

Controlling the Direct Connected Parallel Three-Phase Voltage Source Inverters by Using Neural networks.

Dr.Khalid I .Al – Naimi
University Of Technology

R. O. Kadhim
University Of Technology

Abstract:

This paper demonstrates that neural networks can be used effectively for the controlling the direct connection of parallel three-phase voltage source inverters..

A unique feature of the parallel three-phase inverters is a zero-sequence circulating current. This work proposes a new zero-sequence control for parallel three-phase voltage source inverter using a neural networks to assist the Proportional-Integral controller against the variation on the two inverters to maintain the value of a zero-sequence current within an acceptable value. The controller of the zero-sequence current can be implemented within an individual inverter and it is independent of the other control loops of the inverter, therefore, it greatly facilitates design and expansion of parallel system.

Simulation and results can be obtained by using mathematical software such as MATLAB version seven to show the performance of the controllers and the closed loop system.

Two-layers feedforward neural networks FNN's containing a Levenberg-Marquardt training algorithm are used.

الخلاصة

يبين هذا البحث أماكنية استعمال الشبكات العصبية لأغراض السيطرة المباشرة على منظومات عاكسات ثلاثية الطور مربوطة على التوازي. أن أحد خواص العاكس ثلاثي الطور المربوط على التوازي هو جعل التيار الدوار مساويا" صفر. أن هذا البحث يطرح طريقة جديدة للسيطرة على التيار الدوار في منظومة العاكس الثلاثي الطور المربوط على التوازي بالاعتماد على الشبكات العصبية ليكون مساعد" للمسيطر (التناسبي / التكاملي) بالمحافظة على قيم التيار الدوار لتكون ضمن القيم المقبولة.

أن السيطرة على التيار الدوار يمكن تنفيذه ضمن كل عاكس ، دون الاعتماد على حلقات السيطرة الأخرى وهذا الأسلوب يسهل إمكانية توسيع عدد منظومات التوازي. بالامكان الحصول على نتائج المحاكات باستعمال برامج رياضية مثل (ماتلاب ٧) لبيان أداء المسيطر ومنظومة الحلقة المغلقة. تم استعمال شبكة عصبية ذات طبقتين من نوع التغذية الأمامية تحتوي على خوارزمية ليفنبرغ- ماركواردت لغرض تعليم الشبكة.



Introduction:

Artificial neural networks ANN consist of highly interconnected simple processing element called neurons. A block diagram of a typical neuron is given in Fig.(1).

ANN's can be placed into one of three classes based on their feedback link connection structure [1]; *recurrent* (global feedback connection. e.g. Hopfield neural networks, *locally recurrent*.

A special type of nonrecurrents ANN is the feedforward neural network, or FNN, which consists of layers of neurons with weighted links connecting the outputs of neurons in one layer to the inputs of neurons in the next layer, Fig.2, illustrates the block diagram of a three-layer FNN. The use of parallel three-phase PWM inverter dates back to the late 1980 in motor drivers by *Hashii et al.* [2], and uninterruptible power supply(UPS)applications by *Kawabata* [3] and *Holtz et al.* [4]. The distinguishing feature of parallel three-phase inverters is a zero-sequence circulating current in the implementations without isolation. To avoid the zero-sequence circulating current, the following three approaches are commonly used in present technology.

Kamel [5], *Matsui et al* [6], and *Ueda et al.* [7] used a high impedance inter-phase reactor to provide a high zero-sequence impedance. However, the reactors provide high impedance only at medium and high frequencies and they can not prevent a low-frequency circulating current. The reactor can introduce zero-sequence current if they are not centre-taped. Therefore with the inter phase reactor approach, a zero-sequence current control must be applied to the system.

Lol et al. [8] used a transformer-isolated AC-side when discussing a series of PWM methods of multiple inverter to be adjustable to frequency drive. With these two approaches the overall parallel system is bulky and expensive because of additional power supplies or the AC-line frequency transformer. *Ieto* [9] and *Lee* [10] used a separate DC power supply when they deal with parallel operation of inverters.

Ogasawara et al. [11], *Sukegawa* [12], *Chandorkar* [13] and *Matakas*[14]introduced one converter approach. The parallel converters are basically controlled as one converter. This approach has two implementations, one is to use redundant switching vector introduced by *Ogasawara et al.* [11], *Chandorkar*[13], and *Zhao* [15]. The other is to equalize the current between individual phases of the parallel inverter introduced by *Sukegawa*[12], *Chandorkar* [13], and *Fukuda* [16].

Xing et al. [17] introduced an interleaved discontinuous Space- Vector Modulation (SVM) .which can cause a beat-frequency zero-sequence current.

All the existing controllers using Artificial Neural Networks are applied to the single three phase inverters. The high power inverter such as a three-phase inverters is recommended to operate at low frequencies due to the switching losses. In addition, modeling errors and non-linear loads' with unmodeled dynamics frequently degrade the controller performance resulting in a poor transient response and instability

Burtoon et al. [18] used Artificial Neural Networks (ANNs), which have no off-line pretraining and which can be trained continually on-line to identify an inverter-fed induction motor and control its stator currents.

Cabrera et al [19] used an Artificial Neural Networks (ANNs) to accomplish tuning of the stator resistance of an induction motor. The parallel recursive prediction error and back propagation training algorithms were used in training the neural network for the simulation and experimental result. *Carati et al*[20]presented a Robust Model Reference Adaptive Controller (RMAC) for a single three-phase system un interruptible power supplies.

Crabowski et al.[21] presented the concept and implementation of a new simple direct-torque neuro-control scheme *for* pulse width modulation-inverter-fed *motor* drive.

Sha [22] proposed a Neural Network Robust Controller (NNRC) *to* enhance the robustness of the conventional feedback controller. *Denai and Attia*[23] presented some design approaches *to* hybrid control systems combining conventional control techniques with neural networks. Such a mixed implementation leads *to* a more effective control design with improved system performance robustness. *Elbuluk et al.* [24] applied a Model Reference Adaptive Systems (MRASs) based on the neural networks in the adaptation mechanism for control of a permanent-magnet synchronous motor drive. *Liu et al.* [25] presented a mixed signal feed forward neural network chip with on-chip error-reduction hardware for real-time adaptation. *Sharmeela et al.* [26] used a neural based Proportional Integral (PI) control for active power filters for a single phase inverter system. The PI controller is used to shape the current through the filter inductor such that the line current is in shape with and *of* the same shape as the input voltage. *Rech et el.* [27] proposed a repetitive controller with parameters tuned by a Robust Controller *for* Un interruptible power supply applications.

Mathematical Modeling for Parallel Three-Phase Voltage Source Inverters

The voltage source inverter is usually classified as current bidirectional converter because it shares the same switching cells that are current bidirectional. By turning on either the upper *or* the lower (IGBT) switches in one leg of this inverter it is possible to impose a positive or a negative voltage on the corresponding ac phase. The fast switching capability *of* the IGBT allows creating a Pulse Width Modulated within the limits given by the dc voltage. Thus, the inverter can operate in all four quadrants of the P-Q plane.

Three- phase voltage source inverter is shown in Fig (3).The load is taken to be only a resistive to simplify the modelling and discussion

The switching cells either inherently have anti-parallel diodes, for example IGETs and MOSFETs, *or* have external anti-parallel diodes, for example, GTOs with anti-parallel diodes. The symbolic representation and its voltage and current operational states *are* shown in Fig.(4). The symbol normally stands for an IGST device.

In this paper, the average large-signal models and small-signal models of the parallel three phase voltage source inverter are developed. The switching network averaging is performed on phase-leg basis. In the conventional method, the averaging for a three-phase current-bidirectional converter is based on phase-to-phase (or line-to-line) averaging, which intentionally neglects common-mode components[28]. The common-mode components are generally of no interest in the control design for a single three-phase converter. The phase-leg averaging adopts a reference point in the system, then derives the average values of all other points referring to the reference point. As a result, it allows the model to preserve the common-mode components. After the phase-leg averaging, the average model of a three-phase current-bidirectional converter can be easily obtained by connecting three average phase legs.

The choice of the reference point does not have any particular purpose except for a simple representation for modelling. If another point is chosen, all voltage potentials will be shifted by a constant value.

Fig.(5) shows the parallel three- phase voltage source inverters. Assuming the parallel inverter systems double the power rating, then the output capacitance becomes $2C$, and the output resistance becomes $R/2$, where C and R are the parameters of a single inverter as in Fig.(3). It can be seen in Fig(5), that more than one circulating current path exists. Instead of looking at all individual circulating currents a zero-sequence current is defined to represent the overall circulating currents. The zero- sequence current is defined as the sum of phases currents in one inverter, I_z referring to Fig(3).

$$I_z = I_a + I_b + I_c = I_p + I_n \quad (1)$$

In a single inverter operation, the zero-sequence current is always zero because physically there is no such current path. In parallel inverters, however, the current is no longer always zero because of the circulating current paths. Therefore, appropriate controls have to be applied in order to suppress the zero-sequence current because of its adverse effects, such as additional conduction losses, overrun device rating, distorted waveform, etc..

Average Models:

The switching network averaging is performed on a phase-leg basis to develop the large signal models and small signal models of the three-phase voltage source inverters.

Phase-leg Averaging.

The switching cell can be described by a generic switches, as shown in Fig.(6).

When s is open, that is, neither the switch itself nor the antiparallel diode conducts, then the current i is zero. When s is closed, that is, either the switch itself or the anti- parallel diode conducts, the voltage v is then zero. Therefore, a switching function s can be defined as follows:

$$S = \begin{cases} 0 & i = 0, \text{ if switch is open} \\ 1 & v = 0, \text{ if switch is close} \end{cases} \quad (2)$$

In the current-bidirectional switch based inverters, a generic switching unit is called a phase leg and it can be identified as shown in Fig.(7).

The phase leg is composed of two switching cells, and has a voltage source (or a capacitor) on one side and a current source (or an inductor) on the other. These features make the phase-leg a generic switching unit. There are switching constraints on the two switching cells of the phase-leg. The constraints include that voltage sources or capacitors cannot be short-circuited, and current sources or inductors cannot be open-circuited. The constraints result in a requirement that the two switching cells of the phase-leg are complementary. That is, to prevent the voltage source (or the capacitor) from being short- circuited, only one, can be closed at any time. $s\phi n$ and $s\phi p$ of the two switching cells, Meanwhile, to prevent the inductor from being open-circuited, one of the two switching cells has to be closed at any time. Based on the switching function defined, this complementary relationship can be described as:

$$s\phi p + s\phi n = 1 \quad (3)$$

As a result, the phase-leg can be represented by a single-pole, double-throw switch as shown in Fig(8). The input and output variable of interest are also defined in Fig(8). Because the current of the positive and negative DC rails are not necessity equal in the parallel converters, i_p and i_n are defined as the current of the positive and negative DC rails, respectively,

The PWM of the phase-leg is shown in Fig.(9) where T is the switching period and $d\phi T$ is defined as the duty cycle of the top switch $s\phi p$. The corresponding voltage and current waveform are also shown in fig.(9) . Based on the waveform, one can obtain the voltage and current relationships in average, assuming the current $i\phi$ and the voltage vdc are continuous with small ripples;

$$v\phi = d\phi.vdc \quad (4)$$

$$i_p = d\phi.i\phi \quad (5)$$

the average mode of the phase – leg is depicted in fig.(10)

Average Model of Parallel Voltage Source Invertors

The average model of a three-phase voltage source inverter can be obtained by connecting three averaged phase legs and the rest of the circuit component as shown in Fig.(11), where :

$$I_p = d_a.i_a + d_b.i_b + d_c.i_c \quad (6)$$

$$I_n = i_a + i_b + i_c - i_p = i_z - i_p \quad (7)$$

The state – space equations of the voltage source inverter are :

$$\frac{d}{dt} \begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix} = \frac{1}{L} \begin{bmatrix} d_a \\ d_b \\ d_c \end{bmatrix} v_{dc} - \frac{1}{L} \begin{bmatrix} v_{an} \\ v_{bn} \\ v_{cn} \end{bmatrix} - \frac{1}{L} \begin{bmatrix} v_n \\ v_n \\ v_n \end{bmatrix} \quad (8)$$

$$\frac{d}{dt} \begin{bmatrix} v_{an} \\ v_{bn} \\ v_{cn} \end{bmatrix} = \frac{1}{C} \begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix} - \frac{1}{RC} \begin{bmatrix} v_{an} \\ v_{bn} \\ v_{cn} \end{bmatrix} \quad (9)$$

It can be seen from Fig.(11), the zero sequence current is always zero because physically there is no such current path, In parallel operation, however, a circulating current path is formed, as shown in Fig,(12).

The state – space equations of the voltage source inverter are :

$$\frac{d}{dt} \begin{bmatrix} i_{a1} \\ i_{b1} \\ i_{c1} \end{bmatrix} = \frac{1}{L_1} \begin{bmatrix} d_{a1} \\ d_{b1} \\ d_{c1} \end{bmatrix} v_{dc} - \frac{1}{L_1} \begin{bmatrix} v_{an} \\ v_{bn} \\ v_{cn} \end{bmatrix} - \frac{1}{L_1} \begin{bmatrix} v_n \\ v_n \\ v_n \end{bmatrix} \quad (10)$$

$$\frac{d}{dt} \begin{bmatrix} i_{a2} \\ i_{b2} \\ i_{c2} \end{bmatrix} = \frac{1}{L_2} \begin{bmatrix} d_{a2} \\ d_{b2} \\ d_{c2} \end{bmatrix} v_{dc} - \frac{1}{L_2} \begin{bmatrix} v_{an} \\ v_{bn} \\ v_{cn} \end{bmatrix} - \frac{1}{L_2} \begin{bmatrix} v_n \\ v_n \\ v_n \end{bmatrix} \quad (11)$$

$$\frac{d}{dt} \begin{bmatrix} v_{an} \\ v_{bn} \\ v_{cn} \end{bmatrix} = \frac{1}{2C} \begin{bmatrix} i_{b1} \\ i_{c1} \end{bmatrix} + \frac{1}{2C} \begin{bmatrix} i_{a2} \\ i_{b2} \\ i_{c2} \end{bmatrix} - \frac{1}{RC} \begin{bmatrix} v_{an} \\ v_{bn} \\ v_{cn} \end{bmatrix} \quad (12)$$

In steady state, the output AC phase voltage is regulated as balanced sinusoidal voltage as in equation (14).

$$\begin{bmatrix} v_{an} \\ v_{bn} \\ v_{cn} \end{bmatrix} = \begin{bmatrix} v_m \cos(\omega t) \\ v_m \cos(\omega t - \frac{2\pi}{3}) \\ v_m \cos(\omega t + \frac{2\pi}{3}) \end{bmatrix} \quad (13)$$

In order to obtain a DC steady- state operating point so as to linearize the system to design controllers, the model in the stationary coordinate is usually transferred in to rotating coordinates. The transformation matrix is chosen as follows :

$$T = \sqrt{\frac{2}{3}} \begin{bmatrix} \cos \omega t & \cos(\omega t - \frac{2\pi}{3}) & \cos(\omega t + \frac{2\pi}{3}) \\ -\sin \omega t & -\sin(\omega t - \frac{2\pi}{3}) & -\sin(\omega t + \frac{2\pi}{3}) \\ \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} \end{bmatrix} \quad (14)$$

:

where ω is chosen as the same frequency as the AC line frequency, then T is an orthogonal matrix.

The variable in the stationary coordinates X_{abc} can be transformed into the rotating coordinates X_{dqz} using :

$$X_{dqz} = T \cdot X_{abc} \quad (15)$$

Applying (15) to (8) and (9), we can obtain the average model of the voltage source inverter in the rotating coordinates :

$$\frac{d}{dt} \begin{bmatrix} i_d \\ i_q \\ i_z \end{bmatrix} = \frac{1}{L} \begin{bmatrix} d_d \\ d_q \\ d_z \end{bmatrix} v_{dc} - \frac{1}{L} \begin{bmatrix} v_d \\ v_q \\ v_z \end{bmatrix} - \frac{1}{L} \begin{bmatrix} 0 \\ 0 \\ 3v_n \end{bmatrix} - \begin{bmatrix} 0 & -\omega & 0 \\ \omega & 0 & 0 \\ 0 & 0 & 0 \end{bmatrix} \begin{bmatrix} i_d \\ i_q \\ i_z \end{bmatrix} \quad (16)$$

$$\frac{d}{dt} \begin{bmatrix} v_d \\ v_q \\ v_z \end{bmatrix} = \frac{1}{C} \begin{bmatrix} i_d \\ i_q \\ i_z \end{bmatrix} - \frac{1}{RC} \begin{bmatrix} v_d \\ v_q \\ v_z \end{bmatrix} - \begin{bmatrix} 0 & -\omega & 0 \\ \omega & 0 & 0 \\ 0 & 0 & 0 \end{bmatrix} \begin{bmatrix} v_d \\ v_q \\ v_z \end{bmatrix} \quad (17)$$

where:-

$$\begin{bmatrix} i_d \\ i_q \\ i_z / \sqrt{3} \end{bmatrix} = T \cdot \begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix}, \quad \begin{bmatrix} v_d \\ v_q \\ v_z / \sqrt{3} \end{bmatrix} = T \cdot \begin{bmatrix} v_{an} \\ v_{bn} \\ v_{cn} \end{bmatrix}, \quad \begin{bmatrix} d_d \\ d_q \\ d_z / \sqrt{3} \end{bmatrix} = T \cdot \begin{bmatrix} d_a \\ d_b \\ d_c \end{bmatrix}. \quad (18)$$

And v_z and d_z are defined as:

$$v_z = v_{an} + v_{bn} + v_{cn}, \quad d_z = d_a + d_b + d_c. \quad (19)$$

since $i_z = 0$, in the single inverter. The channel equation is normally dropped from the model. Therefore, the average model of the single voltage source inverter becomes:

$$\frac{d}{dt} \begin{bmatrix} i_d \\ i_q \end{bmatrix} = \frac{1}{L} \begin{bmatrix} d_d \\ d_q \end{bmatrix} v_{dc} - \frac{1}{L} \begin{bmatrix} v_d \\ v_q \end{bmatrix} - \begin{bmatrix} 0 & -\omega \\ \omega & 0 \end{bmatrix} \begin{bmatrix} i_d \\ i_q \end{bmatrix} \quad (20)$$

$$\frac{d}{dt} \begin{bmatrix} v_d \\ v_q \end{bmatrix} = \frac{1}{C} \begin{bmatrix} i_d \\ i_q \end{bmatrix} - \frac{1}{RC} \begin{bmatrix} v_d \\ v_q \end{bmatrix} - \begin{bmatrix} 0 & -\omega \\ \omega & 0 \end{bmatrix} \begin{bmatrix} v_d \\ v_q \end{bmatrix} \quad (21)$$

Applying (15) to (10) - (12), we can obtain the average model of the parallel voltage source inverter in the rotating coordinates:

$$\frac{d}{dt} \begin{bmatrix} id1 \\ iq1 \\ iz1 \end{bmatrix} = \frac{1}{L1} \begin{bmatrix} dd1 \\ dq1 \\ dz1 \end{bmatrix} v_{dc} - \frac{1}{L1} \begin{bmatrix} vd \\ vq \\ vz \end{bmatrix} - \frac{1}{L1} \begin{bmatrix} 0 \\ 0 \\ 3v_n \end{bmatrix} - \begin{bmatrix} 0 & -\omega & 0 \\ \omega & 0 & 0 \\ 0 & 0 & 0 \end{bmatrix} \begin{bmatrix} id1 \\ iq1 \\ iz1 \end{bmatrix} \quad (22)$$

$$\frac{d}{dt} \begin{bmatrix} id2 \\ iq2 \\ iz2 \end{bmatrix} = \frac{1}{L2} \begin{bmatrix} dd2 \\ dq2 \\ dz2 \end{bmatrix} v_{dc} - \frac{1}{L2} \begin{bmatrix} vd \\ vq \\ vz \end{bmatrix} - \frac{1}{L2} \begin{bmatrix} 0 \\ 0 \\ 3v_n \end{bmatrix} - \begin{bmatrix} 0 & -\omega & 0 \\ \omega & 0 & 0 \\ 0 & 0 & 0 \end{bmatrix} \begin{bmatrix} id2 \\ iq2 \\ iz2 \end{bmatrix} \quad (23)$$

$$\frac{d}{dt} \begin{bmatrix} vd \\ vq \\ vz \end{bmatrix} = \frac{1}{2C} \begin{bmatrix} id1 \\ iq1 \\ iz1 \end{bmatrix} + \frac{1}{2C} \begin{bmatrix} id2 \\ iq2 \\ iz2 \end{bmatrix} - \frac{1}{RC} \begin{bmatrix} vd \\ vq \\ vz \end{bmatrix} - \begin{bmatrix} 0 & -\omega & 0 \\ \omega & 0 & 0 \\ 0 & 0 & 0 \end{bmatrix} \begin{bmatrix} vd \\ vq \\ vz \end{bmatrix} \quad (24)$$

For the single inverter, the zero sequence current is always zero because there is no such current path. With the parallel operation, however, a circulating current path is formed as shown in Fig.(12) and noting that:

$$iz = iz1 = - iz2 \quad (25)$$

The model of parallel inverters can be simplified based on equation(3.25) and

$$\Delta dz = dz1 - dz2 \quad (26)$$

$$\frac{d}{dt} \begin{bmatrix} id1 \\ iq1 \end{bmatrix} = \frac{1}{L1} \begin{bmatrix} dd1 \\ dq1 \end{bmatrix} v_{dc} - \frac{1}{L1} \begin{bmatrix} vd \\ vq \end{bmatrix} - \begin{bmatrix} 0 & -\omega \\ \omega & 0 \end{bmatrix} \begin{bmatrix} id1 \\ iq1 \end{bmatrix} \quad (27)$$

$$\frac{d}{dt} \begin{bmatrix} id2 \\ iq2 \end{bmatrix} = \frac{1}{L2} \begin{bmatrix} dd2 \\ dq2 \end{bmatrix} v_{dc} - \frac{1}{L2} \begin{bmatrix} vd \\ vq \end{bmatrix} - \begin{bmatrix} 0 & -\omega \\ \omega & 0 \end{bmatrix} \begin{bmatrix} id2 \\ iq2 \end{bmatrix} \quad (28)$$

$$\frac{dz}{dt} = \frac{\Delta dz \cdot v_{dc}}{L_1 + L_2} \quad (29)$$

$$\frac{d}{dt} \begin{bmatrix} v_d \\ v_q \end{bmatrix} = \frac{1}{2C} \left(\begin{bmatrix} i_{d1} \\ i_{q1} \end{bmatrix} + \begin{bmatrix} i_{d2} \\ i_{q2} \end{bmatrix} \right) - \begin{bmatrix} \frac{1}{RC} & -\omega \\ \omega & \frac{1}{RC} \end{bmatrix} \begin{bmatrix} v_d \\ v_q \end{bmatrix} \quad (30)$$

Neuro- Control Design Results

In this section, the design of the controllers are based on the plant system equations (27)-(30) and the neural networks used are FNN containing on- line (Levenberg Marquardt) training algorithm. The simulation results obtained with help of the MA TLAB software's (SIMULINK and MFILE programs commands).

The general controller architectures are as shown in Fig.(13). The feed-forward controller containing FNNM and a conventional PI controller are used as a mixed controller for the parallel plant model. The conventional PI controllers are used as a voltage and current compensators. The voltage compensator usually has an integrator in order to have a zero steady-state error and current loop also has an Proportional-Integral controller in order to guarantee the unity displacement factor.

Fig.(14) , Fig.(15), and Fig.(16) illustrate the architectures used to evaluation the plant inputs $ddl, dq1, dd2, dq2$, and Δdz . Where $dz = dz_1 - dz_2$. Fig.(17) and Fig.(18) show that the performance of PI controller. only and the performance of mixed (PI + FNNM) controllers corresponding to the reference phase voltage V_{an} .respectively.

Fig.(19) and Fig.(20) illustrate the inverters currents corresponding to the performance of PI controller only and the performance of mixed (PI+FNNM) controllers respectively.

Fig.(21) and Fig (22) show that the sum-squared error between the inverter currents corresponding to the performance of PI controller and mixed (PI+FNNM) controllers respectively.

These , results show the improved performance due to the use of {PI + FNNM} controller as compared with the conventional PI controller as the error difference of the currents is greatly reduced .

Conclusions:

The major accomplishments and some conclusions are summarized below:

This work illustrates the modeling of the parallel three-phase voltage source inverters based on averaging model. This model is equivalent to the conventional three-phase inverter and preserves common-mode information which is critical in the analysis of controlling the parallel inverters.

As part of the parallel models, a unique zero-sequence channel has been explicitly shown. It is found that the zero-sequence current is governed by the difference in the common-mode voltage between the two inverters. Because of the direct parallel operation, a zero-sequence current is formed, as a result, the common-mode voltage has to be controlled so that the zero-sequence current can be minimized.

Artificial Neural Networks (ANN) which have off-line pretraining can be trained continually on – line to identify the inverse model of the plant and use this inverse model as a feedforward controller to provide a good action to improve the overall performance of the PI controller.

Using a mixed controller such as (PI+FNNM) leads to a more effective control design with improved system performance and robustness. While the conventional controller allows different design objectives such as steady state and transient characteristics of the closed loop system to be specified, neural networks are integrated to overcome the problems with uncertainties in the plant parameters.

Finally the use of the neural networks in controlling processing requires a long time via conventional techniques for training method and such controllers are available to high power converters which operate at low frequencies to avoid the switching losses.

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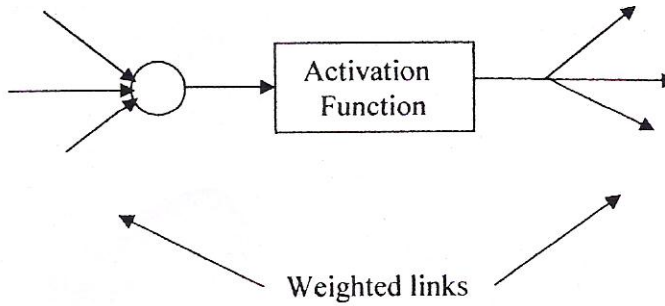


Fig.1. Block diagram of a neuron.

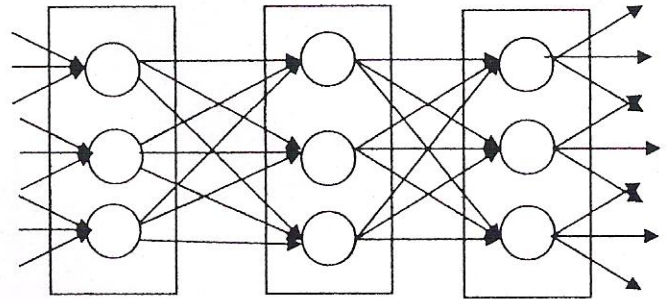


Fig.2. Block diagram of a three-layer FNN.

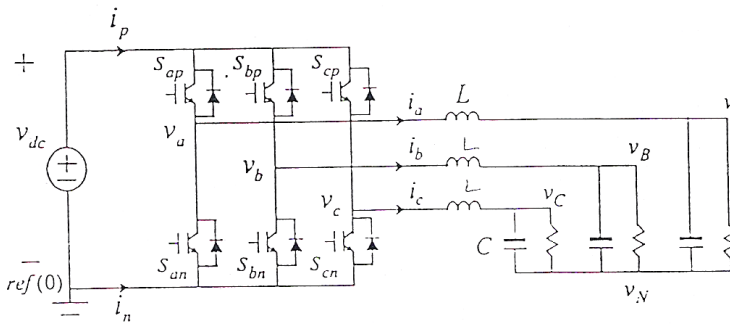


Fig. (3) Three-phase voltage source inverter

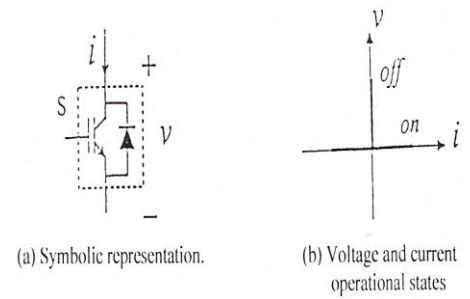


Fig. (4) Current-bidirectional switching cell

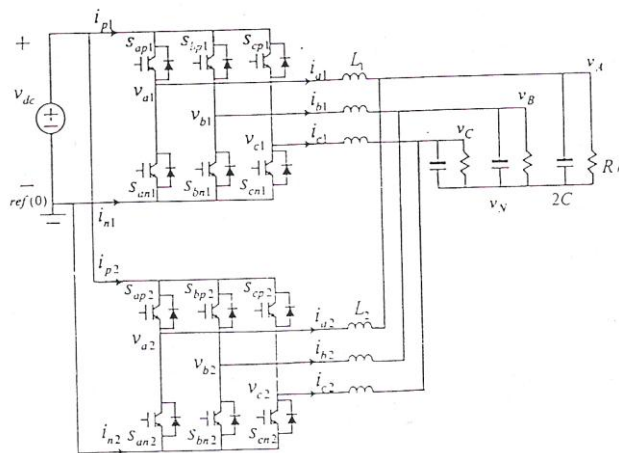


Fig.(5) Parallel three-phase voltage source inverters.

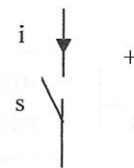


Fig.(6) Generic switch

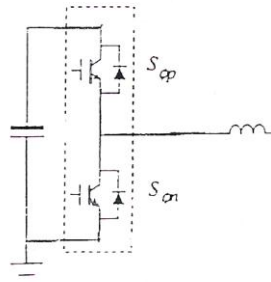


Fig.(7) Generic phase-leg in current-bidirectional inverter

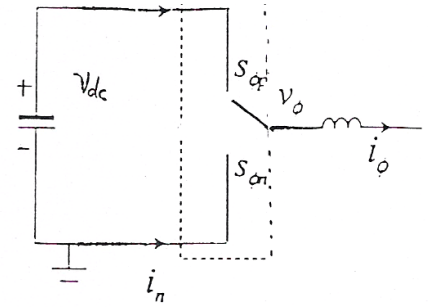


Fig.(8) Phase-leg represented as a single-pole, double-throw switch

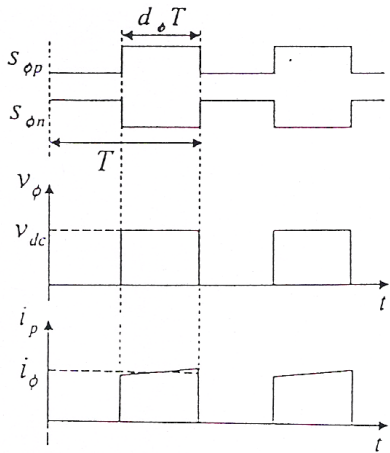


Fig.(9) Phase-leg PWM and corresponding voltage and current waveforms.

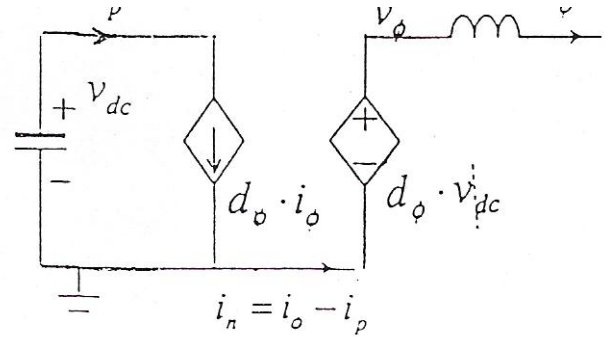


Fig.(10) Phase-leg average model.

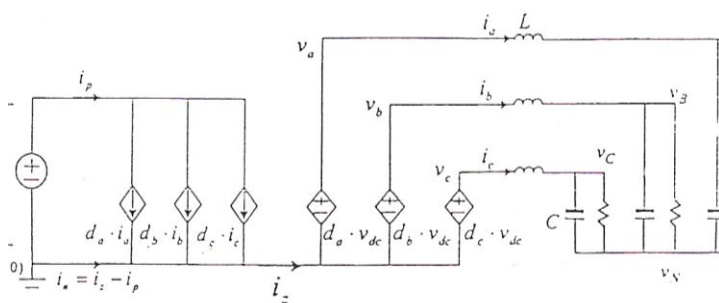


Fig.(11) Voltage-source inverter's average model in stationary coordinates.

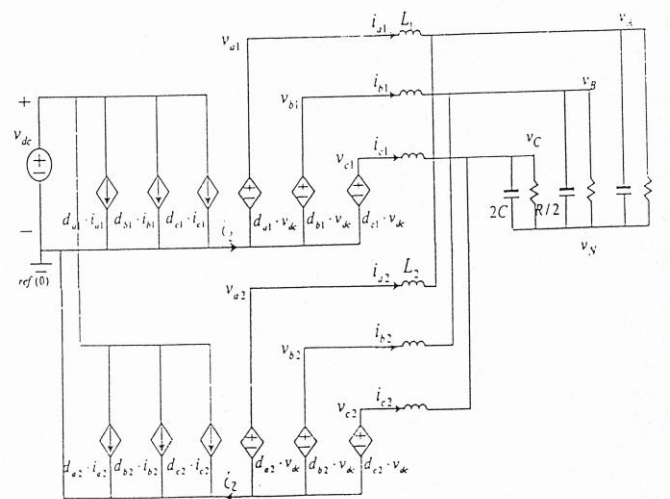


Fig.(12) Parallel voltage-source inverters' average model in stationary coordinates.

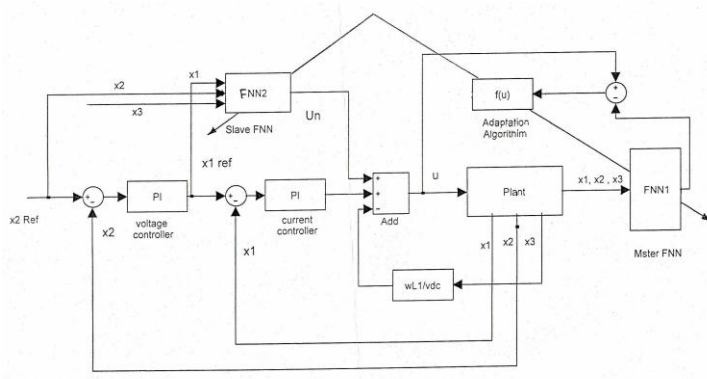


Fig.(13) General controller structure

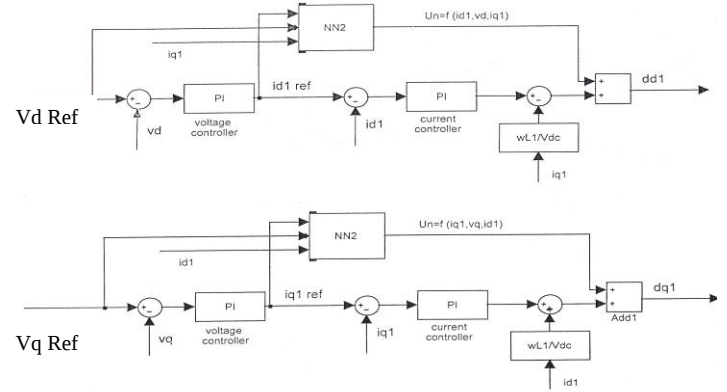


Fig (14) Evaluation of dd1 and dq1

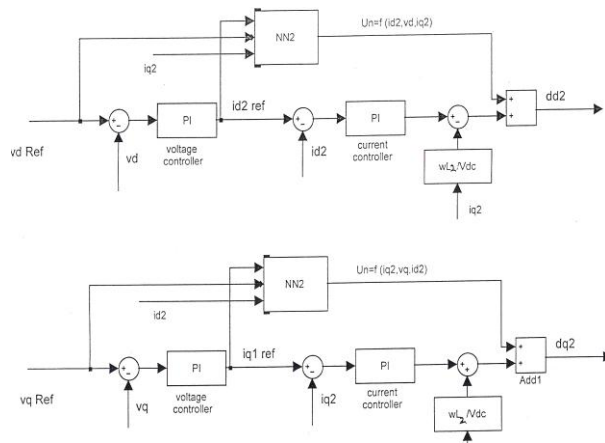


Fig.(15) Evaluation of dd2 and dq2

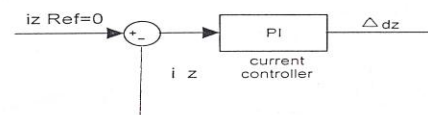


Fig.(16) Evaluation of Δdz

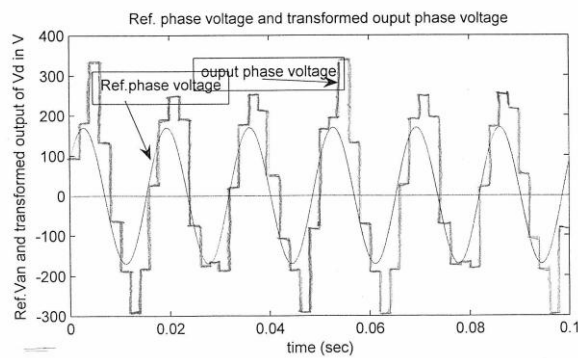


Fig.(17) Reference and actual phase voltages corresponding to performance of PI controller.

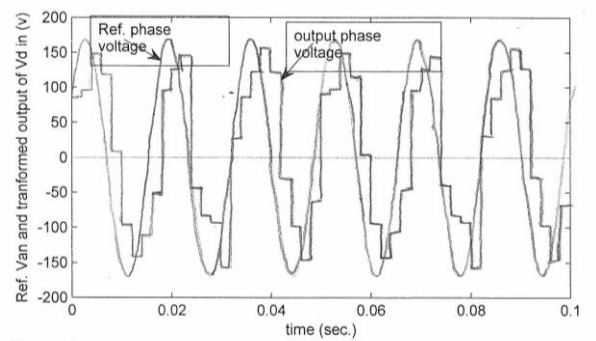


Fig.(18) Reference and actual phase voltages corresponding to performance of (PI+FNNM) controllers.

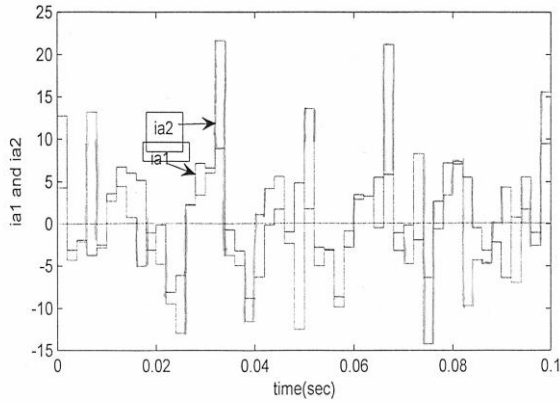


Fig.(19 .) Inverters currents corresponding to performance of PI controller.

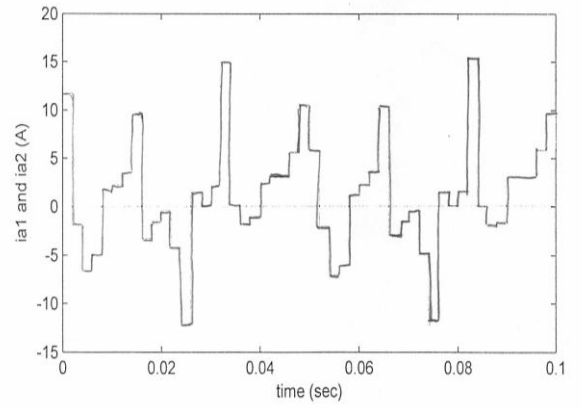


Fig.(20 .) Inverters currents corresponding to performance of (PI+FNNM) controllers

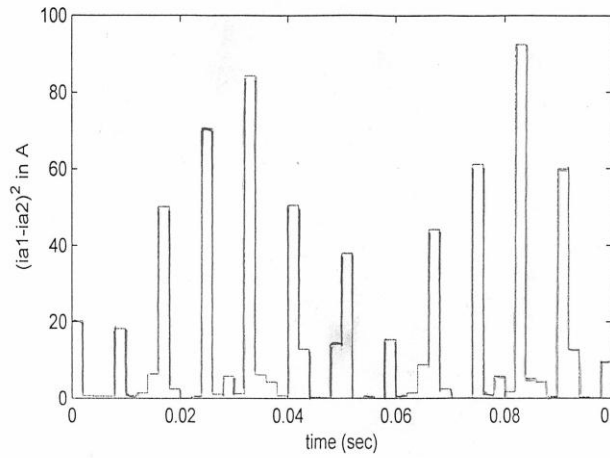


Fig.(21 .) Sum-Squared error between inverters currents correspond to performance of PI controller.

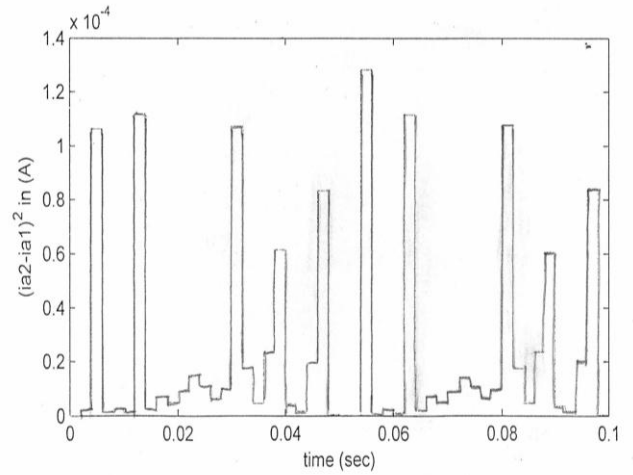


Fig.(22 .) Sum-Squared error between inverters currents corresponding to performance of (PI + FNNM) controllers.