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PROPORTIONAL-RESONANT (PR) CONTROLLERS AND FILTERS FOR GRID-CONNECTED VOLTAGE-SOURCE CONVERTERS

Ankita Singh*1 and Prof S.S.Khule2

*1Student of (M.E) Electrical Engineering, Matoshri College of Engineering & Research Centre Nashik, India

2Assistant Prof & Head/Electrical Matoshri College of Engineering & Research Centre Nashik, India

ABSTRACT

The recently introduced proportional-resonant (PR) controllers and filters, and their quality for current/voltage management of grid-connected converters, are represented in these paper. Victimization the PR controllers, the device reference following performance will be increased and antecedently best-known shortcomings related to standard PI controllers may be relieved. These shortcomings embody steady-state errors in single-phase systems and also the want for synchronous d–q transformation in three-phase systems. Supported similar management theory, PR filters also can be used for generating the harmonic command reference exactly in a full of life power filter, particularly for single-phase systems, wherever d–q transformation theory isn't directly applicable. Another advantage related to the PR controllers and filters is that the chance of implementing selective harmonic compensation while not requiring excessive process resources.

Keywords: *Proportional-Resonant (PR) controllers, grid-connected voltage-source converters etc*

1. INTRODUCTION

Most of those systems embody a grid-connected voltage-source convertor whose practicality is to synchronies and transfer the variable created power over to the grid. This paper is describing the recently introduced proportional-resonant (PR) and their quality for grid-connected converters current control. energy technologies, like wind and solar primarily based energy generation systems, are receiving national and worldwide attention due to the rising rate of consumption of nuclear and fossil fuels Another feature of the adopted convertor is that it's typically pulse-width modulated (PWM) at a high change frequency and is either current- or voltage-controlled employing a designated linear or nonlinear management formula. The deciding criterion once choosing the suitable management theme sometimes involves an optimum exchange between value, complexity and wave form quality required for meeting (for example) new power quality standards for distributed generation in low-voltage grids, this controller will have a big impact on the standard of this equipped to the grid by the PV electrical converter, and thus it's necessary that the controller provides a top quality curving output with smallest distortion to avoid making harmonics. 2 controllers that are employed in current controlled PV inverters are the PI controller with the grid voltage feed-forward and also the PR controller. Exploring the simplicity of PI controllers and to boost their overall performance, several variations are planned within the literature together with the addition of a grid voltage feed forward path, multiple-state feedback and increasing the proportional gain.

Generally, these variations will expand the PI controller information measure however; sadly, they additionally push the systems towards there. In brief, the fundamental practicality of the PR controller is to introduce associate infinite gain at a specific resonant frequency for eliminating steady state error at that frequency, Associate in Nursing is so conceptually almost like a measuring instrument whose infinite DC gain forces the DC steady-state error to zero. The resonant portion of the PR controller will so be viewed as a generalized AC flexibility of calibration the resonant frequency, makes an attempt at victimization multiple PR controllers for by selection compensating low-order harmonics have additionally been for three-phase active power filters, for three-phase uninterruptible power provides (UPS) and for single part electrical phenomenon (PV) inverters. The PR controller provides gain at a definite frequency (resonant frequency) an almost no gain exists at the opposite frequencies. In this paper the PR controllers are introduced and the performances are represented victimization frequency analysis. Then, typical management methods for each single-phase and three-phase RES victimization PI and PR are represented and compared in terms of performance and simple implementation. From the read purpose that electronic power converters can notice increasing grid-interfaced applications either as inverters process DC energy from RES for grid injection or as rectifiers learning grid energy for various load usages, this paper aims to produce a comprehensive reference for readers on the combination of PR controllers and filters to grid-connected converters for enhancing their trailing performances. To begin, the paper reviews frequency domain derivation of the perfect and non-ideal PR controllers and filters, and discusses their similarities as compared to classical PI management. Limits for the present harmonics. Typically, PI controllers with grid voltage feed forward are utilized in order to manage the present of grid-connected converters.

2. PR CONTROL AND FILTERING DERIVATION

The PR controller provides gain at a certain frequency (resonant frequency) and almost no gain exists at the other frequencies. The transfer functions of single- and three-phase PR controllers and filters can be derived using internal model control, modified state transformation or frequency-domain.

2.1 Practical Application to single-phase PR transfer functions

Single-phase grid converters are usually utilized in applications like residential RES (typically PV or FC systems), residential UPS, etc. Fig. 2.1.1 Single-phase equivalent representations of PR and synchronous PI controller typical RES is represented wherever the active and reactive power is controlled within the outer loop. For single-phase PI management, the popularly used synchronous d–q transformation cannot be applied directly, and also the nearest equivalence developed thus far is to multiply the feedback error e(t), in turn, by sine and cosine functions typically synchronous with the grid voltage employing a section locked- loop (PLL), as shown in Fig. This achieves identical result of reworking the element at the chosen frequency to DC; deed all different parts as AC quantities

$$e(t) = E_1 \cos(\omega t + \theta_1) + E_3 \cos(3\omega t + \theta_3)......(1)$$

Where ω, θ_1 and θ_3 represent the fundamental angular frequency, fundamental and third harmonic phase shifts respectively. Multiplying this with cos (ωt) and sin (ωt) gives, respectively:

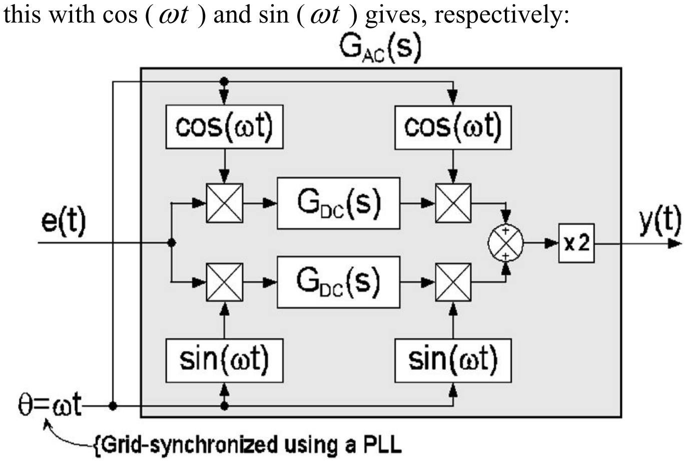

Fig2.1.1 Single-phase equivalent representations of PR and synchronous PI controllers

Here $e_s(t)$

$$e_c(t) = E_1/2\left\{\cos(\theta_1) + \cos(2\omega t + \theta_1)\right\} + E_{3/2} + \left\{\cos(2\omega t + \theta_3)\cos(4\omega t + \theta_{3)}\right\}....(2)$$

$$e_s = E_1/2\left\{\sin(-\theta_1) + \sin(2\omega t + \theta_1)\right\} + E_{3/2}\left\{\sin(-2\omega t - \theta_3) + \sin(4\omega t + \theta_3)\right\}$$

It is determined that the term currently seems as DC quantities cos(θ_1) and sin(θ_1)

The solely complication with this equivalent single-phase conversion is that the chosen frequency element not only seems as a DC amount within the synchronous frame, it additionally contributes to harmonic terms at a frequency of 2 ω (this is not like 3 part synchronous d–q conversion wherever the chosen frequency element contributes solely towards the DC term). All the same, passing $e_c(t)$ and $e_s(t)$ through integral blocks would still force the elemental error amplitude E_1 to zero, caused by the infinite gain of the integral blocks

$$G_{AC}(s) = G_{DC}(s - j\omega) + G_{DC}(s + j\omega).....(3)$$

Where $G_{AC}(s)$ represents the equivalent stationary frame transfer function. Therefore, for the ideal and non-ideal integrators of $G_{DC}(s) = K_i / s$ =s and ($G_{DC}(s) = K_i / (1 + s / \omega_c)$) ((Ki and ω_c << ω represent controller gain and cutoff frequency respectively) represent controller gain and cutoff frequency respectively), the derived generalized AC integrators $G_{AC}(s)$ are expressed as:

$$G_{AC}(s) = Y(s) / E(s) = 2K_i s / s^2 + \omega^2 .....(4)$$

$$G_{AC}(s) = Y(s) / E(s) = 2K_i(\omega_C s + \omega^2{}_c) / s^2 + 2\omega_c s + (\omega_c{}^2 + \omega^2) \approx 2K_i \omega_c s / s^2 + 2\omega_c s + \omega^2 ....(5)$$

$$G_h(s) = \sum_{h=3,5,7} 2K_{ih} s / s^2 + (h\omega)^2 ....(6)$$

$$G_h(s) = \sum_{h=3,5,7} 2K_{ih} \omega_c s / s^2 + 2\omega_c s + (h\omega_c)^2 ......(7)$$

Besides single frequency compensation, selective harmonic compensation also can be achieved by cascading many resonant blocks tuned to resonate at the required low-order harmonic frequencies to be paid for. As AN example, the transfer functions of a perfect and a non-ideal harmonic compensator (HC) designed to atone for the third, fifth and seventh harmonics (as area unit|they're} the foremost distinguished harmonics in a very typical current spectrum) are given as where h is that the harmonic order to be compensated for and K_{ih} represents the individual resonant gain, that should be tuned comparatively high (but among stability limit) for minimizing the steady-state error. A motivating feature of the HC is that it doesn't have an effect on the dynamics of the basic PR controller, because it compensates just for frequencies that square measure terribly near the chosen resonant frequencies

2.2 Derivation of three-phase PR transfer functions for three-phase:

Three-phase grid converters are commonly used in applications like RES (small wind or water turbines, high-power PV plant, etc), residential UPS, active filters, etc. A typical RES is depicted where the active and reactive powers are controlled in the outer loop.

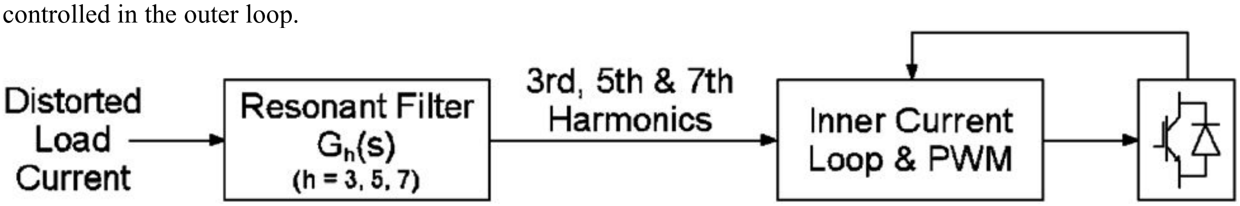

Fig 2.2.1 Resonant filter for filtering 3rd, 5th and 7th harmonics

An alternative simpler method of implementation is therefore desired and can be derived by

Inverse transformation of the synchronous controller back to the stationary $\alpha\beta$ frame $G_{dq}(s) \to G_{\alpha\beta}(s)$. The inverse transformation can be performed by using the following2*2 matrix:

$$G_{\alpha\beta}(s) = 1/2 \begin{pmatrix} G_{dq1} + G_{dq2} & jG_{dq1} - jG_{dq2} \\ -jG_{dq1} + jG_{dq2} & G_{dq1} + G_{dq2} \end{pmatrix} ......(8)$$

$$G_{dq1} = G_{dq}(s + j\omega)$$

$$G_{dq2} = G_{dq}(s - j\omega)$$

Given that $G_{dq}(s) = K_i / s and G_{dq}(s) = K_i(1 + (s / \omega_c))$, the equivalent controllers in the stationary frame for compensating for positive-sequence feedback error are therefore expressed as:

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$$G^{+}_{\alpha\beta}(s) = \frac{1}{2}\begin{pmatrix} \frac{2K_i s}{s^2+\omega^2} & \frac{2K_i \omega}{s^2+\omega^2} \\ \frac{-2K_i \omega}{s^2+\omega^2} & \frac{2K_i s}{s^2+\omega^2} \end{pmatrix} ....(9)$$

$$G_{\alpha\beta}{}^{+}(s) \simeq \frac{1}{2}\begin{pmatrix} \frac{2K_i \omega_c s}{s^2+2\omega_c s+\omega^2} & \frac{2K_i \omega_c \omega}{s^2+2\omega_c s+\omega^2} \\ \frac{2K_i \omega_c \omega}{s^2+2\omega_c s+\omega^2} & \frac{2K_i \omega_c s}{s^2+2\omega_c s+\omega^2} \end{pmatrix} ....(10)$$

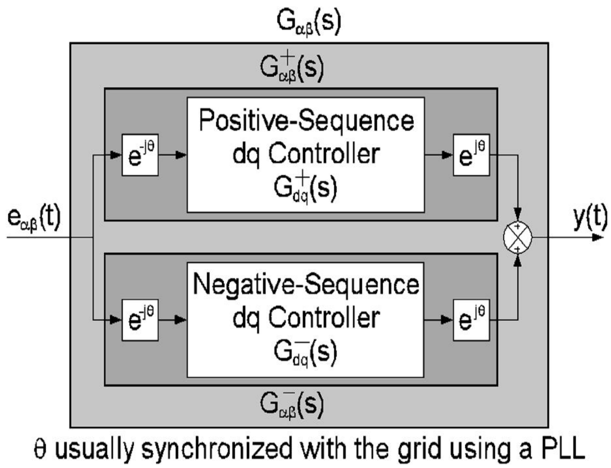

Fig.2.2.2 Three-phase equivalent representations of PR and synchronous PI controllers considering both positive- and negative-sequence components

Three-phase equivalent representations of PR and synchronous PI controllers considering both positive- and negative-sequence components

Similarly, for compensating for negative sequence feedback error, the required transfer functions are expressed as :

$$G_{\alpha\beta}{}^{-}(s) = 1/2\begin{pmatrix} 2K_i s / s^2+\omega^2 & -2K_i \omega / s^2+\omega^2 \\ 2K_i \omega s / s^2+\omega^2 & 2K_i s / s^2+\omega^2 \end{pmatrix} ....(11)$$

$$G_{\alpha\beta}{}^{-}(s) \simeq 1/2\begin{pmatrix} 2K_i \omega_c s / s^2+2\omega_c s+\omega^2 & -2K_i \omega_c \omega / s^2+2\omega_c s+\omega^2 \\ 2K_i \omega_c s+\omega^2 & 2K_i \omega_c s / s^2+2\omega_c s+\omega^2 \end{pmatrix} ....(12)$$

Comparing (9) and (10) with (11) and (12), it is noted that the diagonal terms of $G_{\alpha\beta}{}^{+}(s) and G_{\alpha\beta}{}^{-}(s)$ are identical, but their non-diagonal terms are opposite in polarity. This inversion of polarity can be viewed as equivalent to the reversal of rotating direction between the positive- and negative-sequence synchronous frames.

$$G_{\alpha\beta}(s) = 1/2\begin{pmatrix} 2K_i s / s^2 + \omega^2 & 0 \\ 0 & 2K_i s / s^2 + \omega 2 \end{pmatrix} \text{........(13)}$$

$$G_{\alpha\beta}(s) \simeq 1/2\begin{pmatrix} 2K_i\omega_c s / s^2 + 2\omega_c s + \omega^2 & 0 \\ 0 & 2K_i\omega_c s / s^2 + 2\omega_c s + \omega^2 \end{pmatrix} \text{.......(14)}$$

Therefore, the theoretical knowledge described earlier for single-phase PR control is equally applicable to the three-phase functions expressed in (13) and (14).

3. IMPLEMENTATION OF RESONANT CONTROLLERS

The resonant transfer functions in (4) and (5) (similarly in (13) and (14)) can be implemented using analogue integrated circuits (IC) or a digital signal processor (DSP), with the latter being more popular

3.1 Analogue implementation

The rational function in (4) can be rewritten as
The function in (5) can be rewritten

$$Y(s)/E(s) = 2K_i\omega_c s / s^2 + 2\omega_c s + \omega^2 \Rightarrow \begin{cases} Y(s) = 1/s[2K_i\omega_c E(s) - V(s) - V_2(s)] \\ V_1(s) = 2\omega_C Y(s) \\ V_2(s) = 1/s\omega^2 Y(s) \end{cases} \text{(16)}$$

It may be deduced that the resonant operate will be physically enforced victimization op-amp integrators and inverting/non-inverting gain amplifiers. Note additionally that, whereas implementing (15), parasitic resistance and alternative second-order imperfections would cause it to degenerate into (16), however after all its information measure will only be tuned if further parts are additional for implementing the higher feedback path

3.2 Shift-operator digital implementation

The most commonly used digitization technique is the prewired bilinear (Tustin) transform

$$s = \frac{\omega_1}{\tan(\omega_1 T/2)}\frac{z-1}{z+1} = K_T \frac{z-1}{\tan(\omega_1 T/2)} \text{......(17)}$$

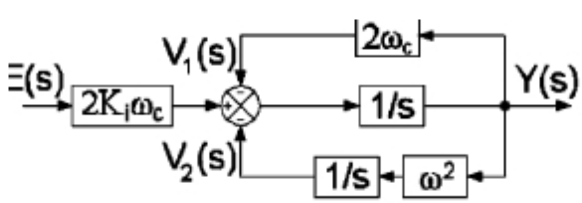

Fig.3.2.1 Decomposition of resonant block into two interlinked integrators

$$Y(z)/E(z) = a_1 z^{-1} - a_2 z / b_0 - b_1 z^{-1} + b_2 z^{-2}$$
$$a_1 = a_2 = 2K_i K_T \omega_c$$
$$b_0 = K_T{}^2 + 2K_T \omega_c + \omega^2 ....(18)$$
$$b_1 = 2K_T{}^2 - 2\omega^2$$
$$b_2 = \begin{Bmatrix} K^2{}_T + 2K_T \omega_C + \omega^2 \\ K_T{}^2 + 2K_T \omega_C + (h\omega)^2 \end{Bmatrix}$$
$$y(n) = 1/b_0 \{a_1[e(n-1) - e(n-2)] + b_1 y(n-1) - b_2 y(n-2)\} ...(19)$$

Equations (18) and (19) will equally be used for implementing the HC compensator when the specified harmonic order h is substituted. The ensuing distinction equation will handily be programmed into a floating-point DSP, however once a fixed-point DSP is employed instead, coefficients of (19) need to be normalized by multiplying them with the most number worth of the chosen word length

4. SINGLE-PHASE PV GRID-CONNECTED INVERTER

Single-phase grid inverters are normally utilized in applications like residential RES (typically PV or cell systems) and UPS. Figure eight shows a typical RES wherever the DC-link voltage, active P and reactive alphabetic character power are management led within the outer control loops (labeled as voltage controller and reference generator within the Figure). The reference current outputs of the outer loops ($i^*{}_{dd}$ and i^* _) are next half-track by an inner current loop whose output is eventually fed to a PWM modulator for switch the electrical converter

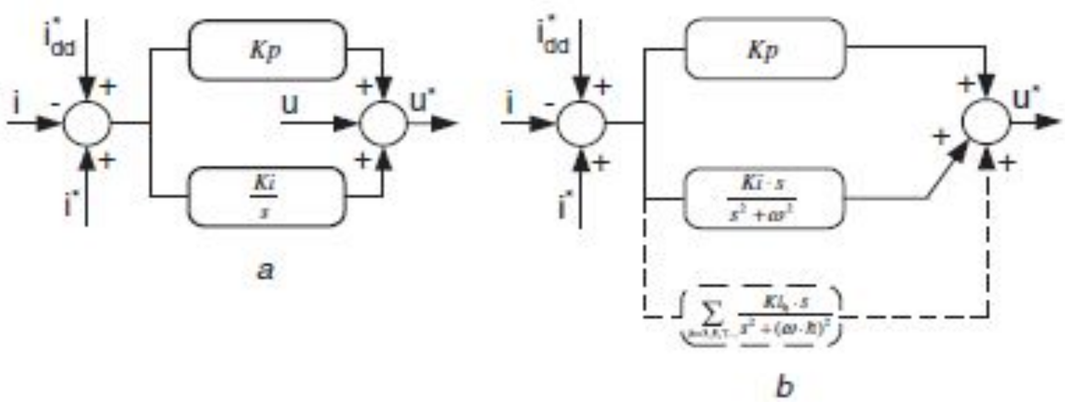

Fig4.1 Single-phase grid inverter control
a) Stationary PI control
b) Stationary PR inner current control

5. CONCLUSION

We can say PR controller can be used in any application where less steady state error is required. As PR controller has less steady error than PI Controller. Selective harmonic can be compensated by using PR filter.

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6. RESULT

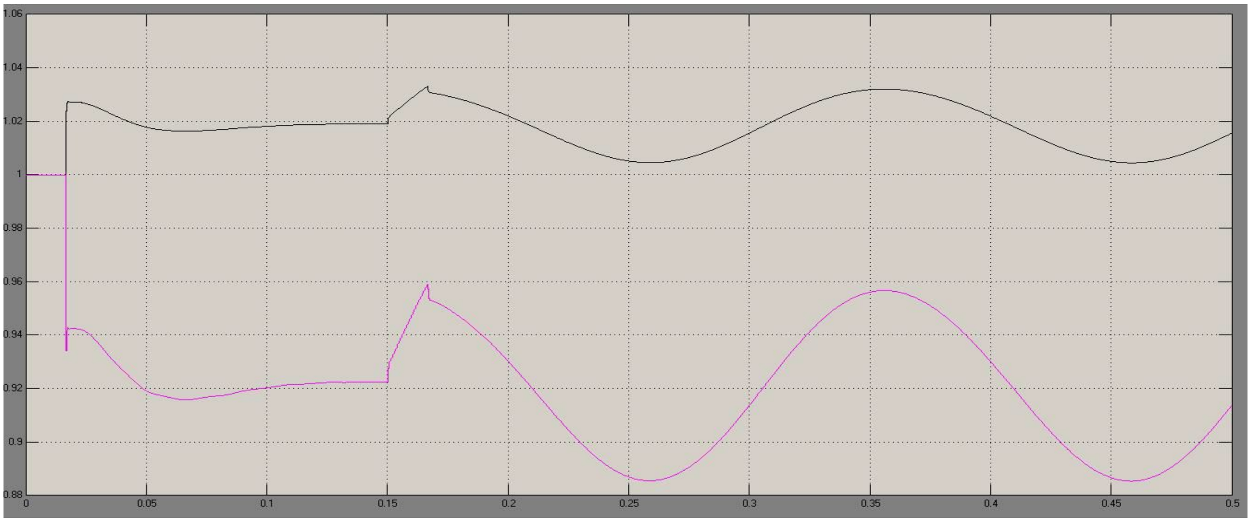

Fig6.1: With Nonlinear Load BUS 1

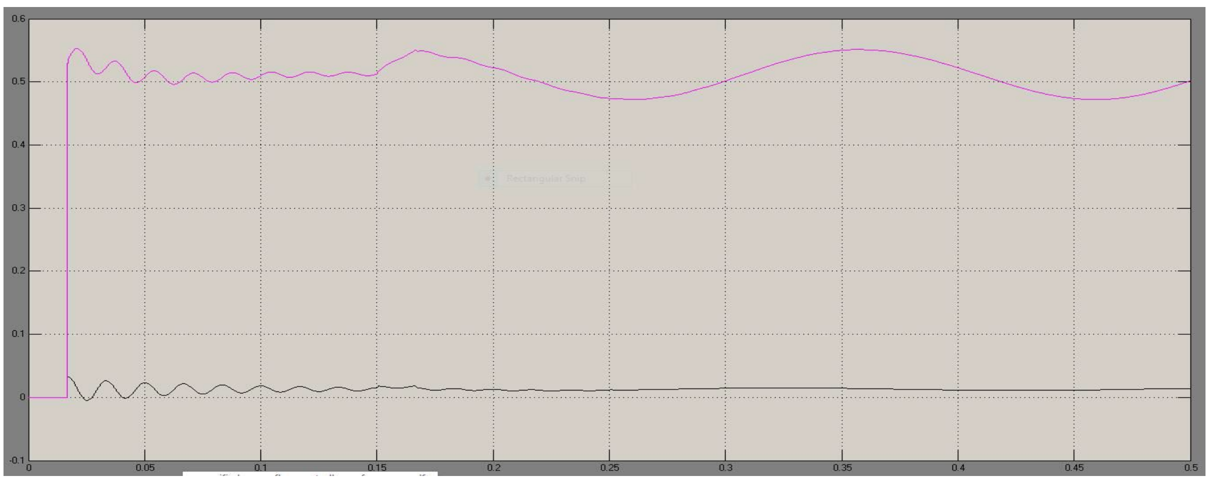

Fig6.2: With Nonlinear Load BUS 2

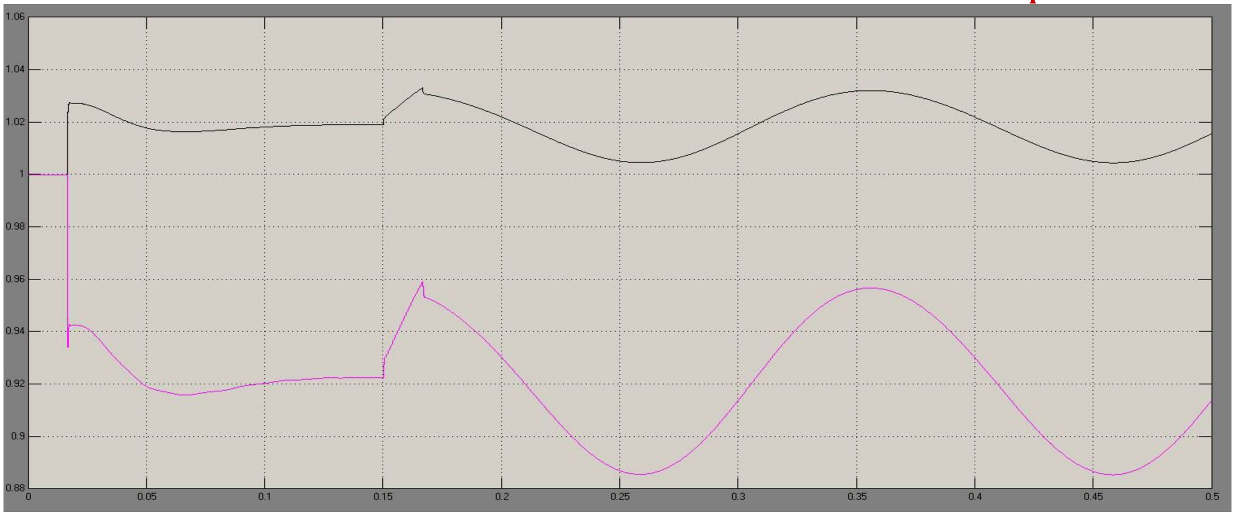

Fig 6.3 with Nonlinear Load BUS 3

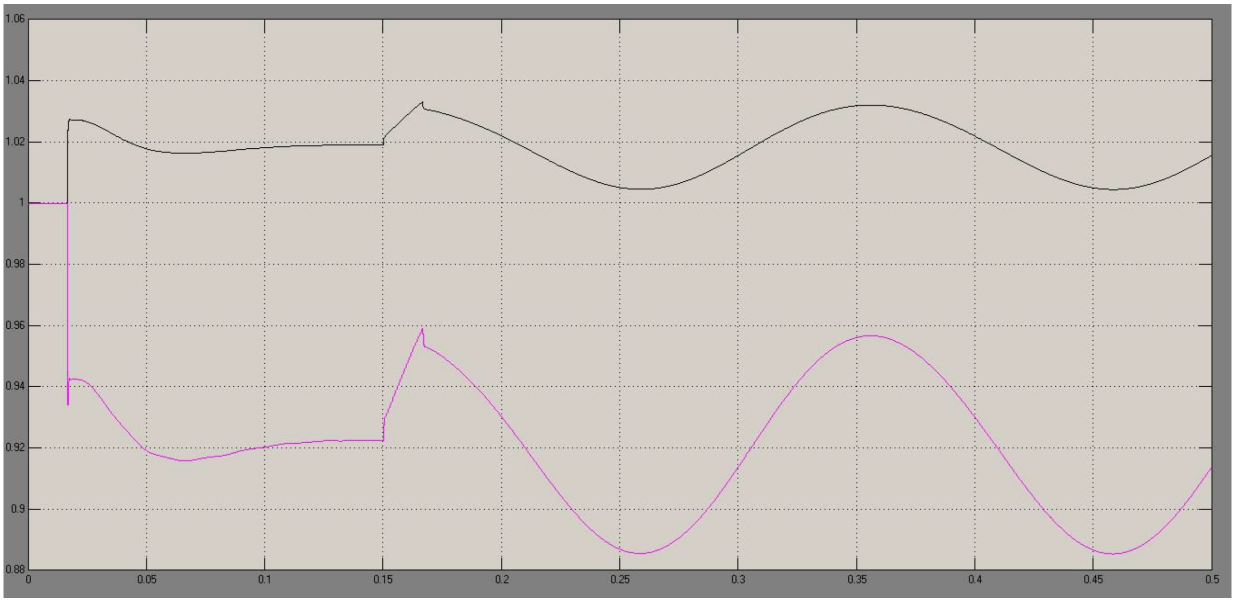

Fig 6.4: With Nonlinear Load and FAULT UPFC 1 and 2

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