ABSTRACT

High performance high bandwidth control of power electronic converters, inverters, and motor drives has become feasible over the past decade. These devices behave as constant power loads over large bandwidths when they are tightly regulated. However, constant power loads have a severe side effect known as negative impedance instability. In order to mitigate the problem of negative impedance instability a new nonlinear system stabilizing controller has been developed. The details of how this controller works along with its implementation is discussed and demonstrated in hardware.

INTRODUCTION

Power electronics based power and propulsion systems are topics of great interest to industry and the military. Applications include submarines, ships, hybrid electric and electric vehicles, aircraft, and spacecraft. The flexibility and potential for high bandwidth control in these systems, which is afforded by the use of power electronic converters/inverters, reduces the need for human operators and at the same time permits a greater degree of optimization, reducing fuel costs. As a result of fuel savings, these systems also offer lower emissions. However, with these benefits come one disadvantage in that these systems tend to be inherently dynamically unstable.

In such a system, most loads take the form of electric drive systems, dc/ac converters, or dc/dc converters. In the past, many different control schemes have been investigated including Sliding Mode Control [2], Fuzzy Logic [3 and 4], Nonlinear Proportional Integral (PI) [5], as well as many others linear and nonlinear [6 and 7]. Regardless of the strategy, these different control laws share the property that if they tightly regulate the converter/inverter output, the converter/inverter presents a constant power load to the system. Thus, the current entering the converter/inverter increases if the voltage at the input decreases, causing the converter/inverter to display a negative impedance characteristic as viewed from the rest of the system - clearly a destabilizing effect.

This has led to an impedance based stability criteria that predicts, on an operating point basis, when the overall system will be unstable. Using this criteria there are several techniques that can be used to eliminate the negative impedance instability. The first of which is designing the converter regulatory control with very low bandwidth so that the negative impedance characteristic is reduced. This can be effective but results in poor regulation of the output of the converters/inverters. Another approach is to incorporate a passive RC damping network to stabilize the system. This is also effective but the extra hardware and space required can be quite expensive, and system efficiency is reduced. The path chosen here is to modulate the commanded input power of the load converters at the appropriate frequencies with a nonlinear system stabilizing controller (NSSC). One possibility to accomplish this is to monitor the power passed down the AC transmission line [8] and from this measurement modulate the commanded converter/inverter input power. This requires additional sensors placed long distances from the converter/inverter. Another possibility is the use of a linear feed forward pole-zero cancellation control as discussed in [9] to eliminate the instability. This linear feed forward control requires a duty cycle controlled converter/inverter with the assumption that the input voltage times the on time of the upper transistor is a constant. The technique chosen here, [10 and 11], uses a new nonlinear feed forward control law that is not only simple to implement but also has minimal effect on the desired performance of the system and at the same time guarantees system stability. This nonlinear control law works well with current controlled converters/inverters, which allow better control in terms of safety and current limiting. The NSSC is augmented to the existing controls which are designed solely on the basis of output regulation. This new control scheme is verified on both the induction motor drive (IMD) of a propulsion system,
and the dc/dc buck boost converter (BBC) of a DC power system.

BASE TEST SYSTEMS

INDUCTION MOTOR BASED ELECTRIC PROPULSION SYSTEM - Figure 1 illustrates the type of electric propulsion system considered herein, [1]. The power source of the system is a diesel engine or turbine (emulated by a dynamometer), which serves as a prime mover for the 3-phase synchronous machine (SM). The 3-phase output of the machine is rectified using an uncontrolled rectifier. The rectifier output voltage is denoted $v_r$. An LC circuit serves as a filter, and the output of this filter is denoted $v_{dcx}$. A voltage regulator / exciter adjusts the field voltage of the SM in such a way that the source bus voltage $v_{dcx}$ is equal to the commanded bus voltage $v_{dcx}^*$. The source bus is connected via a tie line to the load bus, the voltage at which is denoted $v_{dc1}$. The load bus consists of a capacitive filter (which includes both electrolytic and polypropylene capacitance) as well as a 3-phase fully controlled inverter, which in turn supplies an induction motor. The induction motor drives the mechanical load, which is rotating at a speed $\omega_{rm,im}$. Based upon the mechanical rotor speed, and the desired electromagnetic torque $T_{e,des}$ (which is determined by the controller governing the mechanical system), the induction motor controls specify the on/off status of each of the inverter semiconductors in such a way that the desired torque is obtained. Although this system is quite robust with regard to over currents, and simple to design from the viewpoint that the controller governing the mechanical system is decoupled from the control of the electrical system (since the torque can be controlled nearly instantaneously), such systems are prone to be subject to a limit cycle behavior in the dc bus voltage known as negative impedance instability [12].

Before setting forth the implementation of the proposed NSSC controller, it is appropriate to first consider a standard field oriented control such as the rotor flux indirect field oriented control illustrated in Figure 2 (note that the control proposed in this paper, is, however, independent of whether or not the field oriented control is direct or indirect). Therein, an instantaneous torque command $T_{e}^*$ is the input to the controller. This torque command is equal to the torque desired by the controller governing the mechanical dynamics, $T_{e,des}$. As can be seen, based on the torque command $T_{e}^*$ and desired d-axis rotor flux level $\lambda_{dr}^*$, the desired q- and d- axis stator currents, $i_{qs}^*$ and $i_{ds}^*$, are determined. This calculation is a function of the induction motor rotor magnetizing inductance $L_m$, the induction motor rotor inductance (rotor leakage plus magnetizing) $L_{rr}'$, the rotor resistance $r_{r}'$, and the number of poles. Based on the q- and d- axis stator currents the electrical radial slip frequency, $\omega_{im}$, is determined, which is then added to the electrical rotor speed $\omega_{r,im}$ in order to determine the electrical speed of the synchronous reference frame $\Theta_{e,im}$. In addition to the algorithm illustrated in Figure 2, especially in large drives, the field oriented control will often include an on line parameter identification algorithm to compensate for variations of the rotor time constant [13-14].

![Figure 1. System Configuration](image1)

![Figure 2. Rotor Flux Oriented Indirect Field Oriented Control.](image2)
Once the q- and d-axis current commands and the position of the synchronous reference frame are established, these currents may be synthesized in a variety of ways. Herein, the q- and d-axis current command was transformed back into an abc variable current command, which is an input to a hysteresis type current control.

System Behavior – The performance of the propulsion system was tested by ramping the desired electromagnetic torque of the induction motor from 2 to 19 Nm over a period of 100 ms. Figure 3 depicts the commanded a-phase current $i^*_{a}$, the actual a-phase current $i_a$, and the dc inverter voltage $v_{dc}$. The increase in torque can be associated with the linear increase in $i_a$. It can be seen that as the power command increases the dc bus voltage becomes unstable, stressing both the semiconductors and the capacitors. In a typical system such behavior could easily result in the semiconductor and/or capacitor failure.

DC POWER SYSTEM - In order to investigate the control of dc power systems, the small but representative system depicted in Figure 4 was utilized, [1]. As can be seen, this 3.7kW system consists of a generation system, a distribution system, and loads. The loads are the dc/dc converter, which is of special concern herein, and a permanent magnet synchronous motor drive.

The generation system consists of a dynamometer that acts as a prime mover, a 3-phase synchronous machine, an LC filter, and a solid state exciter/voltage regulator. The output of the rectifier is filtered by an LC circuit creating a nearly ripple free dc source. The generation system output is connected to a generation bus, $v_{dc1}$, which is attached to the transmission line. The transmission line transfers energy to the load bus, $v_{dc2}$, which distributes power to the remainder of the system.

Figure 4. DC Power system
The dc system has two loads, a permanent magnet synchronous machine drive (brushless dc motor) and a BBC, Figure 5. The synchronous motor drive is used to represent a propulsion load whereas the BBC represents a distribution type of load in which a dc/dc power converter is used to interface between two voltage levels, and/or provide a voltage regulated bus from an unregulated dc source. The operation and parameter values of this motor drive are as set forth in [17] with the exception that the current control was delta modulated at a frequency of 30 kHz rather than hysteresis modulated [18]. The modeling of this type of drive is set forth in [17].

The nominal converter control scheme utilized herein consists of two separate feedback levels. The outer level consists of a regulating nonlinear PI controller, Figure 6, used to maintain a constant output voltage. It consists of second order low pass filters used to eliminate aliasing in the measured inputs and to remove discritization noise in the controller outputs. The nonlinear block following the PI control converts the output current command to an input current command. System components are protected by limiting the range of the commanded input current following the controller. The conditional block is used to limit the valid operating range of the converter based on the level of the distribution bus voltage, providing additional system protection. The regulating PI controller was designed based on the linearized average value model of the system [15], resulting in controller gains of $K_p$ and $K_i$ being equal to 0 and 3.1 respectively.

The inner level consists of a hysteresis current controller that regulates the input current of the converter to within plus or minus a given hysteresis level of the commanded input current. Advantages of using hysteresis current control are that current ripple is independent of operating conditions and the tight regulation of the input current provides for highly effective current limiting. The two main disadvantages in using this type of control are variable switching frequency and an undefined duty cycle, which makes the average value model difficult to derive as can be seen by the almost complete avoidance of this type of control in literature. However, recently an appropriate average value model has been set forth [15 and 16] and is the approach used herein.

System Behavior – Figure 7 depicts the load variables associated with the BBC for a step change in BBC load from 129.3 to 60.1 Ohms. Depicted are the output voltage of the BBC $v_{dc3}$, the distribution bus $v_{dc}$.
voltage $v_{dc2}$, and the current entering the BBC $i_{dcdc}$. It is shown that the system remains stable and recovers. The current entering the BBC is shown limiting as set by the controls but also recovers as $v_{dc2}$ approaches steady state.

Figure 8 depicts the load variables for a study in which the filter capacitance connected to the generation bus is stepped from 1315.5 µF to 1.4 µF. Note how $v_{dc2}$ and $i_{dcdc}$ begin to oscillate violently once the capacitance is removed, demonstrating that the original control cannot operate without significant generation bus capacitance. This is not surprising in view of the fact that the control design assumed the presence of the generation bus capacitance. This raises the question of how much bandwidth would have to be sacrificed in order to eliminate the generation bus capacitance from the system with the standard PI control.

Figure 8. Loss of Generation Bus Capacitance

As it turns out, the instability demonstrated in Figure 8 was found to be uncorrectable by adjustment of the proportional integral control gains. In order to demonstrate this, the system nonlinear average value model (NLAM) was linearized and the loci of system roots were plotted as the gains $k_p$ and $k_i$ were varied between zero and nine million as shown in Figures 9 and 10. Once the generation bus capacitance is removed there are three pole pairs which can contribute to system instability. The shaded regions in Figure 9 (pole pair 1) shows where the first pole pair can be placed while Figure 10 (pole pairs 2 and 3) shows the areas in which one pole of each of the remaining two interesting pole pairs can be placed. Each pole pair can be made stable but it is impossible to move all poles into the left hand plane simultaneously. A similar situation can arise with other converter configurations as well. This would seem to mandate that at least some level of generation bus capacitance must be present. However, this is not the case. The NSSC algorithm introduced in the next section can achieve system stability when it is augmented to the regulating control regardless of the generation bus capacitance without significantly degrading the transient performance.

Figure 9. System Root Loci for Varying $k_p$ and $k_i$

Figure 10. System Root Loci for Varying $k_p$ and $k_i$

NONLINEAR SYSTEM STABILIZING CONTROLLER

The effect of regulating the constant power output of the IMD and BBC results in the input of each device appearing as a constant power, negative impedance load. Negative impedance loads in many power systems have a destabilizing effect known as negative impedance instability. The physical cause of, a measure for prediction, and a means of mitigating this type of instability are set forth in this section.

NEGATIVE IMPEDANCE INSTABILITY - In order to gain insight into the nature of negative impedance instability, consider the highly simplified representation of both systems depicted in Figure 11. Therein, the source is modeled as an ideal source followed by a low pass filter; the constant power load could be considered as either the IMD or the BBC. It is assumed that the IMD as well as the BBC both compensate instantaneously to changes in the dc link voltage, $V_{in}$, allowing them to be modeled as single dependent current sources which are formulated by assuming that the input power is equal to the commanded input power, $P^*$. 
The stability of this simplified system can be determined by calculating its pole locations. The differential equations governing the system can be expressed as
\[ p\dot{i}_{dceq} = v_{dceq} - R_eq i_{dceq} - v_{in} \tag{1} \]
and
\[ p\dot{v}_{in} = i_{dceq} - \frac{P^*}{C_eq} \tag{2} \]
where \( p \) denotes differentiation with respect to time. Linearizing (1) - (2), finding the eigenvalues, and determining the conditions for which they are in the LHP yields the following necessary and sufficient conditions for stability:

1) \[ P^* < \frac{R_eq C_eq v_{ino}^2}{L_eq} \tag{3} \]
and

2) \[ P^* < \frac{v_{ino}^2}{R_eq} \tag{4} \]

Physically speaking, (4) is normally satisfied in practice and so (3) is the most important constraint. It is convenient to state the stability criteria in terms of the small signal input impedance of the constant power load. This impedance is defined as the linearized transfer function between the constant power load input voltage and input current. In particular, for a constant power load
\[ Z_{in} = \frac{-v_{ino}^2}{P^*} \tag{5} \]
Note that in a small signal sense the constant power load appears as a negative resistance, which would suggest a destabilizing effect. In terms of the linear input impedance the stability criteria (3) may be expressed
\[ -Z_{in} > \frac{L_eq}{R_eq C_eq} \tag{6} \]
As \( P^* \) increases, \( -Z_{in} \) decreases and so eventually the system becomes unstable. Equation (6) immediately suggests several methods for manipulating system stability. First, increasing the capacitance to an appropriate level can insure stability. However, such measures can be expensive in terms of capitol, space, weight and reliability. Alternatively, reducing \( L_eq \) is also a means of satisfying (6). However this technique is limited because \( L_eq \) and \( R_eq \) are both tied to the subtransient inductances of the synchronous machine/generator and the ratio is not readily manipulated. Another method is to manipulate the input impedance of the converter/motor drive. This can be accomplished by adding passive filters at the inputs, although this can again be an expensive solution. Herein an NSSC, Figure 12, is investigated which alters the input impedance characteristics and adds no additional cost. For the IMD drive \( i^*_{in} \) and \( i^*_{inc} \) should be replaced by \( T_{edes} \) and \( T^*_c \), respectively.

EFFECT OF STABILIZING CONTROL ON LOAD IMPEDANCE - In this section the effect of the NSSC on the load input impedance is explored. The commanded input current of the converter can be written in terms of the desired input current as shown in Figure 12 or in terms of the desired input power as
\[ i_{inc}^* = \frac{v_{in} (n-1)}{v_{inf}^n} P^* \tag{7} \]
Assuming that the desired power is constant linearizing equation (7) yields
\[ \Delta i_{inc}^* = (n-1) v_{m0} (n-2) \frac{P^*}{v_{inf}^n} \Delta v_{in} \tag{8} \]
\[ + (n-1) v_{inf} (n-1) \frac{P^*}{v_{inf}^{(n+1)}} \Delta v_{inf} \]
From Figure 12,
\[ \Delta v_{inf} = H(s) \Delta v_{in} \tag{9} \]
This filter is designed such that \( H(0) = 1 \). Therefore
Incorporating (9) and (10) into (8) yields

\[ \Delta i_{\text{inc}}^* = \frac{(n-1)P^*}{v_{\text{in}0}^2} - \Delta v_{\text{in}} + \frac{(-n)P^*}{v_{\text{in}0}^2} H(s) \Delta v_{\text{in}} \]  

(11)

The input admittance can then be determined about the operating point, assuming that the actual input current is always equal to the commanded input current. In particular,

\[ Y_{\text{inc}}^*(s) = \frac{\Delta i_{\text{inc}}^*}{\Delta v_{\text{in}}} = \frac{(n-1)P^*}{v_{\text{in}0}^2} + \frac{(-n)P^*}{v_{\text{in}0}^2} H(s) \]  

(12)

Inverting the admittance yields the input impedance:

\[ Z_{\text{inc}}(s) = \frac{v_{\text{inc}}^2}{n(1-H(s))P^*} \]  

(13)

As can be seen, the NSSC offers many possibilities for input impedance control by adjusting \( n \) and \( H(s) \). First, setting \( n \) equal to zero yields

\[ Z_{\text{inc}}(s) = \frac{-v_{\text{in}0}^2}{P^*} \]  

(14)

whereupon it can be seen that the stabilizing control has no effect. If \( n \) is set equal to one, it can be seen that with proper choice of \( H(s) \) the input impedance can be readily manipulated.

\[ Z_{\text{inc}} = \frac{-v_{\text{in}0}^2}{H(s)P^*} \]  

(15)

Setting \( n \) equal to two yields

\[ Z_{\text{inc}} = \frac{v_{\text{in}0}^2}{[1-2H(s)]P^*} \]  

(16)

In this case, and for higher powers, it can be seen that \( n \) acts as a gain on the filter. Although only integer values have been considered herein, \( n \) is in the set of real numbers and does not have to be an integer. It is interesting to observe that if \( H(s) \) is set equal to \(-6.02dB\) over the frequency range in which the stability criteria is failing, infinite input impedance would occur alleviating the problem. If \( H(s) \) continued to get smaller in magnitude the input impedance would then become positive.

**TEST SYSTEMS WITH NSSC**

**INDUCTION MOTOR BASED ELECTRIC PROPULSION SYSTEM** - The advantage of using this simple though nonlinear stabilizing control algorithm is that it is extremely straightforward to implement yet highly effective in mitigating negative impedance instabilities. In order to illustrate the effect of the algorithm on the system, note that using the control law, input power into the inverter is given by

\[ P = \left( \frac{V_{\text{dc}i}}{\bar{V}_{\text{dc}i}} \right)^n P_{\text{des}} \]  

(17)

where

\[ P_{\text{des}} = T_{\text{e},\text{des}} \omega_{\text{rm}} \]  

(18)

From (24) the input current may be expressed

\[ i_{\text{dc}i} = \frac{V_{\text{dc}i}^{n-1}}{\bar{V}_{\text{dc}i}} P_{\text{des}} \]  

(19)

Linearizing (26) about the desired operating point \( (V_{\text{dc}i} = \bar{V}_{\text{dc}i}) \) yields

\[ i_{\text{dc}i} = \frac{1}{n-1} \frac{V_{\text{dc}i}^{n-2}}{P_{\text{des}}} \]  

(20)

If the low pass filter time constant, \( \tau \), (used in determining \( \bar{V}_{\text{dc}i} \)) were so great as to not interact with the dc link dynamics and \( n \) is selected to be unity then the input impedance presented by the inverter is infinite for the frequency range over which negative impedance instabilities occur, thus avoiding this type of instability.

In order to illustrate the effects of varying \( n \) and \( \tau \), consider the case of a system in which \( V_{\text{dc}eq} = 400V \), \( R_{\text{eq}} = 4.58\Omega \), \( L_{\text{eq}} = 13.9mH \), and \( C_{\text{eq}} = 51.4\mu F \). These parameters correspond to a test system that was used for laboratory verification. Figure 13 illustrates the root loci the characteristic equation as \( \tau \) is varied from
As can be seen, in each case the root locus contains an unstable complex pole (denoted \( A \) and \( A^* \)) for small values of \( \tau \) which becomes stable as \( \tau \) is increased. For all \( n \) shown in Figure 13 the real part of the eigenvalues becomes more negative as \( \tau \) is increased. In addition, the complex part also decreases. In the case of \( n = 5 \), eventually the complex pair becomes real (point B) and then one of these real roots meets the root corresponding to the filter at point C, at which this pair of eigenvalues becomes complex. In the case of \( n = 7 \) the two complex poles eventually become real at point D, after which the pair moves away from each other on the real axis.

System Behavior - Incorporating the link stabilizing control into the field oriented control is quite straightforward. In particular, the only difference in the control is that the instantaneous torque command is generated using Figure 11 (with \( i_{in}^*, i_{inc}^*, \) and \( v_{in} \) replaced by \( T_{c,des}^*, T_{c}^* \), and \( v_{dc}^{ci} \) respectively) rather than being set equal to the desired torque, as is illustrated in Figure 2. The study performed on the propulsion system earlier is repeated in Figure 14 except that now the nonlinear stabilizing control is included. The link stabilizing control parameters were set to \( n = 1 \) and \( \tau = 4ms \), based on the root locus generated in Figure 13. As predicted, the dc bus voltage is well behaved and the dc link bus voltage is stable.

One concern which may arise is a possible reduction in torque bandwidth since a drop in inverter voltage will result in a transient dip in torque. Using a detailed computer simulation this effect is depicted in Figures 15 and 16 with and without the stabilizing control, respectively. This study tests the performance of the field oriented control to a step change in commanded torque from 2 to 19Nm. As can be seen, the electromagnetic torque, in Figure 15, reaches the commanded value in approximately 5ms. The torque response is not instantaneous due to the fact that a step change in current cannot be achieved in practice and because the dip in link voltage causes a temporary loss of current tracking in the hysteresis current control. When the stabilizing controller is implemented, Figure 16, the electromagnetic torque reaches the commanded value in the order of 8ms. Although the link stabilized control is somewhat slower than the standard field oriented control, this slight reduction in bandwidth is not a significant disadvantage in view of the improved dc bus voltage. This is particularly true due to the fact that most propulsion systems have mechanical inertia such that in either case the torque response may be considered to be instantaneous.
DC POWER SYSTEM - The NSSC controller used here sets \( n \) equal to one leaving only the time constant in \( H(s) \) to be chosen, assuming a first order low pass filter with unity gain at dc. In order to facilitate a means of making this choice a root locus as \( \tau \) was varied was calculated by linearizing the NLAM model. This was done with the bus filter capacitance effectively removed from the system. The dominant poles were then plotted in the complex plane creating a root locus in terms of the filter time constant, Figure 17. From the root locus a value of \( \tau = 2 \text{ms} \) was chosen that offered significantly high damping but still maintained a fairly high cutoff frequency so that system stability was guaranteed and system performance degradation was minimized. The resulting nonlinear PI controller with the augmented NSSC is illustrated in Figure 18.

System Behavior – The studies presented earlier on the DC power system are repeated in Figures 19 and 20 except that now the stabilizing controller is included. In Figure 19 the almost identical transient response is observed, compared to Figure 7. While in Figure 20 the system remains stable. Notice that the generation bus voltage does undergo increased variation; however this is due to increased rectifier harmonics, since the source filter capacitance has been effectively removed. The stabilizing controller has the desired effect of maintaining system stability for conditions (very low generation bus capacitance) in which it was determined that a proportional integral control alone could not maintain stability.

CONCLUSION

Power systems and electric propulsion systems with dc links are becoming more prominent in industry and the military. This trend will continue as the need for more efficient and versatile methods of moving energy and designing drive systems progress. A simplified model of a constant power load similar to what is typically found in many of these electric power/drive systems was used to develop a stability criteria and a nonlinear system stabilizing controller (NSSC). Verification of the control was accomplished using transient time domain studies on two such systems, a DC power system and electric propulsion system. It was found that the NSSC offered a means to guarantee system stability without sacrificing significant dynamic performance or introducing extra passive components. In addition, the proposed strategy conveniently separates the component regulatory aspects of the control from the negative impedance system stability.

![Figure 17. System Root Locus for Varying \( \tau \)](image)

![Figure 18. Nonlinear PI controller with attached NSSC](image)

![Figure 19. Step in BBC load with NSSC.](image)
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