Proceedings of the
44th IEEE Conference on Decision and Control, and
the European Control Conference 2005
Seville, Spain, December 12-15, 2005

A Switched–Resistors Implementation of Compensators for Wave Reflections in Transmission Lines

Alessandro de Rinaldis, Alessandro Astolfi, Romeo Ortega and Mark W. Spong

Abstract— **In this paper we propose an approximated current–injection implementation of the wave reflection compensators derived in our previous works that can be easily realized with current technology. First, the (discrete) transfer function of the compensator is approximated by its truncated series. Then, the resulting finite impulse response is realized with *switched–shunt resistors*. Since the resistors are switched on only during brief time periods, and are actually disconnected after the transient wave reflections phenomenon, the extracted power is small making the implementation practically feasible.**

Notation We define the differentiation and advance–delay operators, acting on signals $x : \mathbb{R} \to \mathbb{R}$, as $(p^k x)(t) \stackrel{\triangle}{=} \frac{d^k}{dt^k}x(t)$ and $(q^{\pm k}x)(t) = x(t \pm \frac{kd}{n})$, respectively, where $d \in \mathbb{R}_+$ and $k, n \in \mathbb{Z}_+$. Their Laplace transform counterparts, which are used to define transfer functions, are s and $z = e^{\frac{d}{n}s}$, respectively.

I. Introduction

The problem of compensation of wave effects that appear when the controlled plant, with uncertain dynamic impedance, is coupled to a *fast switching* actuator through long feeding cables is of great practical importance. At high frequency operation the connecting cables behave as a transmission line inducing wave reflections that deform the transmitted signals and degrades the quality of the control. This problem has been studied in [3], [5] where a novel compensator design framework based on the scattering representation of the transmission line is introduced. A family of compensators that requires *only knowledge of the line parameters* and—under some practically reasonable assumptions—guarantees transient performance improvement and asymptotic tracking for all unknown plants with passive impedance is proposed. Unfortunately, in contrast with the standard schemes that can be implemented with RLC circuits, the proposed compensators are *active* and require for their implementation regulated—current *and* voltage—sources. In [4] an active filter implementation is suggested, but its practical realization is still stymied by technological considerations, in particular, the need for fast switching high power devices, not yet available in the market.

The purpose of this paper is twofold, first, we prove that, if plant prior knowledge is available, it is possible to obviate the regulated voltage source, that is, the compensators of [3], [5] can be implemented with a single shunt regulated current source. Second, to make the scheme practically feasible, we propose to *approximate* the action of the controller with *Switched–Shunt Resistors* (SSR). More precisely, the (discrete) transfer function of the compensator is approximated by its truncated series and a certain number of resistors, connected in parallel and placed in shunt, are switched on and off to realize the resulting finite impulse response. Since the resistors are switched on only during brief time periods, and are actually disconnected after the transient wave reflections phenomenon, the current flowing through them is small reducing the extracted power.

The remaining of the paper is organized as follows. To set–up the notation, the model of the overall system, derived in detail in [3], [5], is briefly presented in Section II. In Section III we present the proposed current–injection compensator topology and give necessary and sufficient conditions for systems well–posedness. Section IV is devoted to the derivation of the transfer functions for the regulated current source that implement the controllers of [3], [5], as well as the Half–DC Link scheme proposed in [1]. As indicated above the main drawback is the need of the exact knowledge of the load parameters. In Section V an approximation using SSR is presented. Finally, in Section VI a set of simulations tests the effectiveness of the proposed scheme comparing different number of resistors and switching policies.

II. Systems Model

To model the plant connected to the actuator through long cables we consider the configuration shown in Fig. 1, where we model the *connecting cables* as a two–port system whose dynamics are described via the Telegrapher's equations

$$C\frac{\partial v(t,x)}{\partial t} = -\frac{\partial i(t,x)}{\partial x}\ , \quad L\frac{\partial i(t,x)}{\partial t} = -\frac{\partial v(t,x)}{\partial x}, \qquad (1)$$

where $v(t,x)$, $i(t,x)$ represent the line voltage and current, respectively, $x \in [0, \ell]$ is the spatial coordinate, with $\ell > 0$ the cable length and $C, L > 0$, which are assumed constant, are the capacitance and inductance of the line, respectively.

Université Paris 11, LSS-Supélec, 3 rue Joliot-Curie 91192 Gif-sur-Yvette cedex, France derinaldis(ortega)@lss.supelec.fr

Electrical Engineering Departement, Imperial College, London SW 2BT, UK a.astolfi@imperial.ac.uk

Coordinated Science Laboratory University of Illinois, Urbana, IL 61801, USA m-spong@uiuc.edu

It is well-known that the solution of the above partial differential equations yields the following relation between the port variables of the transmission line (1)

$$\begin{bmatrix} v(t,\ell) \\ i(t,\ell) \end{bmatrix} = W(q) \begin{bmatrix} v(t,0) \\ i(t,0) \end{bmatrix} \tag{2}$$

where

$$\begin{aligned} W(z) &\triangleq T^{-1} \begin{bmatrix} z^{-n} & 0 \\ 0 & z^{n} \end{bmatrix} T \in \mathbb{R}^{2\times 2}(z) \ , \\ T &\triangleq \begin{bmatrix} 1 & Z_0 \\ 1 & -Z_0 \end{bmatrix} \end{aligned}$$

and $d \triangleq \ell\sqrt{LC}$ is the *propagation delay*.
The *actuator* consists of a voltage source, $v_s(t)$, connected in series with a resistor R_a, called the surge impedance.

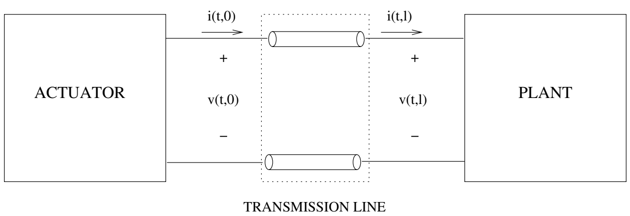

Fig. 1. *Uncompensated systems configuration.*

The transmission line is terminated by the *plant*, which is a one–port, with port variables $(v(t,\ell), i(t,\ell))$. If we assume the plant is LTI the dynamics of the overall system is described by (2) together with

$$\begin{aligned} v(t,0) &= -R_a i(t,0) + v_s(t) \\ v(t,\ell) &= Z_p(p) i(t,\ell), \end{aligned} \tag{3}$$

where $Z_p(s) \in \mathbb{R}(s)$ is the plant impedance—that we assume is strictly stable.
We need the following *generic* assumption:[1]

$$R_p + Z_0 \neq 0. \tag{4}$$

where $R_p \in \mathbb{R}$ is the high–frequency gain of the plant.

Under (4), it is possible to show that the mapping from the source voltage to the plant voltage is given by the linear *delay–differential* operator

$$\begin{aligned} v(t,\ell) = K_a K_p(p) v(t-2d,\ell) + \\ +\frac{1}{2}[1+K_p(p)](1-K_a) v_s(t-d) + \epsilon_t, \end{aligned} \tag{5}$$

where ϵ_t is an exponentially decaying term, that will be omitted in the sequel, and

$$K_a \triangleq \frac{R_a - Z_0}{R_a + Z_0}, \quad K_p(s) \triangleq \frac{1-\alpha(s)}{1+\alpha(s)}, \tag{6}$$

[1]The assumption is needed to ensure these transfer functions are well–defined. If $Z_p(s)$ is an LTI RLC filter then, because of Bruni's Theorem, it is a positive real transfer function with $R_p > 0$, and the assumption may be obviated.

are the so-called actuator and plant *reflection coefficients*, respectively, and for convenience we have defined

$$\alpha(s) \triangleq \frac{Z_0}{Z_p(s)}.$$

For further developments it will be assumed that $K_p(s)$ is also strictly stable. We make at this point the following crucial observation

- the delayed signal $K_a K_p(p) v(t-2d,\ell)$ is added to the filtered (delayed) pulse $v_s(t-d)$ to generate $v(t,\ell)$. This term captures the physical phenomenon of *wave reflection* that deforms the transmitted signals and degrades the quality of the control.

The interested reader is referred to [3], [5] for further details on the wave reflection problem as well as a discussion of the intrinsic limitations of current practice—namely, the introduction of impedance matching filters.

III. Current–Injection Compensation Topology

In [5] it is shown that placing, between the actuator and the transmission line, a regulated current source in shunt and a regulated voltage sources in parallel it is possible to guarantee transient performance improvement and asymptotic tracking for all *unknown plants* with passive impedance. Unfortunately, due to technological constraints related to the power flowing into the system, series devices should be avoided.

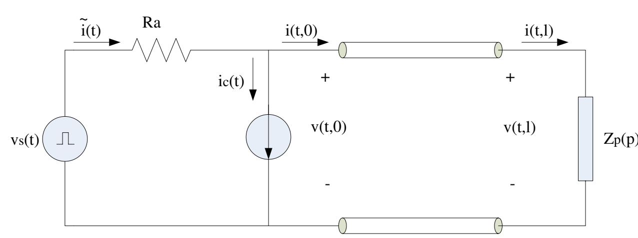

Fig. 2. *Proposed current–injection compensator configuration.*

To comply with this constraint, we consider in this paper the current–injection configuration of Fig. 2, from which we get

$$v_s(t) = v(t,0) + R_a \widetilde{i}(t) \tag{7}$$

$$i(t,0) = \widetilde{i}(t) - i_c(t). \tag{8}$$

Now, the plant connected to the transmission line satisfies the following relation

$$v(t,0) = Z_T(p,q) i(t,0) \tag{9}$$

with

$$Z_T(s,z) \triangleq Z_0 \frac{1+K_p(s)z^{-2n}}{1-K_p(s)z^{-2n}} \tag{10}$$

which is obtained by algebraic manipulations of equations (2) and (3).

The compensator, that we assume is an 2–dimensional LTI system, is described by its generalized impedance

$$v(t,0) = Z_c(p,q)i_c(t) \tag{11}$$

where the operator $Z_c(s,z) \in \mathbb{R}(s,z)$ is to be defined.

The following questions arise immediately:

- Q1 As $Z_c(s,z)$ ranges in $\mathbb{R}(s,z)$ what are the achievable behaviors for the map $v_s(t) \to v(t,\ell)$?
- Q2 What are the conditions on $Z_c(s,z)$ such that the system interconnection is well–posed?[2]

Question Q2 is highly technical and its rigorous answer requires additional assumptions, hence it will be deferred to Section IV.

To answer Q1 we proceed as follows. Replacing (11) into (8) and considering (10) we can re-write (7) as

$$v(t,0) = Q^{Z_c}(p,q)v_s(t)$$

with

$$Q^{Z_c}(s,z) = \frac{Z_T(s,z)}{Z_T(s,z) + R_a[1 + \frac{Z_T(s,z)}{Z_c(s,z)}]} \tag{12}$$

a transfer function parameterized by the compensator $Z_c(s,z)$.

Now, from (2) and the second equation of (3), we obtain

$$v(t,\ell) = \frac{2q^{-n}}{1+\alpha(p) + [1-\alpha(p)]\,q^{-2n}} v(t,0)$$

which, combined with (12), yields the transfer function of interest, namely

$$v(t,\ell) = \mathcal{M}^{Z_c}(p,q)v_s(t-d) \tag{13}$$

where

$$\mathcal{M}^{Z_c}(s,z) = \frac{2}{1+\alpha(s) + [1-\alpha(s)]\,z^{-2n}} Q^{Z_c}(s,z).$$

To prove the interest of the derivations above we will derive in the sequel *current–injection implementations* of the Dynamic Inversion and Current–Decoupling compensators proposed [5], as well as the very interesting Half–DC Link scheme of [1].

A. Dynamic Inversion

As shown in [5], and readily follows from (5), the 2–dimensional filter

$$2\frac{1-K_aK_p(s)z^{-2n}}{[1+K_p(s)](1-K_a)}$$

can be used to cancel the transmission line effects and yield the relation $v(t,\ell) = v_s(t-d)$. Referring to (13) and fixing $\mathcal{M}^{Z_c}_{DI} = 1$ we can then compute, after some lengthy but straightforward calculations, the compensator impedance that achieves dynamic inversion as

$$Z_c^{DI}(s,z) = R_a \frac{1+\alpha(s) + [1-\alpha(s)]z^{-2n}}{[1-\alpha(s)][1-(1-\frac{R_a}{Z_0})z^{-2n}] - \frac{R_a}{Z_0}[1+\alpha(s)]} \tag{14}$$

The relation above defines the current $i_c(t)$ to be injected to the system in order to completely avoid wave reflections. Besides the well–known fragility of exact cancellations, another important drawback of the scheme is the need of the exact knowledge of all (actuator, load, and transmission line) parameters. Adaptive implementations of the scheme are also complicated by the fact that the plant transfer function (9) cannot be discretized without additional assumptions, see [2] for an alternative configuration.

B. A Current–Decoupling Scheme

In [5] it has also been shown that, using current and voltage regulated sources but without knowledge of the plant parameters, it is possible to assign the transfer function

$$\mathcal{M}^{Z_c}_{CD}(s,z) = \frac{1}{1+\alpha(s) - \alpha(s)(1-\frac{R_a}{Z_0})z^{-2n}},$$

for which the ringing phenomenon is considerably reduced. (The name ”Current–Decoupling” stems from the fact that the transfer matrix from the voltage and current port variables of the actuator to the transmission line terminal port variables, i.e., $\left[\begin{array}{c} v(t,\ell), i(t,\ell) \end{array}\right]$, is lower triangular.) Another advantage of achieving current–decoupling is that, as shown in [5], an adaptive implementation that obviated the need of knowing the line parameters is possible.

Working back from $\mathcal{M}^{Z_c}_{CD}(s,z)$ we can compute

$$Q^{Z_c}_{CD}(s,z) \triangleq \frac{1}{2}\frac{1+\alpha(s) + [1-\alpha(s)]\,z^{-2n}}{1+\alpha(s) - \alpha(s)(1-\frac{R_a}{Z_0})z^{-2n}}$$

and the required current–injection impedance

$$Z_c^{CD}(s,z) = R_a \frac{1+\alpha(s) + [1-\alpha(s)]z^{-2n}}{(1-\frac{R_a}{Z_0})(1+\alpha(s))(1-z^{-2n})}. \tag{15}$$

We invite the reader to compare this transfer function with the one of the ”ideal” compensator (14), and remark the plus instead of the minus sign in one of the denominator factors.

C. Half DC–Link Scheme of [1]

In [1] an interesting solution for the over-voltage suppression in AC drives fed through long cables has been proposed. The basic idea is to split the transmitted pulse into two pulses (see Fig. 3)

$$v(t,0) = \frac{1}{2}v_s(t) + \frac{1}{2}v_s(t-2d), \tag{16}$$

to counter, invoking some kind of superposition principle, the effect of the reflected wave.

[2]The importance of this issue, that is sometimes overlooked in the control literature, is thoroughly discussed in [3], [5].

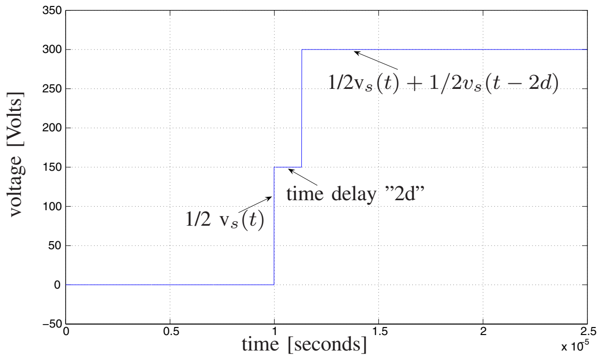

Fig. 3. *Half DC-link compensation scheme. The input voltage* $v(t, 0)$.

This scheme can be also implemented by our current–based approach. Indeed, in this case we have

$$Q_H^{Z_c}(s,z) \triangleq \frac{1}{2}(1+z^{-2n}), \tag{17}$$

and, from (12), an expression for the controlled impedance can be derived. It is interesting to observe that, if $\frac{R_a}{Z_0} \approx 0$, this impedance has the very simple form

$$Z_c^H(z) = R_a \frac{1+z^{-2n}}{1-z^{-2n}}, \tag{18}$$

that can be implemented without plant knowledge. For the sake of completeness we also give the resulting transfer

$$\mathcal{M}_{Z_c}^H(s,z) = \frac{1}{1+\alpha(s)-\alpha(s)\frac{2z^{-2n}}{1+z^{-2n}}}$$

An interpretation of the action of this controller is obtained noting that the ideal dynamic inversion impedance (14) reduces to (18) when $\alpha(s) = 0$, i.e., when the load dynamic $Z_p(s)$ "dominates"—at all frequencies—the line characteristic impedance Z_0.

IV. Well–Posedness Analysis

Since the system is described by delay–differential equations, for which well-posedness seems to be a complex issue, we introduce the following discretization assumption to answer Q2.

Assumption A.1 The sample time $T_s = d/n$ is sufficiently small so that all signals can be suitably described by their piecewise approximation. More precisely, for all signals $x : \mathbb{R}_+ \to \mathbb{R}$

$$x(t) = x(kT_s), \quad \forall\, t \in [kT_s, (k+1)T_s), \quad k \in \mathbb{Z}_+.$$

As discussed in [5], under the "discretization" Assumption A.1 the overall dynamics, at the sampling instants kT_s, is described by a *purely discrete–time system*, for which the well–posedness analysis follows standard lines. Indeed, under Assumption A.1, the plant voltage $v(t,\ell)$ becomes[3]

$$v(t,\ell) = Z_p^*(q)i(t,\ell), \quad \forall\, t \in [kT_s,(k+1)T_s), \quad k \in \mathbb{Z}_+$$

with $Z_p^*(z) \in \mathbb{R}(z)$ the pulse transfer function representation (with sampling time T_s) of the plant impedance. Similarly, we obtain the discrete–time equivalents of (9) and (11) as

$$v(t,0) = Z_T^*(q)i(t,0), \; v(t,0) = Z_c^*(q)i_c(t),$$

respectively.

The proposition below provides an answer to Q2.

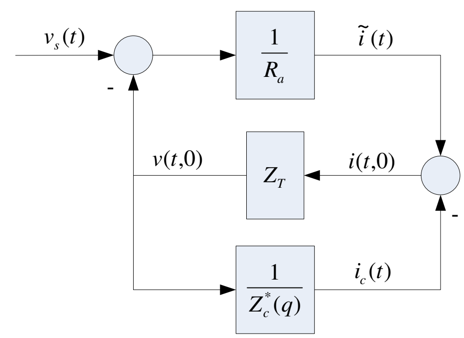

Fig. 4. *Block diagram of the overall system.*

Proposition 1: Consider the overall system described by the equations (7), (8), (9) and (11) with Assumption A.1. The system is well–posed if and only if

$$\begin{aligned} \tilde{Z}_c(\infty) &\neq -Z_0 \\ \tilde{Z}_c(\infty) &\neq \frac{-Z_0 R_a}{R_a + Z_0} \end{aligned}$$

Proof: The block diagram of Fig. 4 depicts the system behavior. From direct inspection of the block diagram we see that there are two loops whose well–posedness needs to be checked. The inner loop is well–posed if and only if

$$1 + \frac{Z_T(\infty)}{Z_c(\infty)} \neq 0,$$

and $Z_T(\infty) = Z_0$, which gives the first condition above. Well–posedness of the outer loop imposes

$$1 + \frac{1}{R_a}\frac{Z_T(\infty)}{1+\frac{Z_T(\infty)}{Z_c(\infty)}} \neq 0,$$

that, after some simple algebraic manipulations, gives the second condition. ■

Applying this result to the compensators proposed in Section III we get that Z_c^{DI} yields a well posed interconnection provided $Z_p^*(\infty) \neq 0$, i.e., the plant transfer function is proper but not strictly proper, and $Z_0 \neq 0$. On the other hand, well–posedness for Z_c^{CD} and Z_c^H imposes $Z_0 \neq 0$ and $Z_0 \neq R_a$.

[3]To simplify the notation we preserve the continuous–time notation $(\cdot)(t)$ for all signals, in the understanding that they are constant along the sampling periods.

V. An Approximate Implementation

In this section we propose a procedure to approximate the controllers transfer function so that the scheme can be practically implemented. Namely, we propose to the action of the controller by a purely discrete transfer function that is then described by its truncated series. To realize the resulting finite impulse response a certain number of resistors, connected in parallel and placed in shunt, are switched on and off—yielding the SSR implementation shown in Fig. 5.

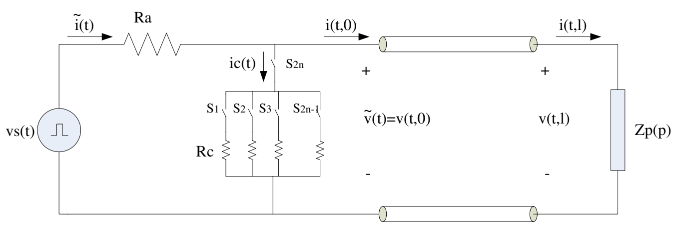

Fig. 5. *Electrical scheme of the system with SSR compensator placed on the actuator side.*

An additional approximation step, that allows us to obtain much simpler expressions for all the proposed compensators, is needed for our derivations. Taking into account that the actuator resistance R_a is generally of the order of a few Ohms while the characteristic impedance of the line, which does not depend on the length of the cable, is of the order of hundreds of Ohms it is reasonable to make the following assumption.

Assumption A.2 $\frac{R_a}{Z_0} \approx 0$.

Our motivation for the SSR scheme stems from the Half DC–Link scheme, whose impedance reduces, under Assumption A.2, to (18). Using the fact that

$$\frac{1}{1-z^{-2n}} = \sum_{i=0}^{\infty} z^{-i2n},$$

we obtain the series expansion

$$Z_c^H(z) = R_a \left[\sum_{i=0}^{\infty} z^{-i2n} + z^{-2n} \sum_{i=0}^{\infty} z^{-i2n} \right].$$

We propose now to truncate this series retaining only n terms. For $n = 2$, we see that the controller can be implemented with a single resistor $R_c = R_a$, connected for $0 \le t < 2d$ and disconnected for $2d \le t < T^P$, with T^P being the period of the train of pulses. In this way we have

$$\begin{aligned} Z_c(1,2) &= R_a, \\ Z_c(3,4,...) &= \infty. \end{aligned}$$

A similar treatment is done for the remaining compensators. Under Assumption A.2, the following approximated expression is obtained for (14)

$$Z_c^{DI}(s,z) \cong R_a \frac{1+\alpha(s)+[1-\alpha(s)]z^{-2n}}{[1-\alpha(s)](1-z^{-2n})} \tag{19}$$

that can be expanded as

$$Z_c^{DI}(s,z) = R_a \left[K_p^{-1}(s) \sum_{i=0}^{\infty} z^{-i2n} + z^{-2n} \sum_{i=0}^{\infty} z^{-i2n} \right].$$

Similarly, the impedance of the current-decoupling scheme (15), under Assumption A.2, becomes

$$Z_c^{CD}(s,z) \cong R_a \frac{1+\alpha(s)+[1-\alpha(s)]z^{-2n}}{[1+\alpha(s)](1-z^{-2n})} \tag{20}$$

which, can be written as

$$Z_c^{CD}(s,z) = R_a \left[\sum_{i=0}^{\infty} z^{-i2n} + K_p(s) z^{-2n} \sum_{i=0}^{\infty} z^{-i2n} \right].$$

Because of the presence of the plant reflection coefficient $K_p(s)$, the expressions above involve continuous and discrete dynamics—even after truncation of the series. However, if we assume that the plant is purely resistive, then the transfer function is purely discrete and the coefficients of the series can be explicitly computed. This is the approach we adopt in the simulations below.

VI. Simulation results

We consider the same AC motor benchmark presented in [3] to which we refer for the numerical values of the electrical components. Assuming $n = 2$, the compensator has been implemented with three resistors, with values

$$R_{c_1} = 1.7R_a,\ R_{c_2} = R_a,\ R_{c_3} = 8R_a,$$

connected in parallel as depicted in Fig. 5, where each resistor is controlled by his related ideal switch s_1, s_2, s_3, and the all set of switched-shunt resistors is controlled by another ideal switch s_4 able to disconnect the active filter yielding the equivalent shunt impedance $R_c \to \infty$. The switching policy depends on the particular choice of compensation scheme adopted. Of course not all scheme's implementations require the use of 4 switches—for example in the sequel we present the switching policy for the implementation of (18) where, clearly, switches s_1 and s_3 don't play any role—, but our intention is to show that a compensator adopting just three SSRs lead to good approximation of the proposed schemes and, above all, a significant attenuation of the wave reflection phenomenon.

In order to implement the scheme proposed in (18) we defined the following switching policy:

- for $0 \le t < \frac{2d}{n}$ then s_1, s_3 are *open* and s_2, s_4 are *closed*
- for $\frac{2d}{n} \le t < T^P$ then s_2, s_4 are *open* and s_1, s_3 are *closed*

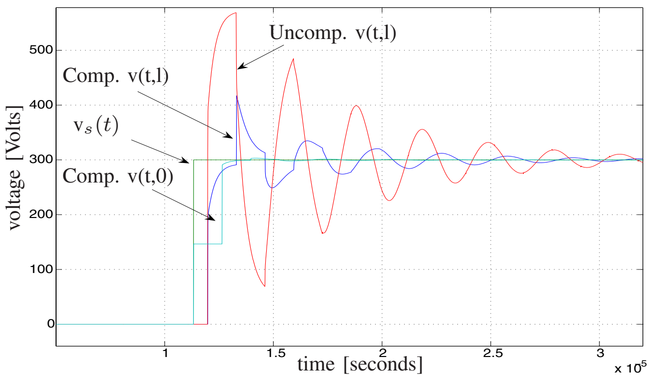

Fig. 6. Simulations results using one switched shunt resistor.

By observing the simulation results in Fig. 6, one should notice that the overshoot has been reduced around 60 %, unfortunately decreasing the rise time. Although the implementation of SSR is very simple, it inherits the drawback of the passive filters approach—whose main action is to slow down the rise time.

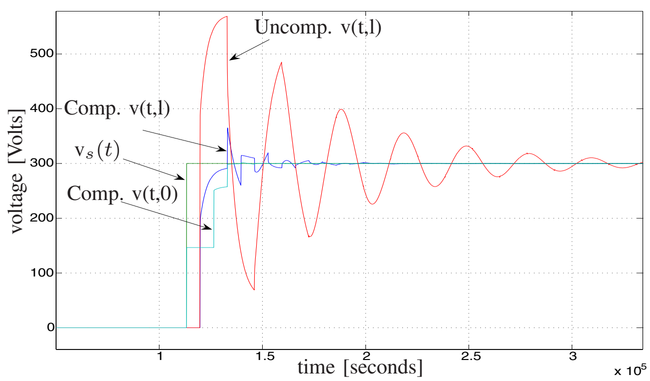

Fig. 7. Simulations results using two switched shunt resistors.

Applying the current-decoupling scheme, approximated by in (20), we try to implement it using the following switching policy:

- for $0 \leq t < \frac{2d}{n}$ then s_1, s_3 are *open* and s_2, s_4 are *closed*
- for $\frac{2d}{n} \leq t < \frac{3d}{n}$ then s_1, s_2 are *open* and s_3, s_4 are *closed*
- for $\frac{3d}{n} \leq t < T^P$ then s_4 is *open* and s_1, s_2, s_3 are *closed*

Remark that in this case s_1 does not play any role. By inspection of Fig. 7 compared to Fig. 6 one can observe that the use of the switched-resistor R_{c_3} improved remarkably the performance of the compensator, leading to an overshoot reduction of 80 % without affecting the rise time.

Furthermore, a rough approximation of the ideal-compensator scheme can be also implemented enabling the use of the switch s_1, as depicted in Fig. 8, according with the following switching policy:

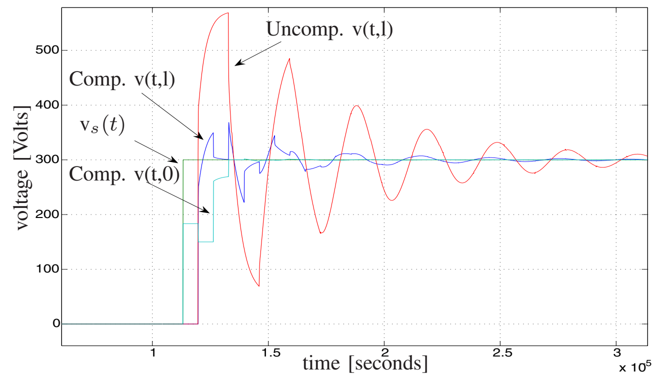

Fig. 8. Simulations results using three switched shunt resistors.

- for $0 \leq t < \frac{d}{n}$ then s_2, s_3 are *open* and s_1, s_4 are *closed*,
- for $\frac{d}{n} \leq t < \frac{2d}{n}$ then s_1, s_3 are *open* and s_2, s_4 are *closed*,
- for $\frac{2d}{n} \leq t < \frac{3d}{n}$ then s_1, s_2 are *open* and s_3, s_4 are *closed*,
- for $\frac{3d}{n} \leq t < T^P$ then s_4 is *open* and s_1, s_2, s_3 are *closed*,

So, even if the overshoot reduction has been slightly improved up to 85 % the principal advantage of using the former strategy concerns the speed of the load-side voltage rise time. Indeed, we have seriously reduced this secondary effect keeping good transient performances.

Acknowledgements

The work was (partially) supported through a European Community Marie Curie Fellowship and in the framework of the CTS, contract number: HPMT-CT-2001-00278.

References

[1] S. C. Lee and K. H. Nam, "An Overvoltage Suppression Scheme for AC Motor Drives Using a Half DC-Link Voltage Level at each PWM Transition", *IEEE Trans. Ind. Elec.*, Vol. 49, No. 3, pp. 549-557, June, 2002.

[2] R. Narashima, M. Spong and R. Ortega, "A Feedback Linearization Scheme To Cancel Wave Reflections In Transmission Lines", *SCL, University of Illinois*, Int. Rep., May 2005.

[3] R. Ortega, A. de Rinaldis, M. Spong, S. C. Lee and K. Nam, "On Compensation of Wave Reflections in Transmission Lines and Applications to the Overvoltage Problem in AC Motor Drives", *IEEE Trans. Aut. Cont.*, Vol. 49, pp. 1757-1762, Oct, 2004.

[4] R. Ortega, A. de Rinaldis, G. Escobar and M. Spong, "A Hybrid Active Filter Implementation of an Overvoltage Suppression Scheme", *IEEE Proceedings ISIE-2004*, pp. 1141-1146, May, 2004.

[5] A. de Rinaldis, R. Ortega and M. Spong "A Compensator for Attenuation of Wave Reflections in Long Cable ActuatorPlant Interconnactions with Guaranteed Transient Performance Improvement", *IFAC World Congress 2005*, Prague, Czech Republic, July 4–8, 2005.