Reactive Power, Unbalance and Harmonics Compensation using D–Statcom with a Dissipativity-Based Controller

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Abstract

In this paper we present a solution to the problem of reactive power compensation and harmonic compensation in the general case when *both* the source voltages and the load currents are unbalanced and contain an arbitrary number of harmonics. For the compensation purpose, a D–Statcom is connected in parallel to inject the required currents, so that from the source terminals the *same* apparent resistance is observed in all phases and at all frequencies. A controller based on the ideas of passivity theory, to which we have added adaptation to compensate for the unavoidable uncertainty in some of the parameters, is suggested as a solution. One of the major advantages of this solution compared to conventional ones is that we are able to perform precise tracking (also for high order harmonics) even in the presence of a relatively low switching frequency, i.e., in presence of an active filter with limited bandwidth. Simulation results are provided to illustrate the performance of our controller.

1 Introduction

Statcom(*Sta*tic *com*pensator) is one of the most important Flexible AC transmission system (FACTS) device [4, 6] because of its ability to regulate voltages in transmission lines, to improve transient stability and to compensate variable reactive power. These devices are also interesting at the distribution level, where fast compensation of large, fluctuating industrial loads (such as electric arc furnaces and rolling mills) can be achieved, substantially improving upon the performance achievable with conventional thyristor-based converters. For low-medium power applications at the distribution level, such devices (sometimes called D-Statcom (*D*istribution Statcom) or also active filters) are able also to perform compensation for the load current harmonics produced by distorting load, to comply with harmonic standards such as IEEE 519.

Design and control of Statcom's and active filters have been investigated in several papers [1, 2, 10, 5], covering different issues such as unbalanced and harmonic control, compensation strategies, modulation techniques, and so on. In order to overcome the problem of the control delay [3], particularly relevant in high-power applications with low switching frequency, this paper investigates a solution to the problem of reactive power compensation and harmonic compensation in the case when *both* the source voltages and the load current have unbalanced and harmonic components. For this purpose, a D–Statcom is connected in parallel to inject the "necessary" currents, so that from the source terminals the *same* apparent resistance is observed in all phases and at all frequencies. In other words, the current provided by the source will be proportional to the source voltages so as to achieve unity power factor for the compensated load. In order to achieve this objective an energy storage capacitor is needed on the dc-side of D–Statcom's since also some amount of higher harmonic real (active) power is exchanged between the compensator and the load. The design of such "necessary" currents constitutes a key part of the controller design.

In this paper we propose a controller based on the ideas of passivity theory [8] to which we have added adaptation to compensate for the unavoidable uncertainty in some of the parameters. Using a proper estimation scheme for the evaluation of the derivative of current reference, a precise tracking and compensation is achieved even in presence of inverters with limited control bandwidth. Simulation results are provided to assess the performance of our controller.

2 System Model

The dynamics of a three-phase three-wire D–Statcom shown in Fig. 1 is described by the following model

$$L\frac{d}{dt}i = -ri - \frac{v_C}{2}Bu + v_S \tag{1}$$

$$C\frac{d}{dt}v_C = \frac{1}{2}u^T i - \frac{v_C}{R} \tag{2}$$

and its connection to the line source is defined in terms of currents by

$$i_s = i_L + i$$

where L is the inductance of each input filter winding and r represents its parasitic resistance. For the sake of simplicity, we assume that r and L are identical on each branch, i.e., the input filter has balanced parameters. Without this assumption, bulky matrix computations would be involved which, however, would give similar results. On the dc-side of Fig. 1, C is the output capacitor, while R is only representing inverter switching and other losses, here simply collected as a resistive element. All parameters

(L,r,C,R) in (1) are considered unknown constants; $i(t) = [i_1(t),\, i_2(t),\, i_3(t)]^T$ is the vector of compensating currents; $i_L(t) = [i_{L1}(t),\, i_{L2}(t),\, i_{L3}(t)]^T$ is the vector of currents produced by the load; $v_S(t) = [v_{S1}(t),\, v_{S2}(t),\, v_{S3}(t)]^T$ is the vector of voltages coming from the source; $u(t) = [u_1(t),\, u_2(t),\, u_3(t)]^T$ represents the vector of switch positions acting as the control input vector, $u_i \in \{-1, 1\}$, that is, if the upper switch in Fig. 1 is *on* (the lower switch *off*) then $u_1 = 1$ otherwise $u_1 = -1$, the same applies for the other two phases; B is a matrix defined by

$$B = \frac{1}{3}\begin{bmatrix} 2 & -1 & -1 \\ -1 & 2 & -1 \\ -1 & -1 & 2 \end{bmatrix}$$

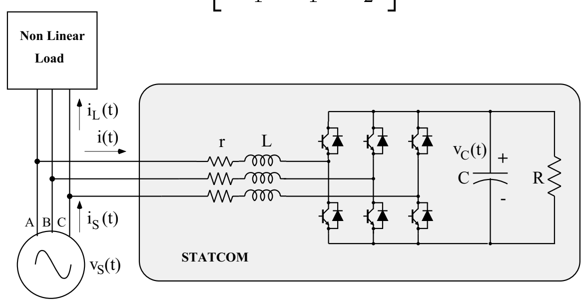

Figure 1: Shunt connection of the D–Statcom to the line

The system is mainly designed to cancel/compensate the reactive power, unbalanced and harmonics contents in the distribution line caused by reactive and distorting loads. For simplicity, we model the distribution lines with voltage sources and distorting loads with currents sources, even if their dynamics may add other control problems in some specific cases [9]. The compensation of reactive, unbalanced and harmonics currents can be performed by injecting into the line the compensating currents $i_i(t)\,(i \in \{1, 2, 3\})$ which force the currents $i_{Si}(t)$ on each line to be proportional to the respective source line voltage $v_{Si}(t)$, so as to achieve unity power factor. To accomplish this, the capacitor C is charged to a certain constant value V_d, which is the dc component of voltage v_C during steady-state conditions. Finally, in order to analyze generic conditions, both the load currents and the source voltages are allowed to be unbalanced, and thus amplitudes and phase angles could take arbitrary values.

Since our three-phase system is without the neutral wire, model (1)-(2) is firstly expressed into conventional stationary $\alpha\beta$ coordinates, using following $3 \to 2$ transformation

$$\begin{bmatrix} x_\alpha \\ x_\beta \end{bmatrix} = \underbrace{\frac{2}{3}\begin{bmatrix} 1 & -\frac{1}{2} & -\frac{1}{2} \\ 0 & \frac{\sqrt{3}}{2} & -\frac{\sqrt{3}}{2} \end{bmatrix}}_{\mathcal{A}} \begin{bmatrix} x_1 \\ x_2 \\ x_3 \end{bmatrix} \tag{3}$$

whose inverse transformation is defined by $\mathcal{A}^{-1} \stackrel{\triangle}{=} \frac{3}{2}\mathcal{A}^T$.

The transformed model is therefore given by

$$L\frac{d}{dt}i_{\alpha\beta} = -ri_{\alpha\beta} - \frac{1}{2}v_C u_{\alpha\beta} + v_{S\alpha\beta} \tag{4}$$

$$C\frac{d}{dt}v_C = \frac{3}{4}u_{\alpha\beta}^T i_{\alpha\beta} - \frac{v_C}{R} \tag{5}$$

$$i_{S\alpha\beta} = i_{L\alpha\beta} + i_{\alpha\beta}$$

where we have used the fact that $\mathcal{A}B\mathcal{A}^{-1} = \mathcal{I}_2$.

The control objective consists in injecting the compensating currents so that from the source terminals the *same* apparent resistance is observed in all phases and at all frequencies. Thus, the currents provided by the source is proportional to the voltages generated by the source.

$$i^*_{S\alpha\beta} = g_T v_{S\alpha\beta} \tag{6}$$

with g_T an equivalent conductance determined by power balance conditions.

3 Inner Control Loop

The control problem presented above can be restated as a tracking problem on the current state $i_{\alpha\beta}$, being the current reference computed as

$$i^*_{\alpha\beta} = i^*_{S\alpha\beta} - i_{L\alpha\beta} = g_T v_{S\alpha\beta} - i_{L\alpha\beta} \tag{7}$$

The compensator contains no net energy sources, so it has to preserve the dc component of the original active power after compensation, besides inverter losses. Thus, we propose to compute g_T as follows

$$g_T \stackrel{\triangle}{=} g'_T + g \;\;,\;\; g'_T \stackrel{\triangle}{=} \frac{\langle v^T_{S\alpha\beta} i_{L\alpha\beta}\rangle_0}{\langle v^T_{S\alpha\beta} v_{S\alpha\beta}\rangle_0} \tag{8}$$

where g'_T represents the equivalent conductance of the load, g the D–Statcom losses in r and R, and $\langle y \rangle_k$ is the k^{th} harmonic of the (scalar) waveform $y(\cdot)$.

Let $i^*_{\alpha\beta}$ denote the current reference with time derivative $\frac{d}{dt}i^*_{\alpha\beta}$. Assuming that v_C is bounded away from zero, a simple dissipative controller that forces the current $i_{\alpha\beta}$ to follow $i^*_{\alpha\beta}$ asymptotically is

$$u_{\alpha\beta} = \frac{2}{v_C}\left[-\hat{r}i^*_{\alpha\beta} + v_{S\alpha\beta} + K_1\tilde{i}_{\alpha\beta} - \hat{L}\frac{d}{dt}i^*_{\alpha\beta}\right] \tag{9}$$

where $\tilde{i}_{\alpha\beta} \stackrel{\triangle}{=} i_{\alpha\beta} - i^*_{\alpha\beta}$, $\hat{r}$ and $\hat{L}$ are estimates for r and L, respectively, which are computed with the following gradient update laws

$$\dot{\hat{r}} = \gamma_1(r_{max} - \hat{r})(r_{min} - \hat{r})\tilde{i}^T_{\alpha\beta} i^*_{\alpha\beta} \tag{10}$$

$$\dot{\hat{L}} = \gamma_2(L_{max} - \hat{L})(L_{min} - \hat{L})\tilde{i}^T_{\alpha\beta}\frac{d}{dt}i^*_{\alpha\beta} \tag{11}$$

where γ_1 and γ_2 are positive constants, $r_{min} \le r \le r_{max}$ are known upper and lower bounds for r, and $L_{min} \le r \le L_{max}$ are known upper and lower bounds for L.

For the generation of such an adaptive law the following energy storage function is considered

$$V_1 = \frac{L}{2}\tilde{i}^2_{\alpha\beta} + \frac{1}{\gamma_1}\log\left|\frac{(\hat{r} - r_{max})^{\tau_r}}{(\hat{r} - r_{min})^{\tau_r + 1}}\right| + $$

$$+\frac{1}{\gamma_2}\log\left|\frac{(\hat{L} - L_{max})^{\tau_L}}{(\hat{L} - L_{min})^{\tau_L + 1}}\right|$$

where $\tau \triangleq \frac{r-r_{max}}{r-r_{min}} < 0$ and $\tau \triangleq \frac{L-L_{max}}{L-L_{min}} < 0$.

Note that the controller (9) deals in a natural way with the problem of injection of a second harmonic, which appears on the dc voltage during unbalanced conditions. This comes from a partial inversion of the system, since $u_{\alpha\beta}$ contains v_C^{-1} which cancels the corresponding term in the product $\frac{1}{2}v_C u_{\alpha\beta}$.

Design of the reference current time derivative

The main issue in the controller above is the requirement to have available the time derivative of the reference current vector. Since we are dealing with unbalanced three phase system, even for the harmonic components, the computation of the time derivative $\frac{d}{dt}i^*_{\alpha\beta}$ demands a more elaborate procedure.

Consider the description of the k^{th} harmonic in a-b-c coordinates of a signal x:

$$x_k = \begin{bmatrix} \alpha_1 \cos(kwt + \phi_1) \\ \alpha_2 \cos(kwt + \phi_2) \\ \alpha_3 \cos(kwt + \phi_3) \end{bmatrix}$$

where α_i, ϕ_i, $(i \in \{1,2,3\})$ are constants, and x is either i_L or v_S .

The time derivative of such vector is

$$\dot{x}_k = -kw \begin{bmatrix} \alpha_1 \sin(kwt + \phi_1) \\ \alpha_2 \sin(kwt + \phi_2) \\ \alpha_3 \sin(kwt + \phi_3) \end{bmatrix}$$

The difference between the elements of the vector x and their time derivatives, besides amplitude and sign, is just a phase delay of $\pi/2$ rad, i.e., $\cos(wt - \pi/2 - \phi_i) = \sin(wt - \phi_i)$. Thus, we propose to reconstruct $\sin(kwt - \phi_i)$ from $\cos(kwt - \phi_i)$ as follows

$$\sin(kwt - \phi_i) \cong -kwG(\frac{p}{k})\cos(kwt - \phi_i) \quad (12)$$

$$G(\frac{p}{k}) = \frac{\psi w}{(p/k)^2 + \psi(p/k) + w^2} \quad (13)$$

where $p \triangleq \frac{d}{dt}$ is the time derivative operator.

Note that ψ should be chosen closed to the critical damping $(2w)$ of the second order filter $G(\frac{p}{k})$ so as to guarantee fast non-oscillatory response. Note that $G(\frac{p}{k})$ acts as an observer (with relatively fast convergence governed by ψ) which introduces a phase shift of $\pi/2$ and unit amplitude when the signal has frequency kw.

Thus the time derivative $\dot{x}$ can be reconstructed as

$$\frac{d}{dt}x_k \cong -kw\mathcal{G}_3(\frac{p}{k})x_k$$

where $\mathcal{G}_3(\frac{p}{k}) = \text{diag}\{G(\frac{p}{k}),\, G(\frac{p}{k}),\, G(\frac{p}{k})\}$

This approximation can be expressed in terms of conventional coordinates $\alpha\beta$ using transformation (3) as follows

$$\begin{aligned} \frac{d}{dt}x_{\alpha\beta,k} &\cong -kw\mathcal{A}\mathcal{G}_3 x_k = -kw\mathcal{G}_2(\frac{p}{k})\mathcal{A}x_k \\ &= -kw\mathcal{G}_2(\frac{p}{k})x_{\alpha\beta,k} \end{aligned}$$

where $\mathcal{G}_2(\frac{p}{k}) = \text{diag}\{G(\frac{p}{k}),\, G(\frac{p}{k})\}$. The result above follows from the linearity of both operators $\mathcal{A}$ and $\mathcal{G}_3(\frac{p}{k})$.

Using the approximation above, the load currents $i_{L\alpha\beta}$ and the source voltages $v_{S\alpha\beta}$ (rich in possibly unbalanced harmonics) can be decomposed as

$$i_{L\alpha\beta} = \sum_{k=1}^{\infty} i_{L\alpha\beta,\,k} \;\;,\;\; v_{S\alpha\beta} = \sum_{k=1}^{\infty} v_{S\alpha\beta,\,k}$$

where $(\cdot)_k$ states for the k^{th} harmonic component, obtained with a band pass filter tuned at that frequency.

Then the reference currents (7) can be rewritten as follows

$$i^*_{\alpha\beta} = i^*_{S\alpha\beta} - i_{L\alpha\beta} = \sum_{k=1}^{\infty} (g_T v_{S\alpha\beta,\,k} - i_{L\alpha\beta,\,k}) \quad (14)$$

and the time derivative can be reconstructed as

$$\frac{d}{dt}i^*_{\alpha\beta} \cong -w\sum_{k=1}^{\infty} k\mathcal{G}_2(\frac{p}{k})\,(g_T v_{S\alpha\beta,\,k} - i_{L\alpha\beta,\,k}) \quad (15)$$

4 Outer Control Loop

In the previous case we needed v_C bounded away from zero in order to ensure boundedness of the control signal $u_{\alpha\beta}$. In addition, due to physical constraints, the amplitude of $u_{\alpha\beta}$ is restricted to reside in a constrained region that we describe shortly. This imposes another (larger) value on the lower bound of v_C. In other words, the current tracking can not be achieved unless enough energy is stored in the capacitor. Hence, a second objective consists in design an "outer control loop" ensuring that the average of voltage (taken over a period of the fundamental) maintains a sufficiently high value. Since we are mainly interested in the DC component of v_C, we will consider the following subsystem for the design

$$C\frac{d}{dt}v_{C0} = \frac{3}{4}\langle u^T_{\alpha\beta} i_{\alpha\beta}\rangle_0 - \frac{v_{C0}}{R} \quad (16)$$

where $\langle x(t)\rangle_0 \triangleq \frac{1}{T}\int_{(t-T)}^{t} x(\tau)d\tau$ stands for the possibly time–varying dc component, and $v_{C0}(t) = \langle v_C(t)\rangle_0$.

The design of such an "outer control loop", is reduced to the computation of the gain g used in the definition of $i^*_{\alpha\beta}$ in (7), as we clarify shortly. This "outer control loop" should add damping to reinforce the asymptotic stability of the closed loop system, and should incorporate adaptation via a gradient law to compensate for the uncertainties in R.

For the sake of simplicity, we will assume that the dynamics of the "current control loop" is much faster than the dynamics of "outer control loop". This assumption is typically very well satisfied in practice. Thus, after a relatively short time, we can assume that the currents have reached perfect tracking on their references. Otherwise, we can use the vanishing perturbation theory to relax this decoupling assumption [7]. Thus, in the development that follows we assume that $\tilde{i}_{\alpha\beta} \approx 0$, $\hat{r} \approx r$ and $\hat{L} \approx L$. Direct substitution of the control (9) in (16), under these assumptions, yields the following expression for the output capacitor voltage dynamics

$$v_{C0}C\dot{v}_{C0} = \frac{3}{2}\langle L(i^*_{\alpha\beta})^T \frac{d}{dt}i^*_{\alpha\beta} - ri^{*2}_{\alpha\beta} + (i^*_{\alpha\beta})^T v_{S\alpha\beta}\rangle_0 - \frac{v^2_{C0}}{R}$$

where the term $\langle (i^*_{\alpha\beta})^T v_{S\alpha\beta}\rangle_0$ plays the role of the actual control signal.

We will consider the term $\langle ri^{*2}_{\alpha\beta}\rangle_0$ as an unknown constant to which we will refer as a, as the computation of the control signal becomes unnecessarily complex otherwise. The term $\langle L(i^*_{\alpha\beta})^T \frac{d}{dt}i^*_{\alpha\beta}\rangle_0$ is the dot (inner) product of two perpendicular vectors, so it can be considered very close to zero. Nevertheless, because of the approximation used for the time derivative, a small error is introduced that will be absorbed in a.

From the definition of $i^*_{\alpha\beta}$ given in (7), the term $\langle (i^*_{\alpha\beta})^T v_{S\alpha\beta}\rangle_0$ can be expressed as

$$\langle (i^*_{\alpha\beta})^T v_{S\alpha\beta}\rangle_0 = g'_T\langle (v_{S\alpha\beta})^T v_{S\alpha\beta}\rangle_0 + g\langle (v_{S\alpha\beta})^T v_{S\alpha\beta}\rangle_0 \tag{17}$$

where $\langle (v_{S\alpha\beta})^T v_{S\alpha\beta}\rangle_0 = \sum_{k=1}^{\infty}(v_{S\alpha\beta,\,k})^T v_{S\alpha\beta,\,k}$ is a constant to which we will refer as E. Notice that, $Eg'_T = \sum_{k=1}^{\infty}(i_{L\alpha\beta,\,k})^T v_{S\alpha\beta,\,k}$, which is a constant.

Let us introduce the change of variables $z = v^2_{C0}/2$ and $\theta = 1/R$. Notice that, if $v^*_{C0} = V_d$, then $z^* = V_d^2/2$. The model is then reduced to

$$C\dot{z} = \frac{3}{2}(-a - Eg'_T + Eg) - 2\theta z \tag{18}$$

where it is evident that the actual control depends only on the scalar function g. Thus, the control design procedure is reduced to the computation of the amplitude g which is used in $i^*_{\alpha\beta}$ (7).

For the system (18) we propose the controller

$$g = \hat{\alpha} - k_p\tilde{z} \tag{19}$$

where $\tilde{z} \triangleq z - z^*$, $k_p > 0$ is a design parameter, and we have defined $\hat{\alpha} \triangleq \frac{\hat{a}}{E} + \frac{4\hat{\theta}z^*}{3E} + \hat{g}'_T$ to combine the sum of estimates into a single parameter which is updated according to

$$\dot{\hat{\alpha}} = -k_i\tilde{z} \tag{20}$$

Note that we are not using the fact that g'_T is a known constant, hence no cancellation[1] occurs; instead, we are compensating the unknown term through an integral action.

The error model for this subsystem is

$$C\dot{\tilde{z}} = -(2\theta + \frac{3EK_p}{2})\tilde{z} + \frac{3E}{2}\tilde{\alpha}$$

where $\tilde{\alpha} \triangleq \hat{\alpha} - \alpha$, with $\alpha \triangleq \frac{a}{E} + \frac{4\theta z^*}{3E} + g'_T$.

Using the quadratic storage function $V_2 = \frac{C}{2}\tilde{z}^2 + \frac{3E}{4k_i}\tilde{\alpha}^2$, it is easy to see that the adaptive law (20) makes $\dot{V}_2$ negative semi-definite.

[1] For cancellation use g'_T instead of $\hat{g}'_T$ and $\hat{\alpha} \triangleq \frac{\hat{a}}{E} + \frac{4\hat{\theta}z^*}{3}$.

The "outer control loop" composed by (19) and (20) represents a typical PI controller, where k_p and k_i are the proportional and integral gains, respectively. Recall that in the "inner/current control loop" we assumed that the dynamics of g and v_C is much slower, so that they can be treated as constants. Thus the PI controller should be tuned in a way that does not invalidate this assumption.

5 Simulation results

For the purpose of simulation, we use a system model with the following parameters: $L = 2$mH, $C = 2200\mu$F, $r = 0.025\Omega$, $R = 1K\Omega$, at a constant frequency $f = 60$ Hz. The control design parameters are: $V_d = 500$ Volts, $r_{max} = .2\Omega$, $r_{min} = .0025\Omega$, $\gamma_1 = 17.25$, $k_i = 0.0004$, $k_p = 0.000375$ and $k_1 = 7.5$; with the initial conditions $v_{C0}(0) = 200$ Volts, and $i(0) = [0,0,0]^T$ Amps, $\hat{r}(0) = 0.05\Omega$. In the simulations shown here, we do not estimate L.

We will consider two conditions of the current load – at the beginning the system takes the form shown in Fig. 2, then after time $t = 1$ sec the second loading shown in Fig. 3 is applied. Both current loads are composed by a fundamental, 3^{rd} and 5^{th} harmonics, which are completely unbalanced. The voltage source is composed of the fundamental, 3^{rd} and 5^{th} harmonics that are also unbalanced – a scaled plot $0.5 * v_{S\alpha\beta}(t)$ is provided in both figures for comparison.

The compensation for harmonics and reactive power is *not* activated until time $t = 0.5$ sec. Thus, during the first 0.5 seconds the system is working as a simple synchronous rectifier, which allows the system to reach perfect regulation of v_{C0} to $V_d = 500$ Volts, as shown in Fig. 5 (without harmonics or reactive power compensation) . At time $t = 0.5$ sec, the compensation for the first load current condition commences. In Fig. 4, the compensated current $i_{S\alpha\beta}(t)$ is presented, together with the the scaled source voltage $0.25 * v_{S\alpha\beta}(t)$. Notice in Fig. 5 that, after a relatively short transient, the voltage v_C maintains its average on $V_d = 500$ Volts.

At time $t = 1$ sec., the load current takes the shape shown in Fig. 3. In Fig. 6 we present the phase space plot of the vector $i_{S\alpha\beta}$ –the inner trajectory corresponds to the first current condition before the transient, and the outer trajectory is the operation after the second condition. Note that both trajectories, after relatively short transient, obtain a form proportional to $v_{S\alpha\beta}(t)$ which is also presented, scaled by 0.35, in the same figure for comparison. In Fig. 7 the response of voltage $v_C(t)$ is shown following the change on the load current. Again we can observe that the controlled voltage reaches its desired value after a short transient.

The top panel of Fig. 8 displays the plots of the load $\langle v^T_{S\alpha\beta}i_{L\alpha\beta}\rangle_0$ and source $\langle v^T_{S\alpha\beta}i_{S\alpha\beta}\rangle_0$ active (real) power. Here we can observe that the active load and source power are very close to each other before and after compensation. The slight difference comes from the active power necessary to maintain the capacitor voltage in its desired average value in the presence of switching and other losses (modeled by the resistor R). In the lower diagram of the same figure

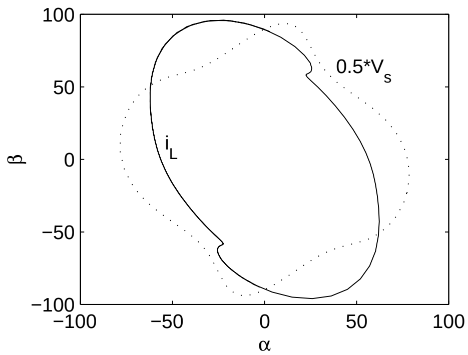

Figure 2: First current load condition $i_{L\alpha\beta}$ and scaled $0.5 * v_{S\alpha\beta}(t)$

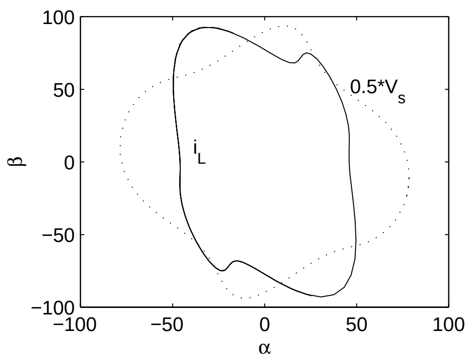

Figure 3: Second current load condition $i_{L\alpha\beta}$ and scaled $0.5 * v_{S\alpha\beta}(t)$

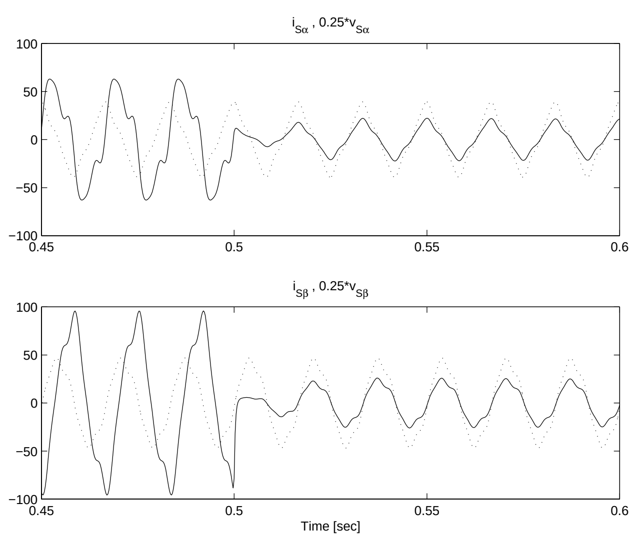

Figure 4: Transient response of $i_{S\alpha\beta}$ and scaled $0.25 * v_{S\alpha\beta}(t)$ when the compensation is enabled at $t = 0.5$ sec

we show the behavior of $g_T(t)$. Notice that in all variables there appears a small ripple coming from a second harmonic in v_C, we observe from our simulations that using a low pass filter to derive v_{C0} may cause large oscillations.

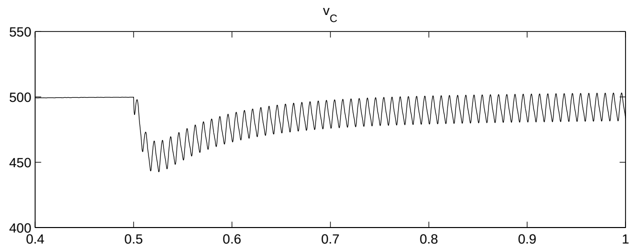

Figure 5: Transient response of $v_C(t)$ when the compensation is enabled at $t = 0.5$ sec

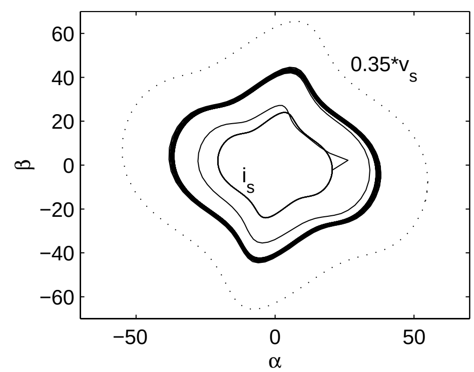

Figure 6: Transient response of the compensated $i_{S\alpha\beta}$ during the change of current load, and scaled $0.35 * v_{S\alpha\beta}(t)$

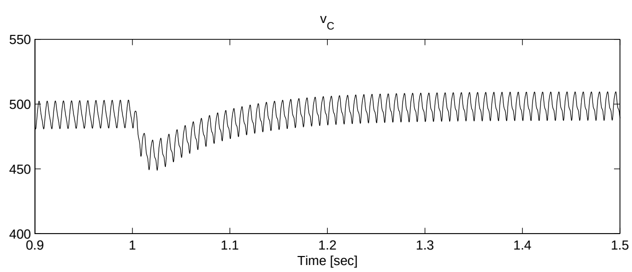

Figure 7: Transient response of $v_C(t)$ when the load current condition changes at $t = 1$ sec

Figs. 9 and 10 shown the phase space plots of the control signal in steady state for the two current loads. We have also displayed the hexagon where the control is constrained to reside[2]; note that in both cases the control signal generated by our algorithm is fully realizable, since it is completely contained inside the hexagon. We remark that the example presented here uses strongly deformed $v_{S\alpha\beta}$, and $i_{L\alpha\beta}$,

[2]These constraints arise due to the physical restrictions in voltage sourced inverters.

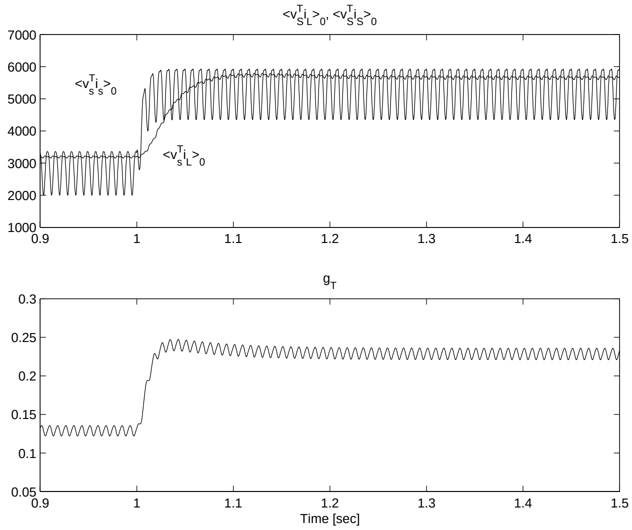

Figure 8: **(Top)** Transient response of the power $\langle v_{S\alpha\beta}^T i_{L\alpha\beta} \rangle_0$ and $\langle v_{S\alpha\beta}^T i_{S\alpha\beta} \rangle_0$ and **(Bottom)** Transient response of $g_T(t)$ when the load current condition changes at $t = 1$ sec

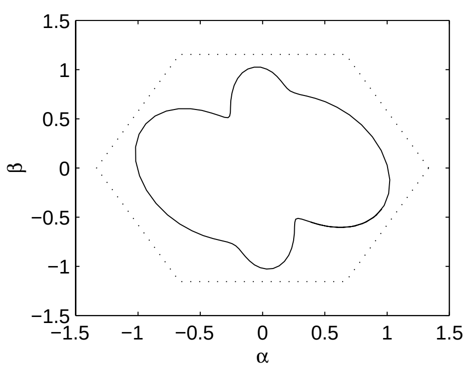

Figure 9: Steady state behavior of the control signal $u_{\alpha\beta}$ during the first current load condition

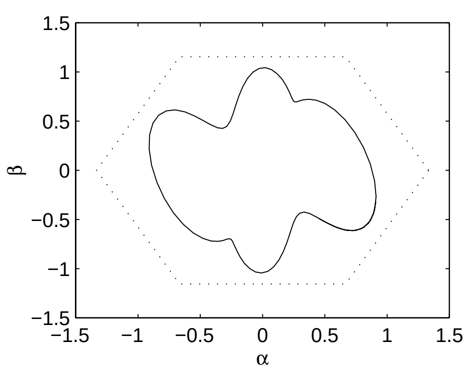

Figure 10: Steady state behavior of the control signal $u_{\alpha\beta}$ during the second current load condition

which in a control signal that is very different from a circle that corresponds to a balanced single harmonic case. In the case that even more deformed signals were considered, a larger V_d would be necessary to keep the control vector inside the hexagon, at least in steady state.

Conclusions

This paper presents a dissipativity–based controller for D-Statcom that simultaneously achieves compensation of reactive power, load harmonics and unbalances. A two–loop control design is described, together with simulations of a realistic model of a D-Statcom.

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