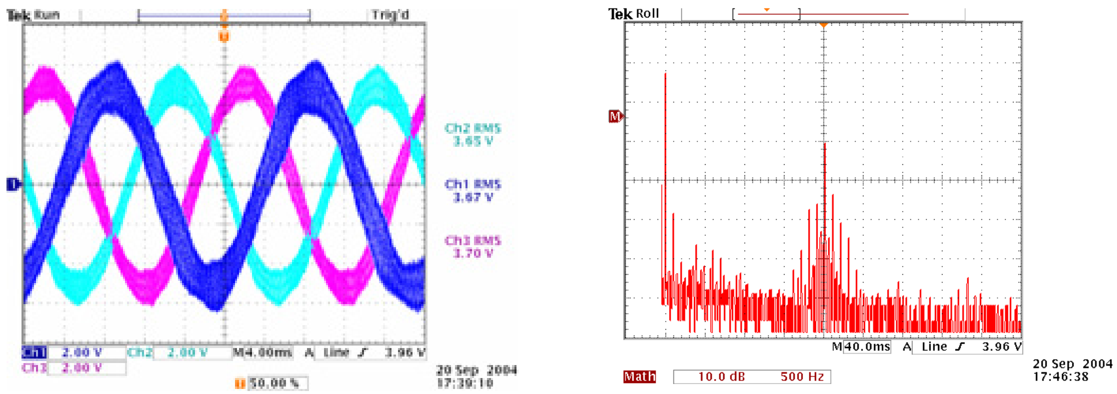
triggering LCL-filter resonance



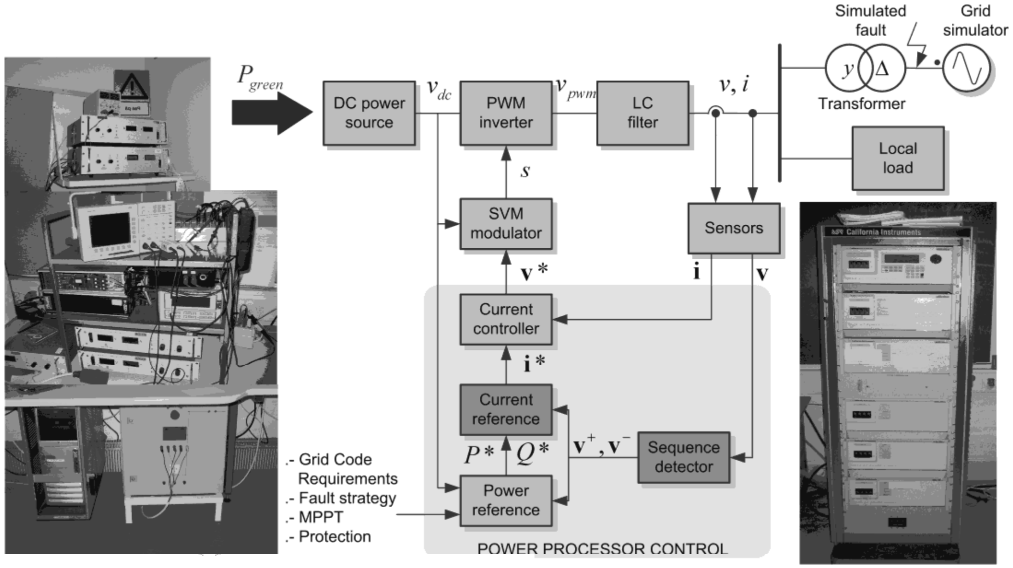
P+resonant in stationary frame

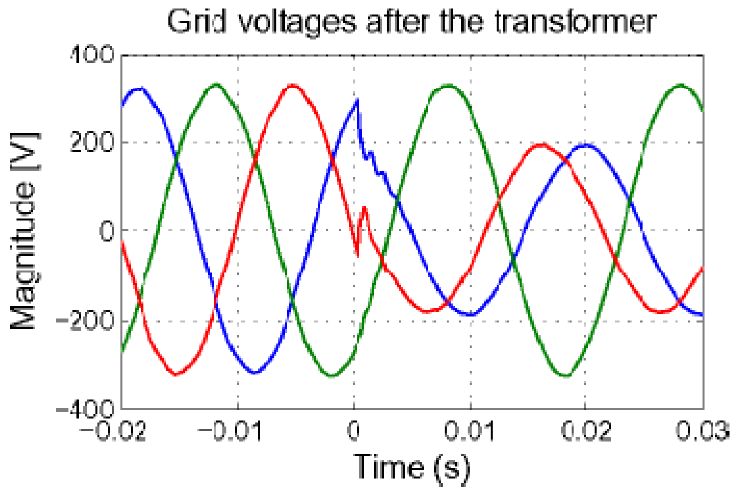


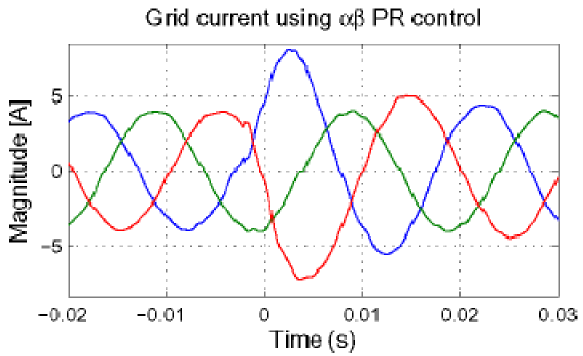
triggering LCL-filter resonance

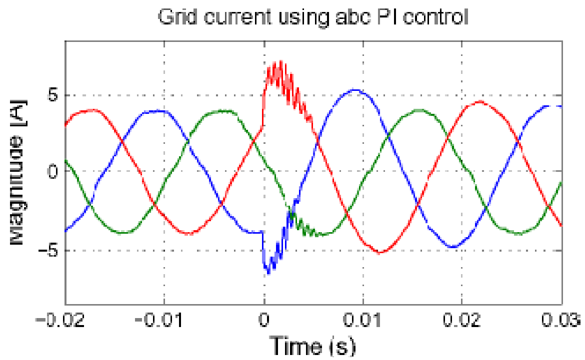


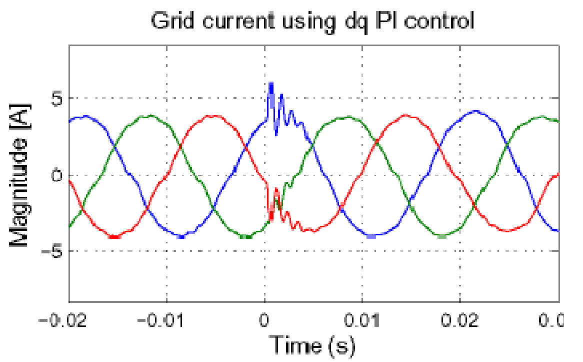
Comparison under fault

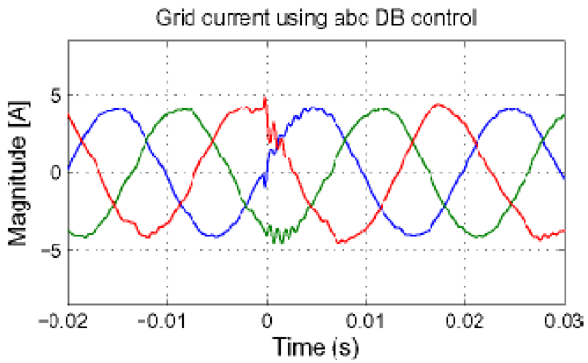












Single-phase fault:

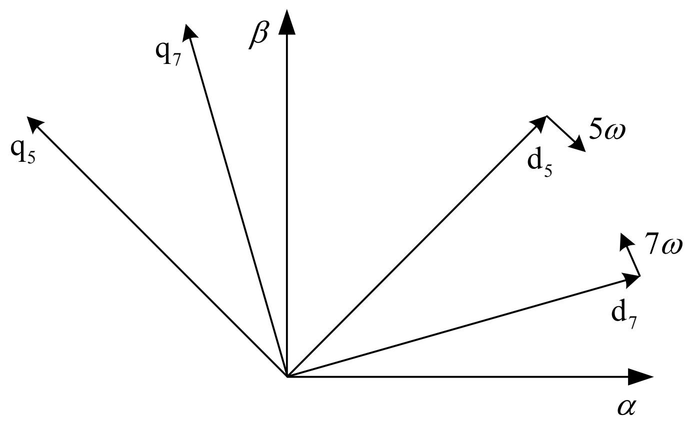
With dead-beat controller the current overshoot is limited

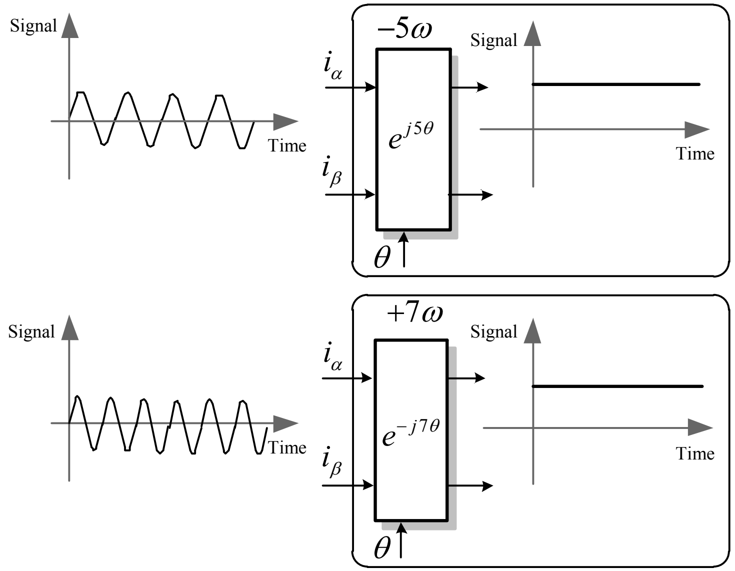
Harmonic compensation

- The decomposition of signals into harmonics with the aim of monitor and control them is a matter of interest for various electric and electronic systems
- There have been many efforts to scientifically approach typical problems (e.g. faults, unbalance, low frequency EMI) in power systems (power generation, conversion and transmission) through the harmonic analysis
- The use of Multiple Synchronous Reference Frames (MSRFs), early proposed for the study of induction machines, allows compensating selected harmonic components in case of two-phase motors, unbalance machines or in grid connected systems

Harmonic compensation

- The harmonic components of power signals can be represented in stationary or synchronous frames using phasors
- In case of synchronous reference frames each harmonic component is transformed into a dc component (frequency shifting)

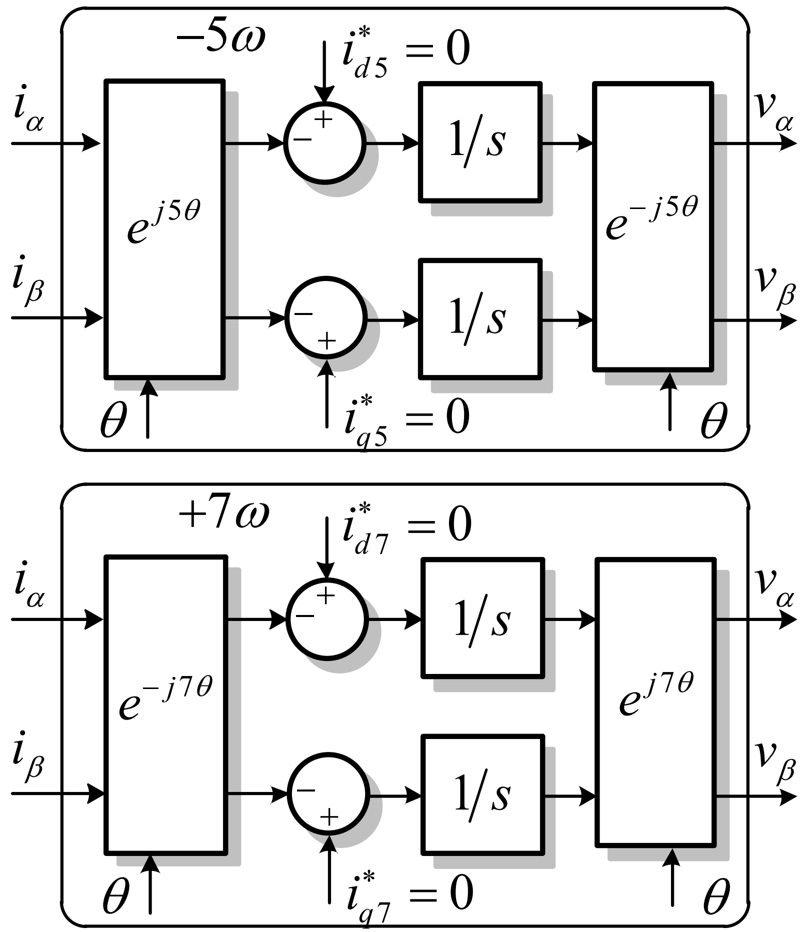




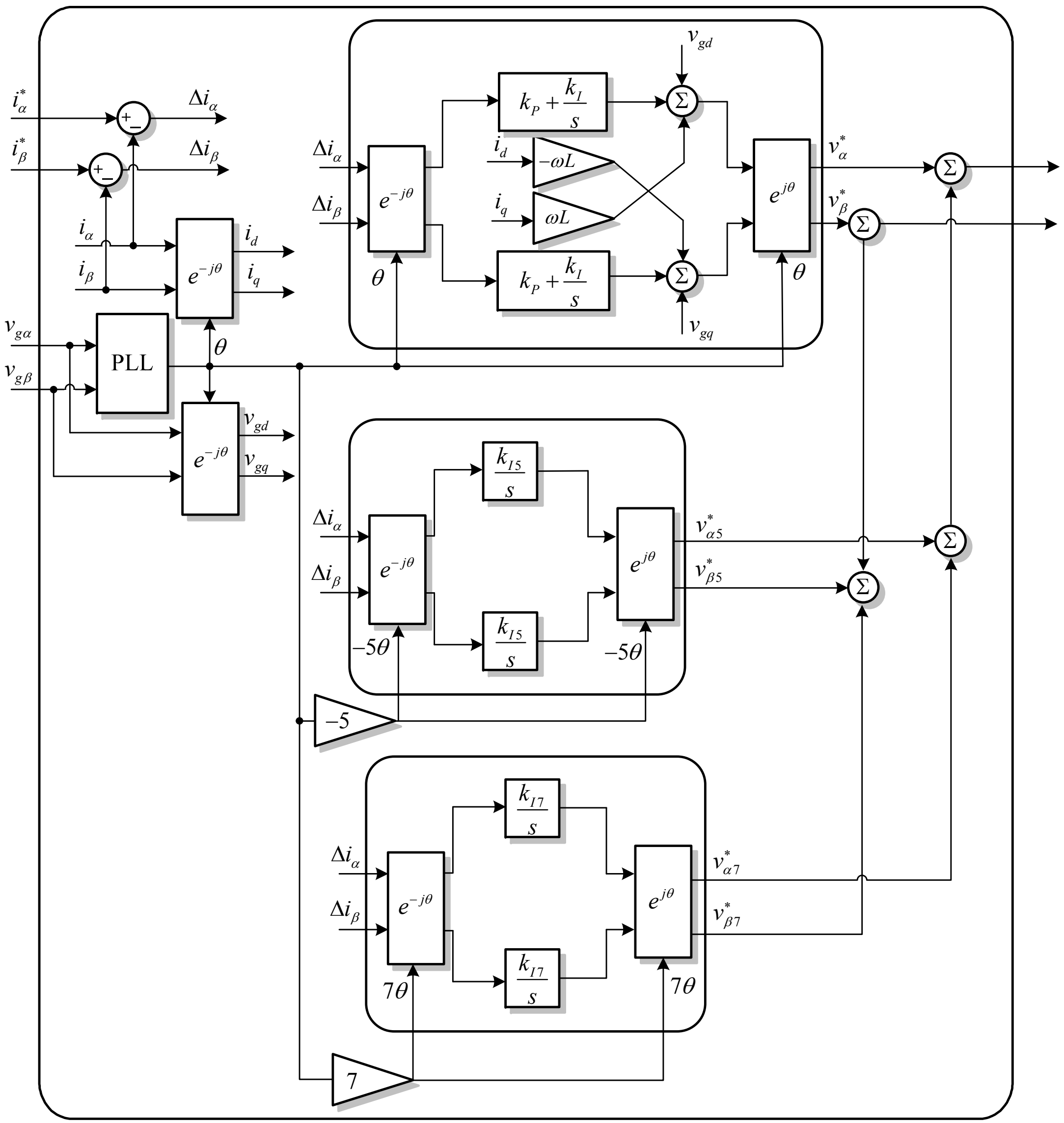
- If other harmonics are contained in the input signal, the dc output will be disturbed by a ripple that can be easily filtered out

Harmonic compensation by means of synchronous *dq*-frames

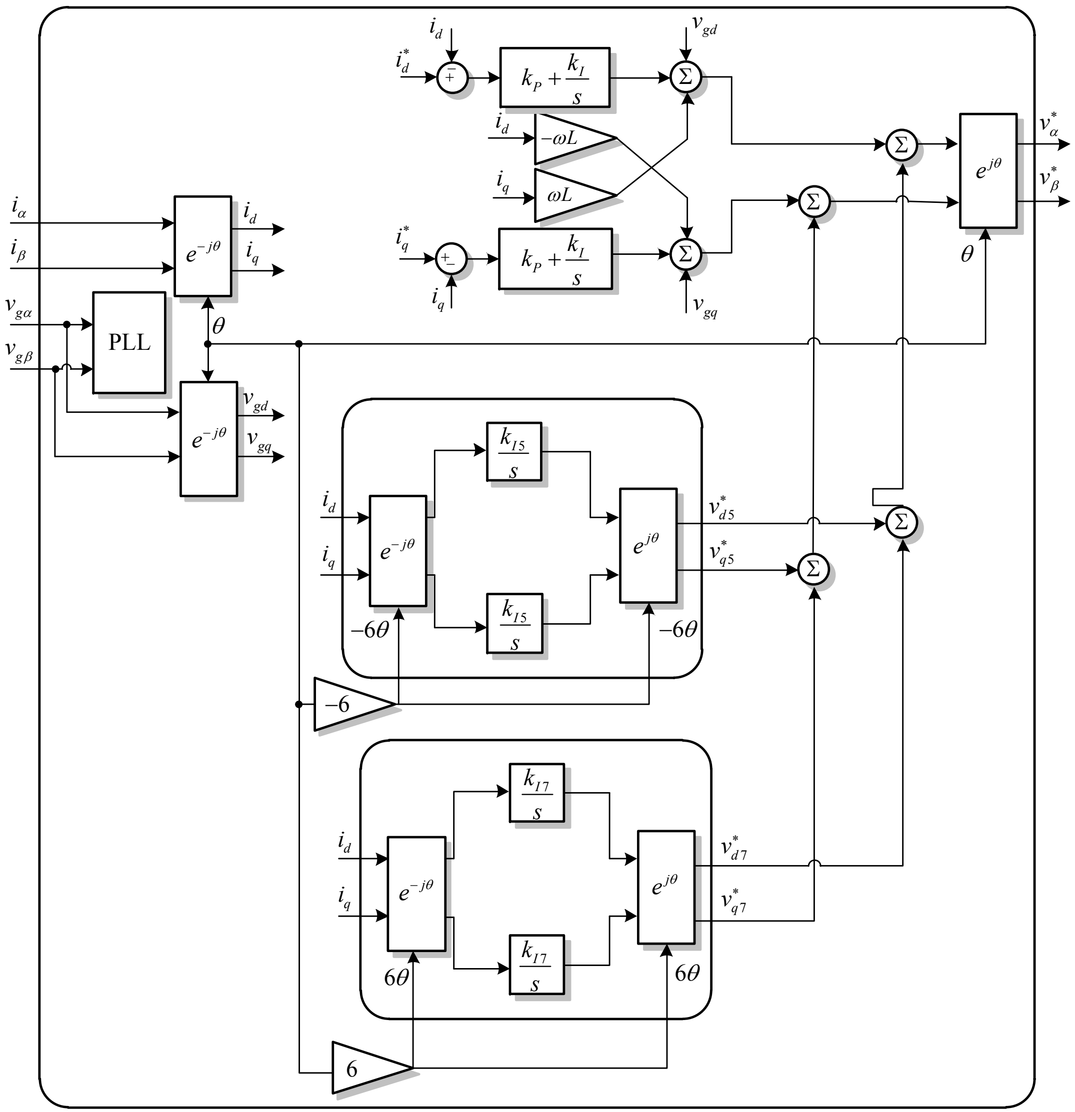
- Two controllers should be implemented in two frames rotating at -5ω and 7ω
- Or nested frames can be used i.e. implementing in the main synchronous frame two controllers in two frames rotating at 6ω and -6ω
- Both solutions are equivalent also in terms of implementation burden because in both the cases two controllers are needed



Harmonic compensation by means of synchronous *dq*-frames



Harmonic compensation by means of synchronous *dq*-frames

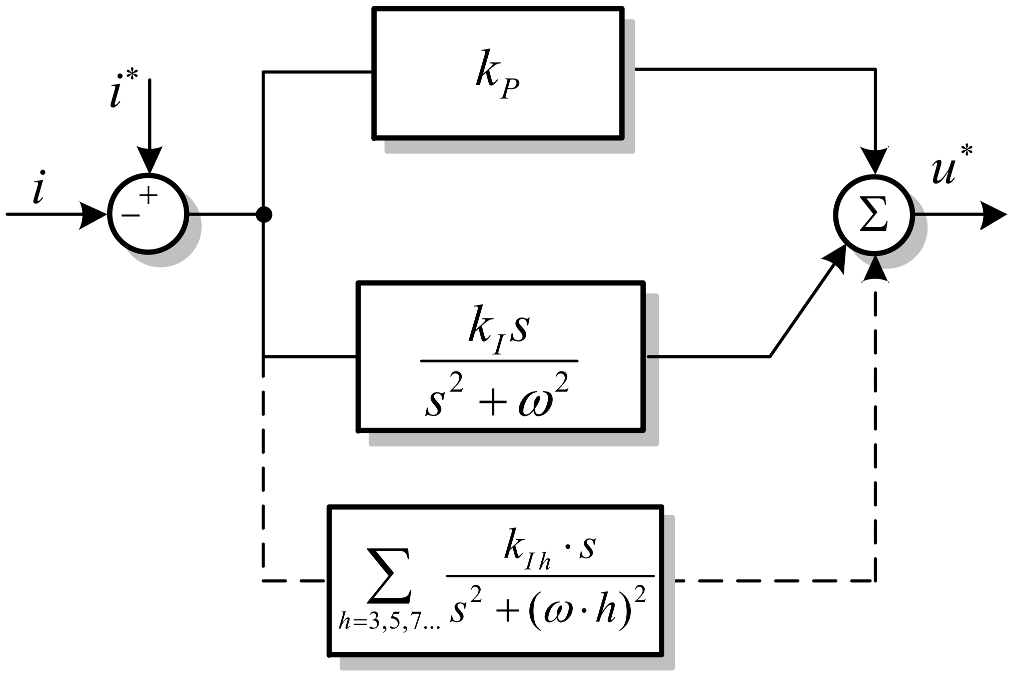


Harmonic compensation by means of stationary $\alpha\beta$-frame

Besides single frequency compensation (obtained with the generalized integrator tuned at the grid frequency), selective harmonic compensation can also be achieved by cascading several resonant blocks tuned to resonate at the desired low-order harmonic frequencies to be compensated.

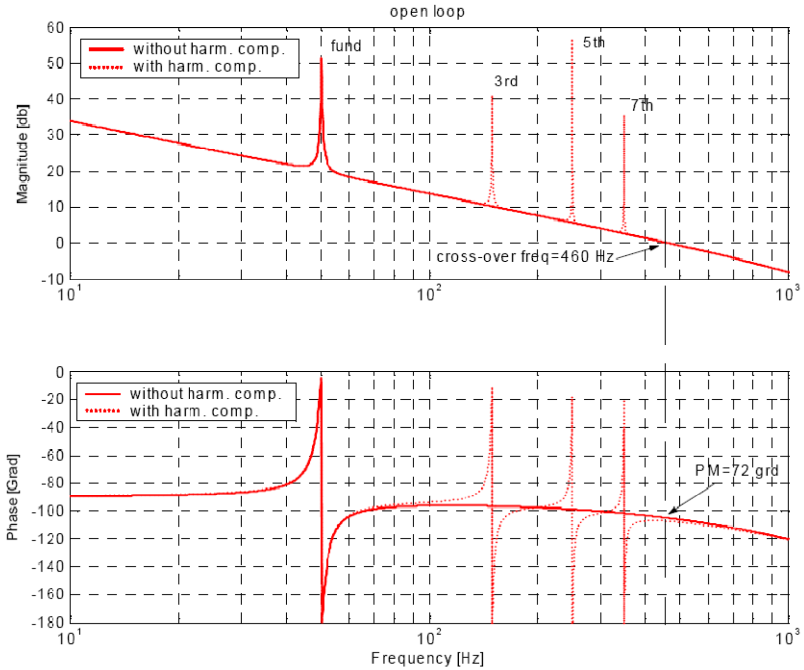
As an example, the transfer functions of a non-ideal harmonic compensator (HC) designed to compensate for the 3rd, 5th and 7th harmonics is reported.

$$G_h(s) = \sum_{h=3,5,7} \frac{2k_{Ih}\omega_c s}{s^2 + 2\omega_c s + (h\omega)^2}$$

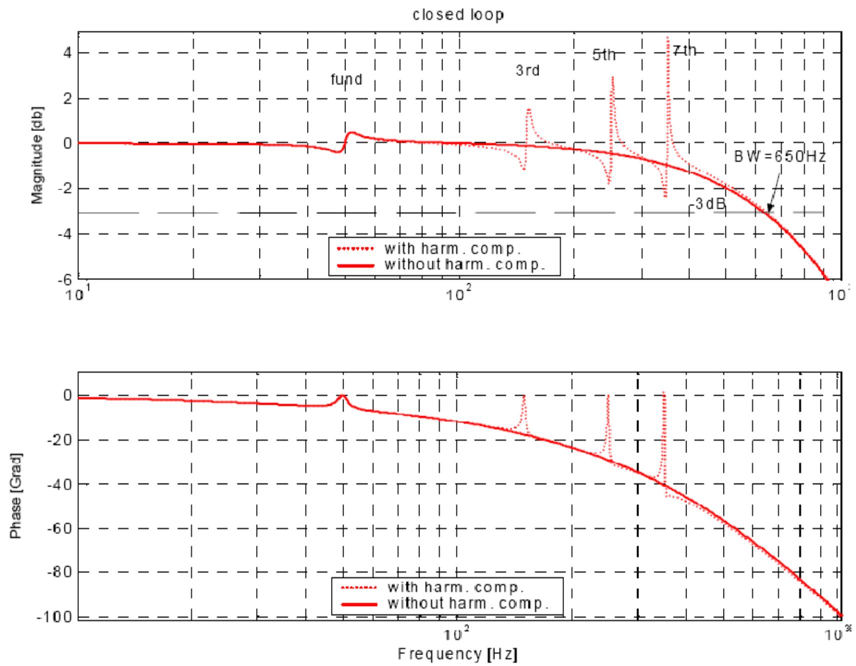


Harmonic compensation by means of stationary $\alpha\beta$-frame

Open-loop PR current control system with and without harmonic compensator



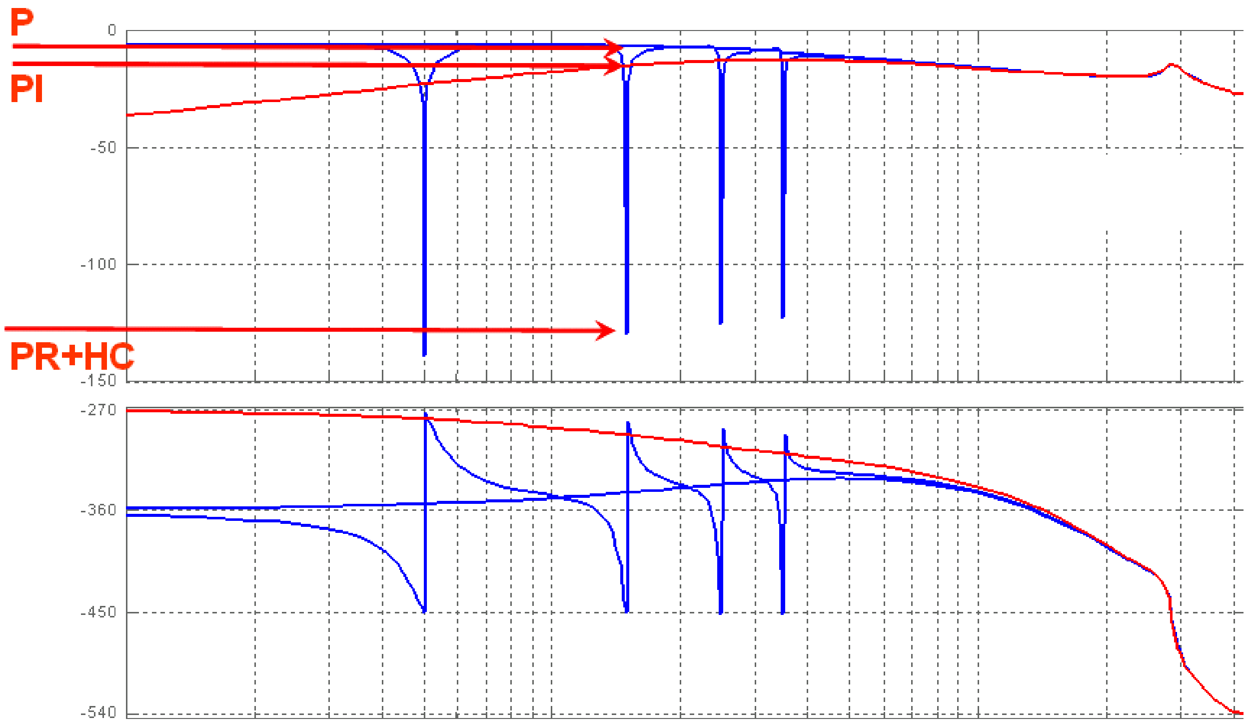
Closed loop PR current control system with and without harmonic compensator



- Having added the harmonic compensator, the open-loop and closed-loop bode graphs changes as it can be observed with dashed line. The change consists in the appearance of gain peaks at the harmonic frequencies, but what is interesting to notice is that the dynamics of the controller, in terms of bandwidth and stability margin remains unaltered

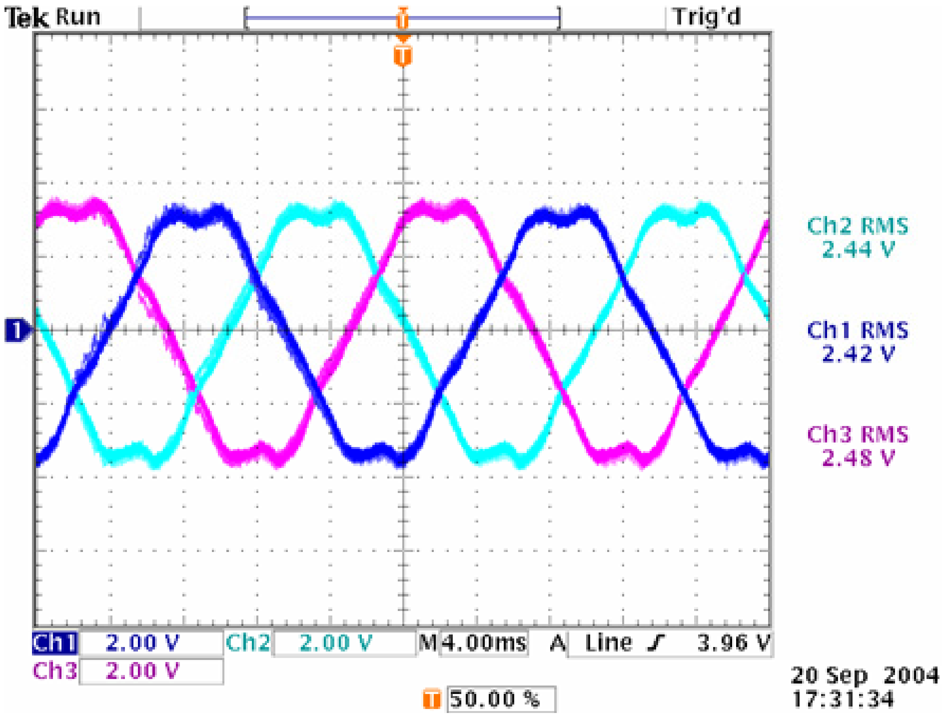
Disturbance rejection comparison

Disturbance rejection (current error ratio disturbance) of the PR+HC, PR and P

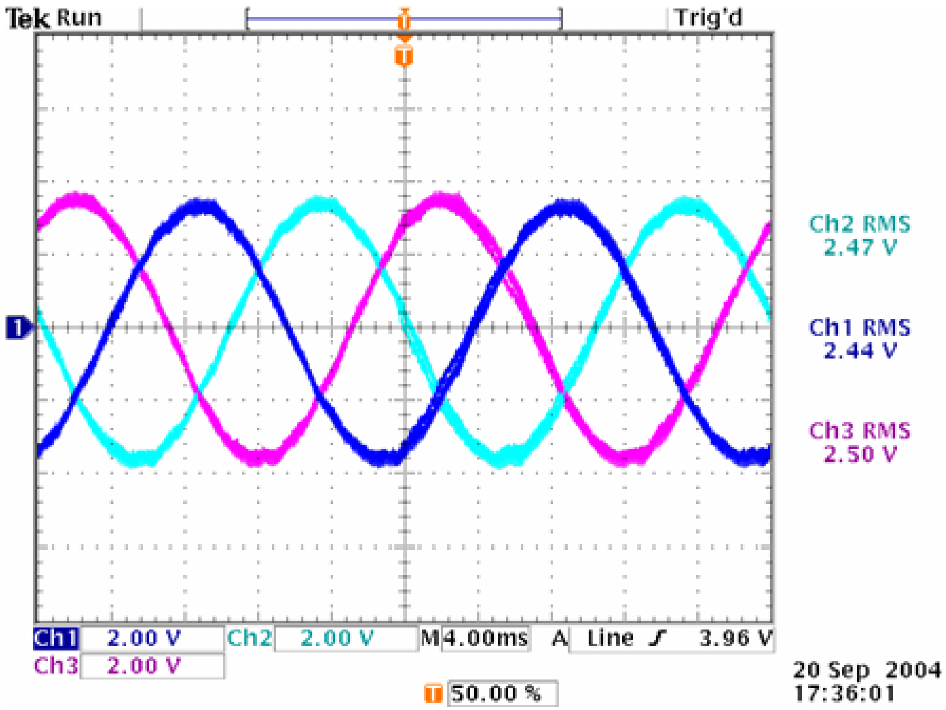


- Around the 5th and 7th harmonics the PR attenuation being around 125 dB and the PI attenuation only 8 dB. The PI rejection capability at 5th and 7th harmonic is comparable with that one of a simple proportional controller, the integral action being irrelevant
- PR +HC exhibits high performance harmonic rejections leading to very low current THD!

Results: grid voltage background distortion

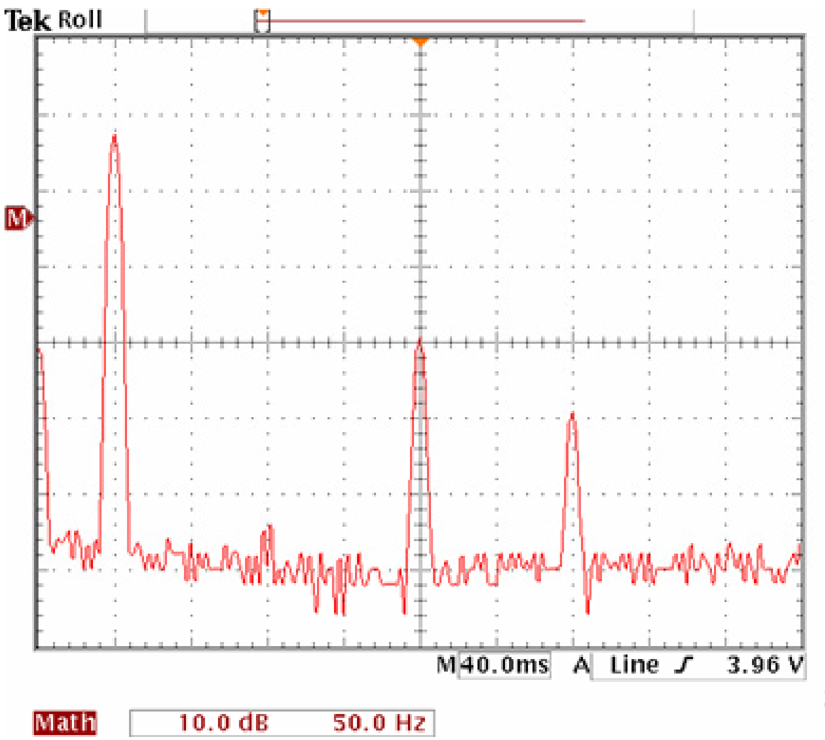


Effect of the grid voltage background distortion on the currents

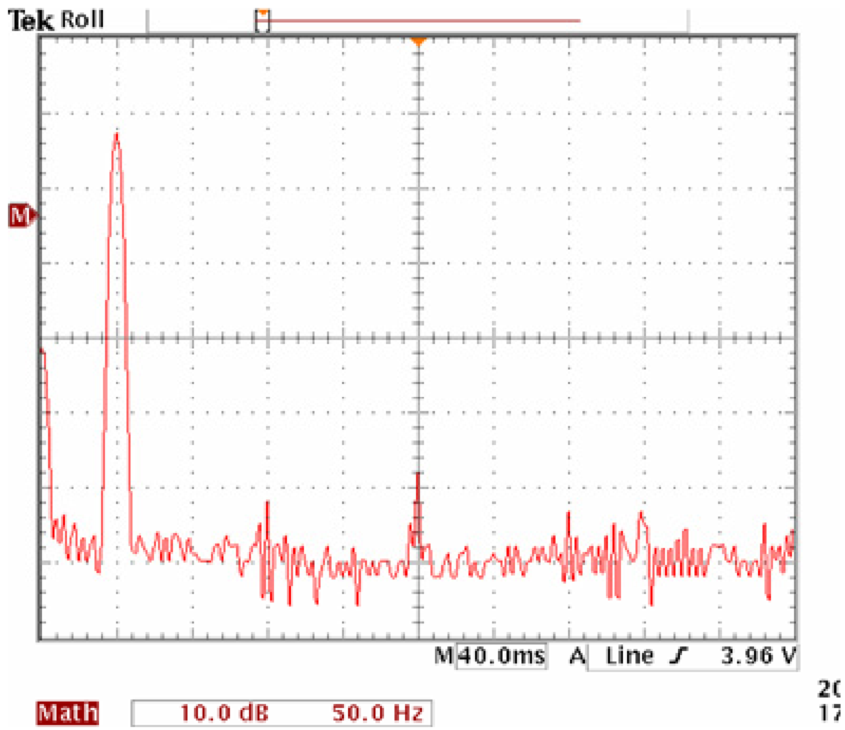


Use of harmonic compensators

Results: grid voltage background distortion



Effect of the grid voltage background distortion on the currents



Use of harmonic compensators

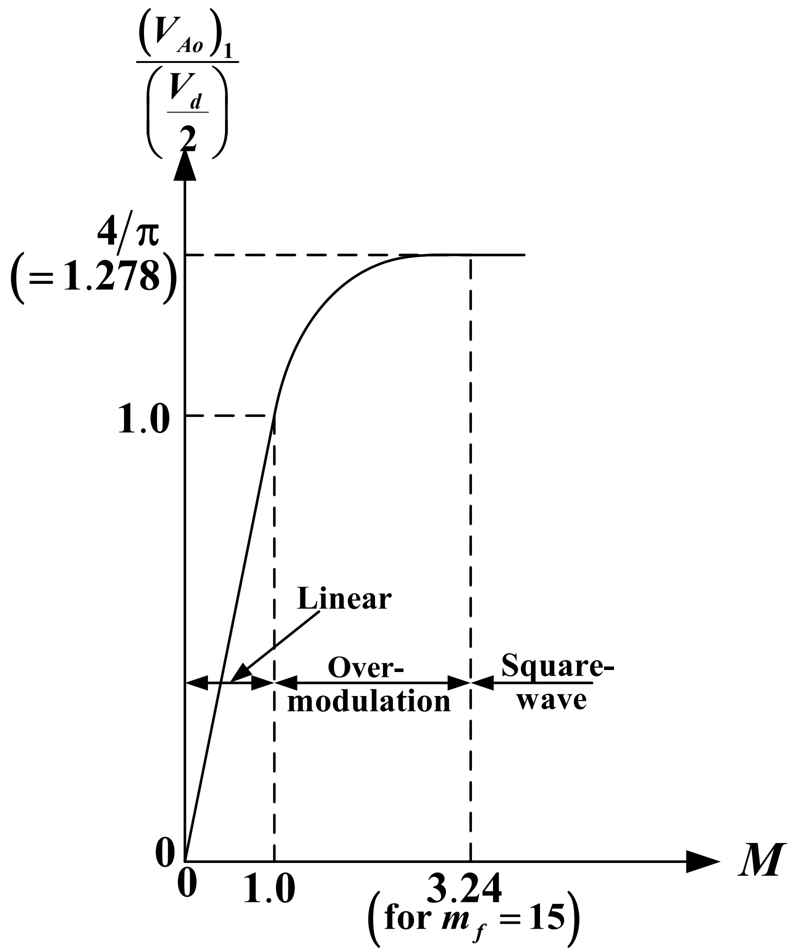
Modulation techniques

- Characteristic parameters of these strategies are:
  - The ratio between amplitudes of modulating and carrier waves (called modulation index M)
  - The ratio between frequencies of the same signals (called carrier index m)
- These techniques differ for the modulating wave chosen with the goal to obtain
  - A lower harmonic distortion
  - To shape the harmonic spectrum
  - To guarantee a linear relation between fundamental output voltage and modulation index in a wider range
- The space vector modulations are developed on the basis of the space vector representation of the converter ac side voltage

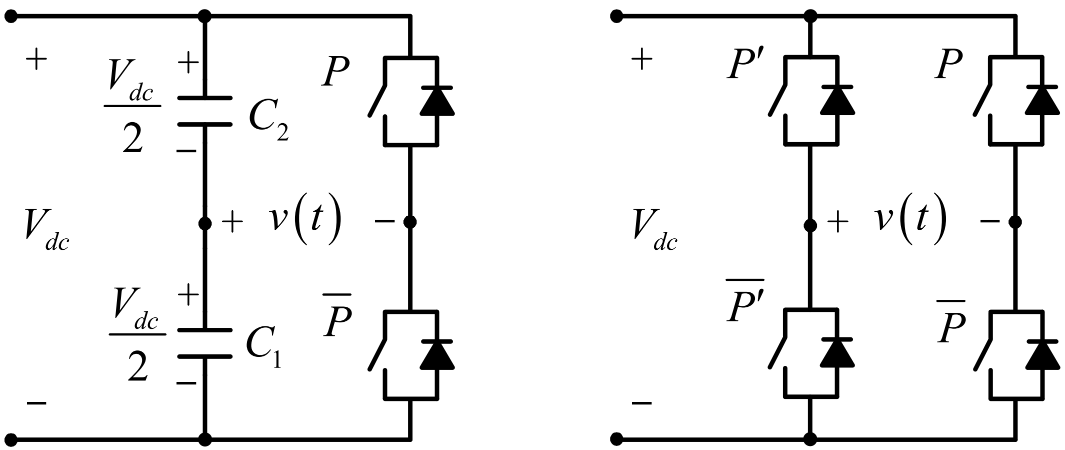
Modulation techniques

- Analogic or digital
- Natural sampled or regular sampled
- Symmetric or asymmetric

Optimization both for the linearity and harmonic content



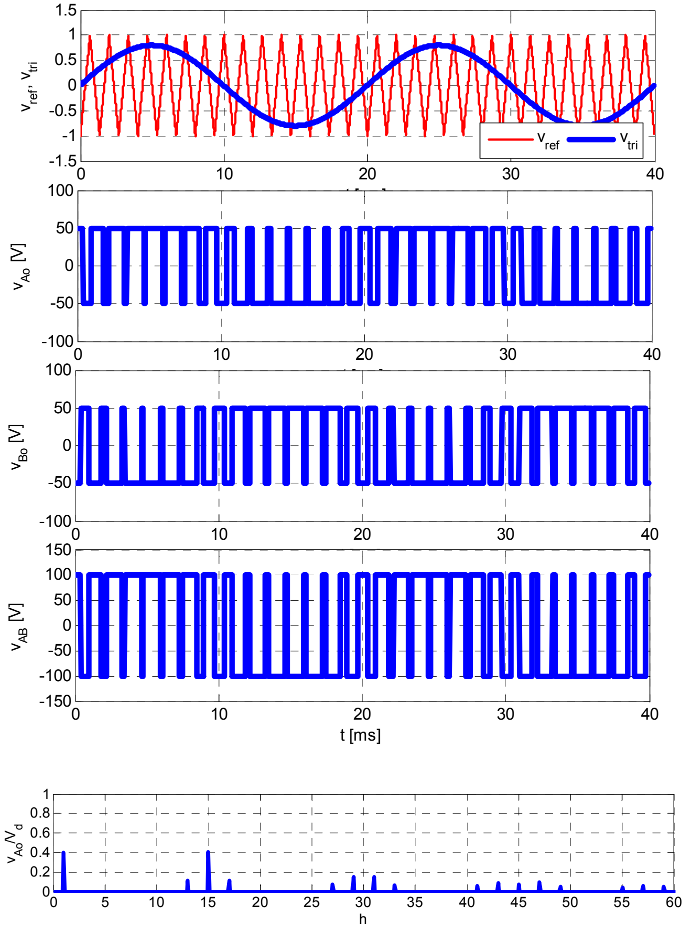
Bipolar and unipolar modulations

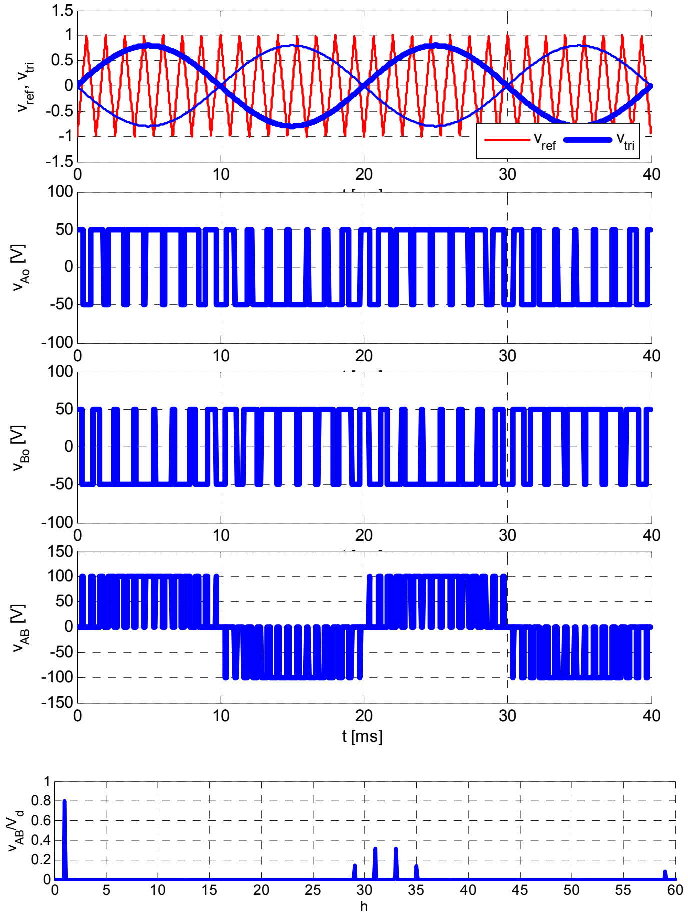


- Bipolar
$$v(t) = \frac{4V_{dc}}{\pi} \sum_{\substack{m=0 \leftrightarrow n=1 \\ m>0 \leftrightarrow n=-\infty}}^{\infty} \sum^{\infty} \frac{1}{q} J_n\left(q\frac{\pi}{2}M\right) \sin\left([m+n]\frac{\pi}{2}\right) \cos(m\omega_c t + n\omega_0 t)$$

- Unipolar
$$v(t) = 2V_{dc}M\cos(\omega_0 t) + \frac{8V_{dc}}{\pi}\sum_{m=1}^{\infty}\sum_{n=-\infty}^{\infty}\frac{1}{2m}J_{2n-1}(m\pi M)\cos([m+n-1]\pi)\cos(2m\omega_c t + [2n-1]\omega_0 t)$$

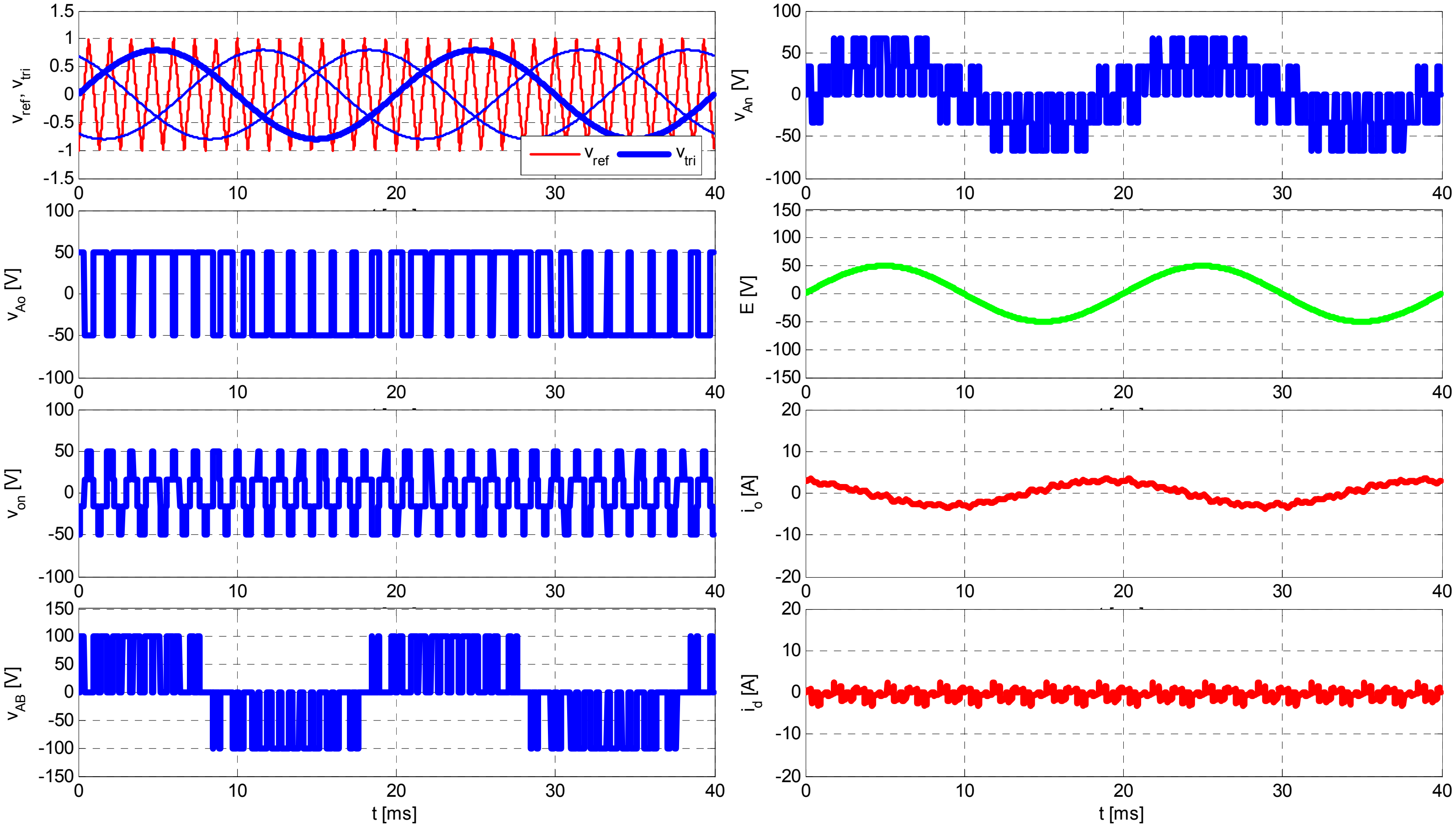
Bipolar and unipolar modulations





Due to the unipolar PWM the odd carrier and associated sideband harmonics are completely cancelled leaving only odd sideband harmonics (2n-1) terms and even (2m) carrier groups.

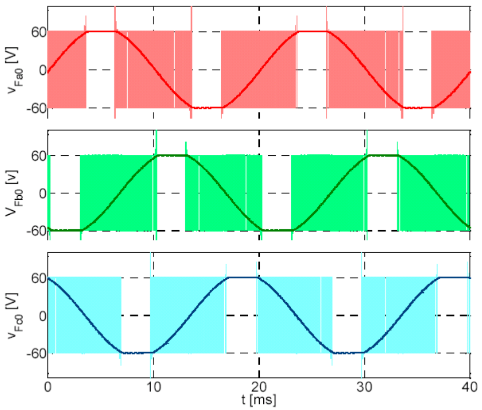
Three-phase modulation techniques

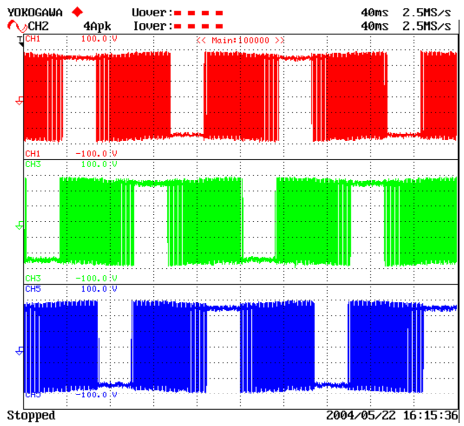


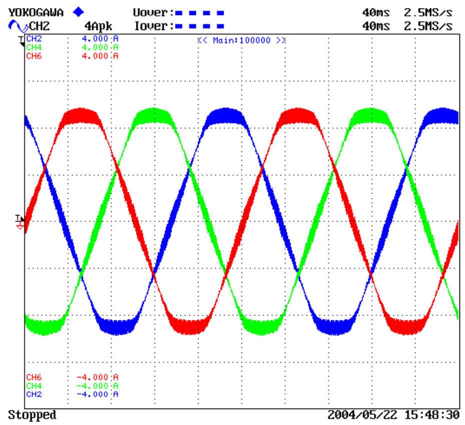
The basic three-phase modulation is obtained applying a bipolar modulation to each of the three legs of the converter. The modulating signals have to be 120 deg displaced. The phase-to-phase voltages are three levels PWM signals that do not contain triple harmonics. If the carrier frequency is chosen as multiple of three, the harmonics at the carrier frequency and at its multiples are absent.

Extending the linear range (m=1,1)

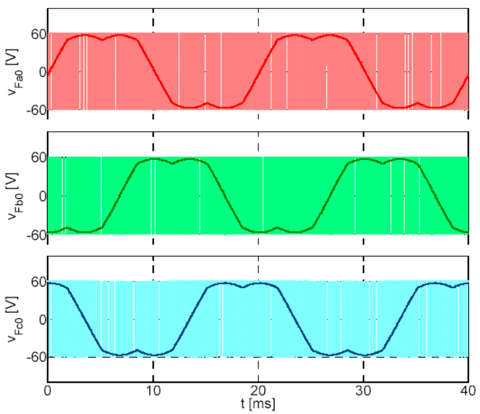
SPWM

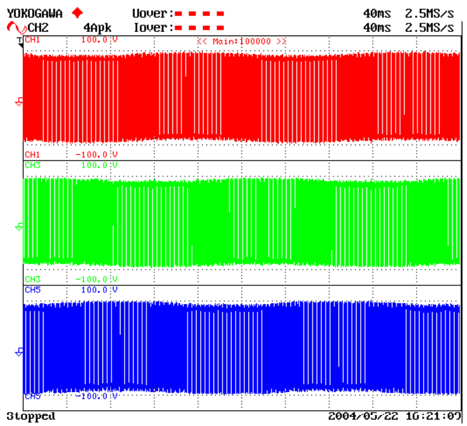


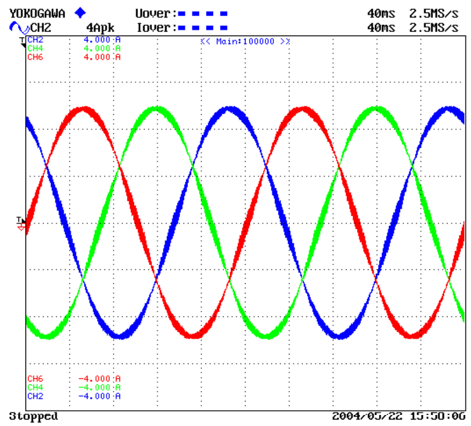




SVM



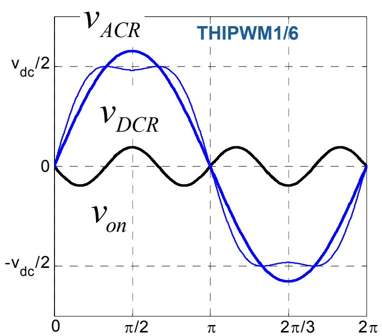




Three-phase continuous modulation techniques

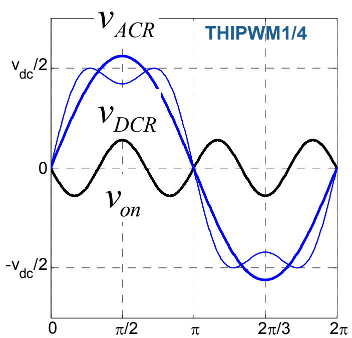
Continuous modulations

- Sinusoidal PWM with Third Harmonic Injected – THIPWM. If the third harmonic has amplitude 25% of the fundamental the minimum current harmonic content is achieved; if the third harmonic is 17% of the fundamental the maximal linear range is obtained
- Suboptimum modulation (subopt). A triangular signal is added to the modulating signal. In case the amplitude of the triangular signal is 25% of the fundamental the modulation corresponds to the Space Vector Modulation (SVPWM) with symmetrical placement of zero vectors in sampling time



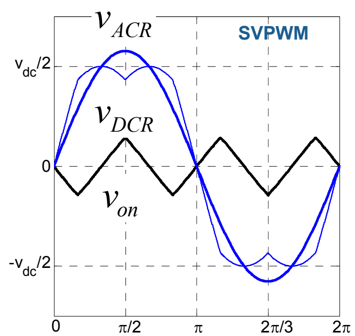
Triple harmonic injection 1/6

Maximum linear range



Triple harmonic injection 1/4

Minimum current harmonics



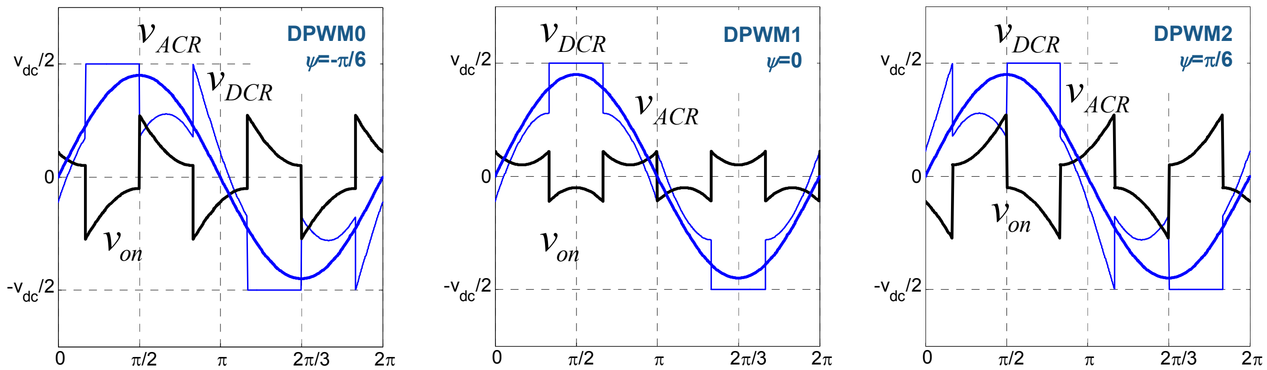
Space vector - Triangular 1/4

Maximum linear range and almost optimal current harmonics

Three-phase discontinuous modulation techniques

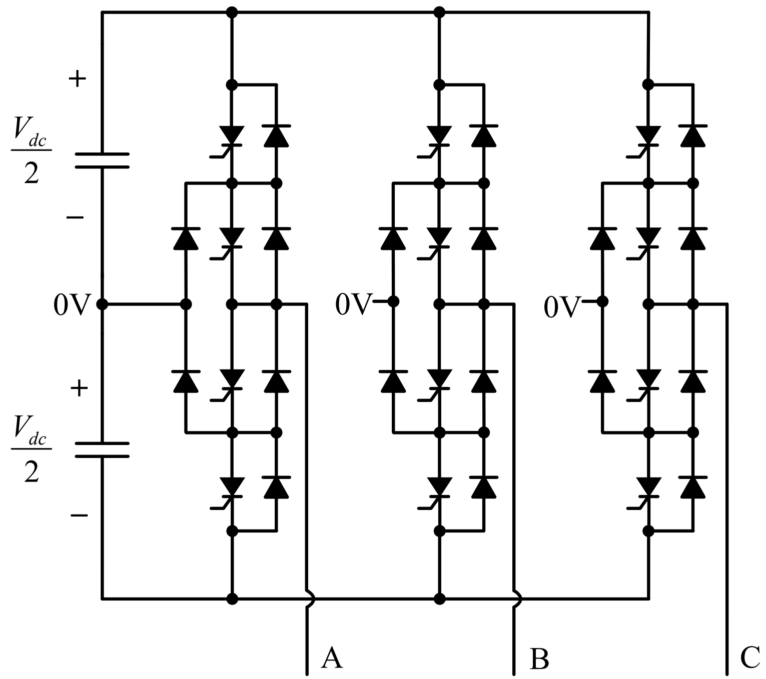
The discontinuous modulations formed by unmodulated 60 deg segments in order to decrease the switching losses

- Symmetrical flat top modulation, also called DPWM1
- Asymmetrical shifted right flat top modulation, also called DPWM2
- Asymmetrical shifted left flat top modulation, also called DPWM0



Multilevel converters and modulation techniques

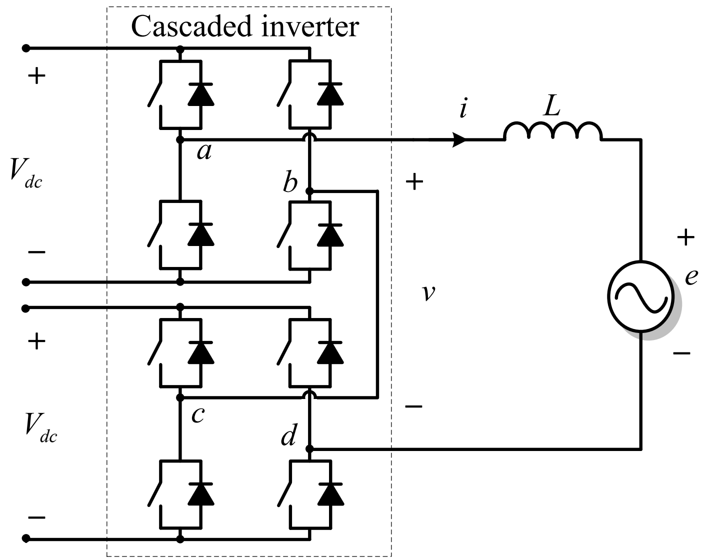
- Wind turbine systems: high power -> 5 MW Alstom converter
- Photovoltaic systems: many dc-links for a transformerless solution

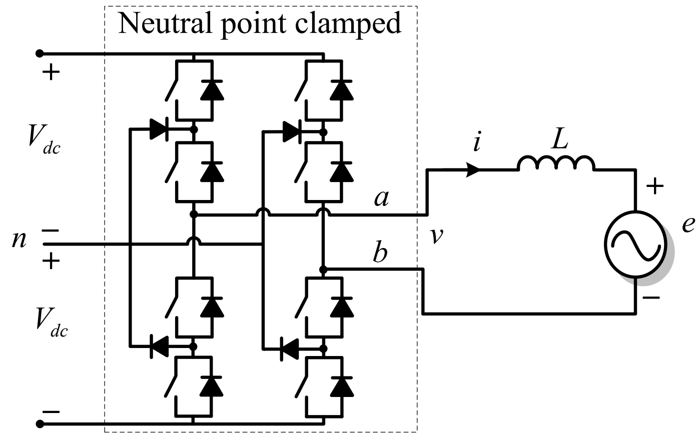


Different possibilities:

- Alternative phase opposition (APOD) where carriers in adjacent bands are phase shifted by 180 deg
- Phase opposition disposition (POD), where the carriers above the reference zero point are out of phase with those below zero by 180 deg
- Phase disposition (PD), where all the carriers are in phase across all bands

Multilevel converters and modulation techniques



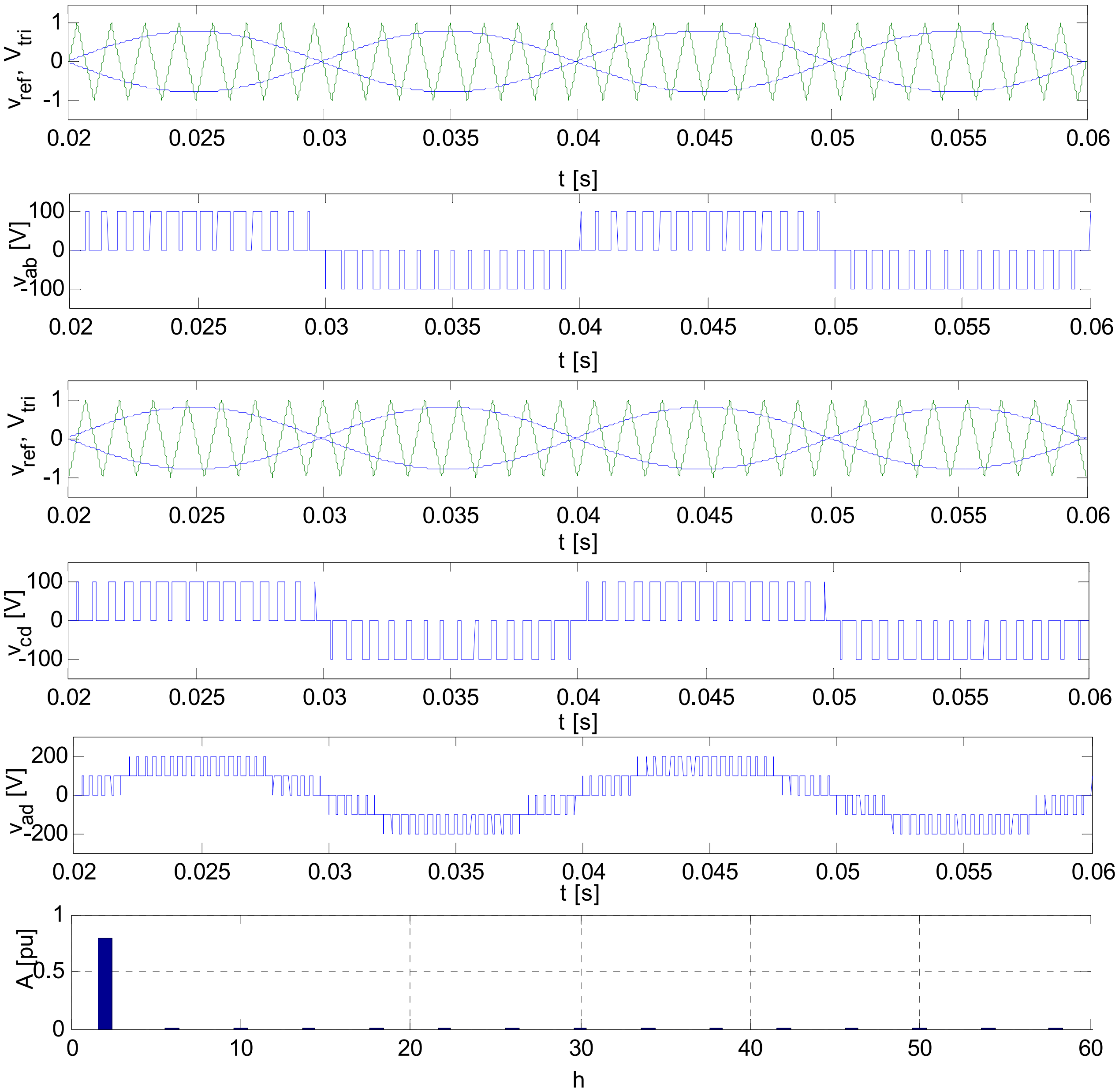


$$v(t) = NV_{dc}M\cos(\omega_0 t) +$$

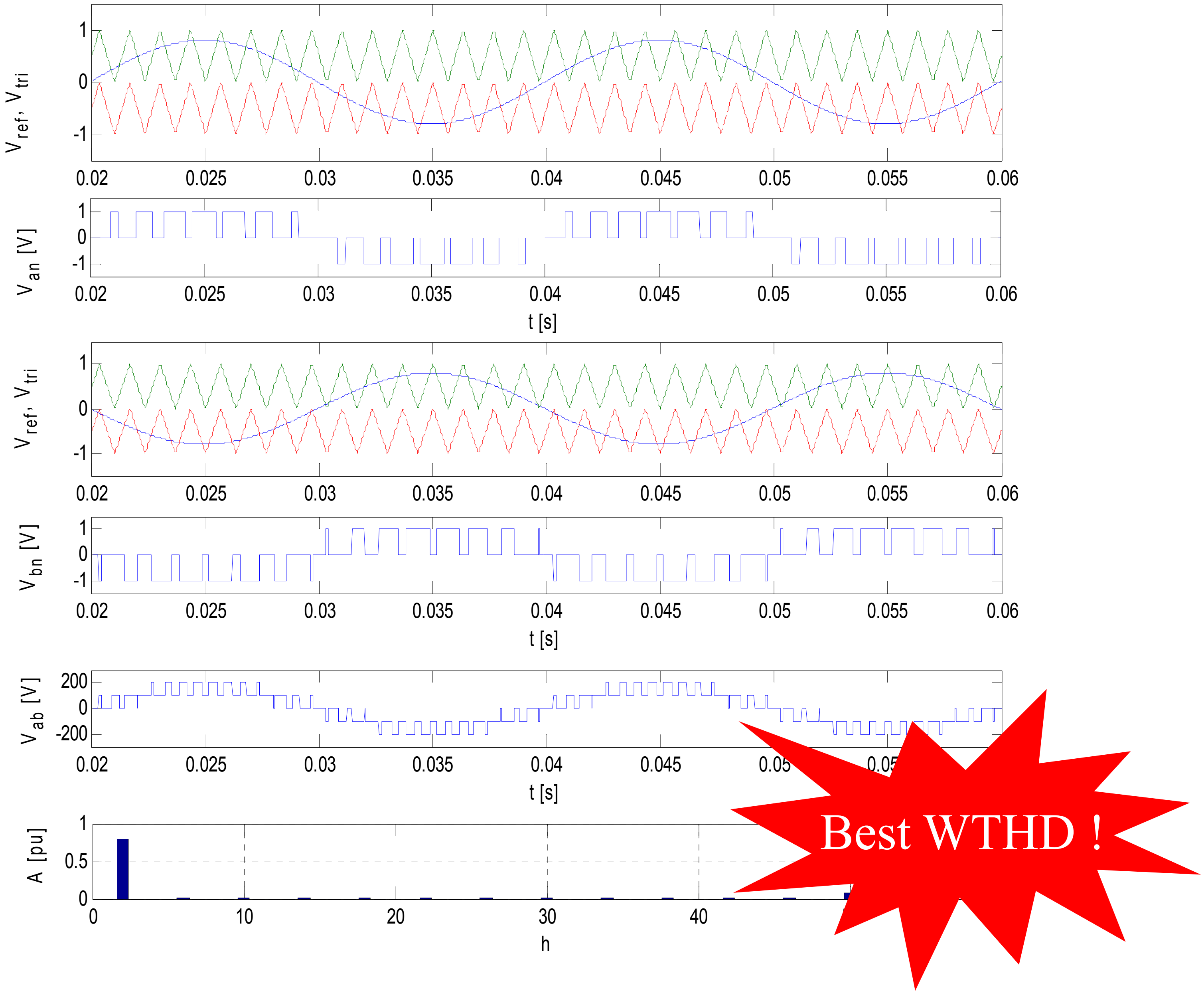
$$+\frac{4V_{dc}}{\pi}\sum_{m=1}^{\infty}\sum_{n=-\infty}^{\infty}\frac{1}{2m}J_{2n-1}(m\pi M)\cos([m+n-1]\pi)\sum_{i=1}^{N}\cos(2m\omega_c t + [2n-1]\omega_0 t + 2m\theta_i)$$

- Carrier shifting $\qquad \theta_i = \frac{(i-1)\pi}{N} \qquad \forall m \neq kN, k = 1, 2, 3\ldots$

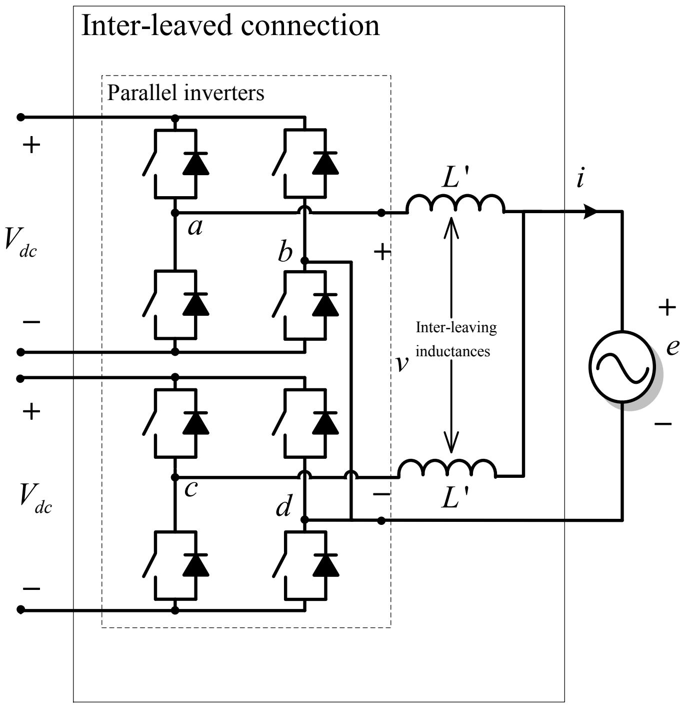
Carrier shifting



PD Modulation for NPC



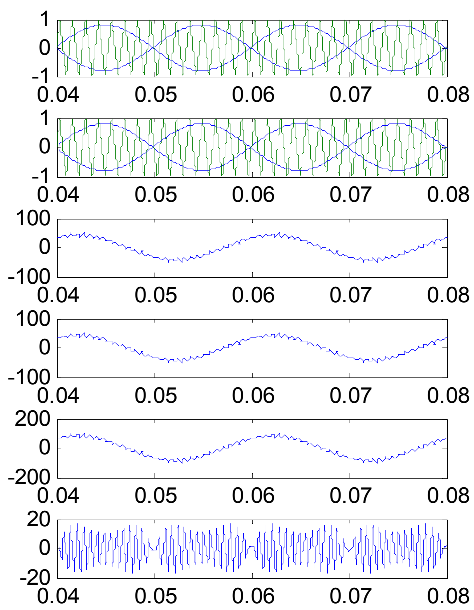
Inter-leaved modulation



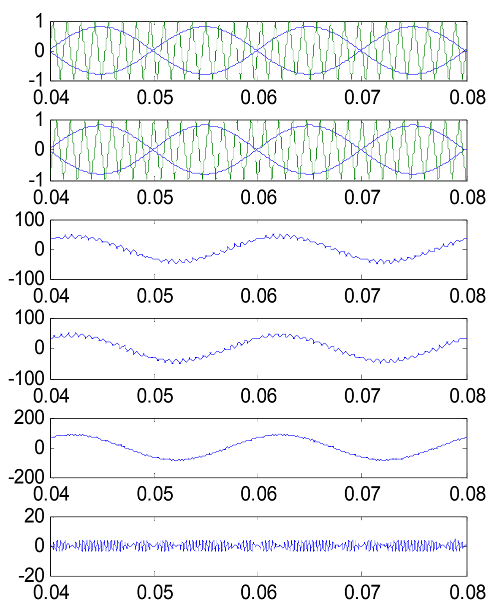
Shifting the carrier signals (interleaving modulation) reduces the harmonic content in the current with the same principle that leads cascaded converters to have a reduced harmonic content.

However here it is the current to have a reduced harmonic content.

Without interleaved



With interleaved



Operating limits of the grid converter

Power processed by the converter

$$P = 3E\frac{\sqrt{\left(M^2 V_{dc}^2/8\right) - E^2}}{\omega L}$$

This means that the higher the dc voltage and the smaller the inductance, the higher the power rating of the converter.

Current sharing ratio transistor/diodes

$$M\cos\delta = \frac{2}{\sqrt{3}k_{boost}}$$

The transistors are conducting about 93% of the current while the load on the diodes is very low. In this situation the load on the transistors is slightly higher than for normal inverter operation on an induction motor. This is due to the small displacement angle of the grid inverter compared to a typical phase angle of the stator current of an induction motor.

Conclusions

- The PR uses Generalized Integrators (GI) that are double integrators achieving very high gain in a narrow frequency band centered on the resonant frequency and almost null outside
- This makes the PR controller to act as a notch filter at the resonance frequency and thus it can track a sinusoidal reference without having to increase the switching frequency or adopting a high gain, as it is the case for the classical PI controller
- PI adopted in a rotating frame achieves similar results, it is equivalent to the use of three PR's one for each phase
- Also single phase use of PI in a *dq* frame is feasible
- Dead-beat controller can compensate current error in two samples but it is affected by PWM limits and parameters mismatches
- Dead-beat controller is faster in limiting overcurrent during faults
- A review of modulation techniques has been given