Digital non-interleaved high-power totem pole PFC based on double integral sliding mode

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Abstract: A PFC circuit for applications higher than 2 kW is generally configured by an interleaved structure, which reduces component stress and the ripple of inductor currents and output voltage. However, when implemented with GaN HEMT and controlled by a double sliding mode, a high-power totem pole PFC could achieve better performance without an interleaved structure.

Keywords: PFC, GaN HEMT, totem pole, double sliding mode

Classification: Power devices and circuits

References


1 Introduction

A linear small signal model is convenient for modeling and analysis; however, due to the hypothesis of a small signal, its accuracy is unsatisfactory when the PFC works over the entire load range of an operation and across the universal input voltage range. Moreover, the small signal model fails to present the stability characteristics of the whole work area of the PFC. According to [1], a system whose stability is derived from a small signal model may be unstable. Sliding mode control, when applied in a variable structure control system featuring strong robustness and stability in the presence of parametric uncertainty [2], is suitable for a nonlinear system such as a switch power supply. It should be mentioned that switch components of power supplies are incapable of achieving infinite switch frequency, so chattering is unavoidable. To reduce the chattering, a high-switch frequency functions admirably.

Compared with Si MOSFET, GaN HEMT’s advantages can be summarized as a smaller driving loss, a smaller switching loss, a smaller reverse recovery loss and a smaller voltage oscillation, which yields high switch frequency without sacrificing efficiency. These characteristics make GaN HEMT suitable for applying to a totem pole PFC controlled by a sliding mode.

Researchers and engineers have made efforts to improve the performance of PFCs, primarily paying attention to high efficiency, high power density and total harmonic distortion (THD) [3, 4, 5, 6, 7, 8] related to the power factor and EMI. The trend in PFC development is to apply fewer components while achieving better performance and satisfying the proposed requirements. Among all PFC topologies,
totem pole PFC has the fewest components, which means the least conduction loss. However, before the industry application of GaN HEMT, totem pole PFC was hardly ever used in practice due to the unavoidable drawbacks of Si MOSFET. The large switch loss and parasitic ringing derived from the reverse recovery effect of the body diode limits the application of totem pole PFC with Si MOSFET to low power levels and low frequency in either discontinuous conduction mode (DCM) or critical conduction mode (CRM) operation rather than to the high power levels and high frequency in hard switching CCR that are required by high-power PFC.

Interleaved structure, which is usually applied to PFCs that are more than 2 kW [9, 10, 11], consists of two or more sets of boost converter cells, which operate in phase-shift mode. As a result, the superposition of current waveforms has lower ripple than each individual current waveform generated by a single boost converter cell. This technique reduces not only component stress and volume but also current ripple and EMI [12, 13]; however, it also increases the number of components and the volume of PFC and reduces power density by adding one or more sets of boost converter cells. Nevertheless, totem pole PFC with GaN HEMT controlled by DSP can be made without interleaved structure, even when the power is as high as 2.4 kW, due to low EMI caused by the smaller voltage oscillation of GaN HEMT and the low current ripple caused by high switch frequency. Compared with traditional interleaved PFC with Si MOSFET, non-interleaved totem pole PFC with HEMT has better performance in such areas as efficiency, THD and power density. A 2.4-kW prototype was made to verify the theoretical analysis; the peak efficiency is 99.07% (50 kHz), and the minimum values of THD are 2.8% (115 V) and 2.7% (230 V). In addition to high efficiency and low THD, totem pole PFC with GaN HEMT could achieve high power density; according to [14], 130 W per inch$^3$ has been obtained by a 3.2 kW prototype.

2 Design of algorithm

The transfer function of a boost converter working in continuous conduction mode (CCM) has zero points in the right half plane (RHPZ), which causes the dynamic response of the controlled system to be hysteretic, especially when controlled only by voltage. Current control is usually applied to improve the dynamic response of an RHPZ system. Although integral control could improve steady error, sliding mode with integral control could hardly ever do a satisfactory job in improving steady error [15, 16]. Usually, increasing the control order of a system contributes to improving steady error, so double integral items would be introduced. The topology and algorithm of totem pole PFC are shown in Fig. 1. $Q_1$ and $Q_2$ are GaN HEMTs, and T1 and T2 are Si MOSFETs.

![Fig. 1. Topology and algorithm of totem pole PFC](image-url)
The analysis of the case in which \( v_i < 0 \) is similar to the analysis when \( v_i > 0 \): when \( v_i < 0 \), we obtain the absolute value of \( v_i, i_L \). In this paper, we only analyze the case in which \( v_i > 0 \). When we have \( v_i > 0 \), \( T_2 \) on, \( Q_1 \) on, \( Q_2 \) and off, then \( v_L = v_i - v_o \) and \( u = 1 \); with \( T_2 \) on, \( Q_1 \) off, \( Q_2 \) on, then \( v_L = v_i \), and \( u = 0 \). So \( v_L = v_i - u v_o \).

The controlled state variables are current error \( x_1 \); voltage error \( x_2 \); integral of current and voltage error \( x_3 \); and double integral of current and voltage error \( x_4 \); they are written as follows:

\[
\begin{align*}
  x_1 &= i_{ref} - i_L \\
  x_2 &= V_{ref} - \beta v_o \\
  x_3 &= \int (x_1 + x_2) dt \\
  x_4 &= \int \int (x_1 + x_2) dt
\end{align*}
\]

where \( i_{ref} = K(V_{ref} - \beta v_o) \), \( K = |v_i|/V_{IRMS} \), and \( |v_i| \) and \( V_{IRMS} \) are the absolute values of instantaneous value and RMS of input voltage, respectively. The sliding manifold is expressed as:

\[
S = a_1 x_1 + a_2 x_2 + a_3 x_3 + a_4 x_4
\]

Taking the derivatives of the variables in (1), the result is:

\[
\begin{align*}
  \dot{x}_1 &= -\frac{K\beta}{C} i_C - \frac{v_i - u v_o}{L} \\
  \dot{x}_2 &= -\frac{\beta}{C} i_C \\
  \dot{x}_3 &= (K + 1)(V_{ref} - \beta v_o) - i_L \\
  \dot{x}_4 &= \int [(K + 1)(V_{ref} - \beta v_o) - i_L] dt
\end{align*}
\]

Assume \( \dot{S} = a_1 \dot{x}_1 + a_2 \dot{x}_2 + a_3 \dot{x}_3 + a_4 \dot{x}_4 = 0 \)

Then, the equivalent sliding control law \( u_{eq} \) could be written as:

\[
u_{eq} = \frac{1}{v_o} \left\{ K_1 i_C + K_2 (V_{ref} - \beta v_o) + K_3 \int (V_{ref} - \beta v_o) dt + K_2 [K(V_{ref} - \beta v_o)] - i_L \right\} + K_3 \int [K(V_{ref} - \beta v_o) - i_L] dt + v_i
\]

where \( K_1 = \frac{\beta L}{C} (K + \frac{a_2}{a_1}) \), \( K_2 = -\frac{a_3 L}{a_1} \), \( K_3 = -\frac{a_4 L}{a_1} \).

To guarantee the existence of sliding state, \( \lim_{\delta \to 0} S \cdot \dot{S} < 0 \) should be satisfied; according to (3) and (4), the existence condition is:

\[
\begin{align*}
  \left\{ K_1 i_{C\text{max}} + K_2 (x_{1\text{min}} + x_{2\text{min}}) + K_3 x_{3\text{min}} < v_{o\text{max}} - v_{i\text{max}} \right\} \\
  -K_1 i_{C\text{min}} - K_2 (x_{1\text{max}} + x_{2\text{max}}) - K_3 x_{3\text{max}} < v_{i\text{min}}
\end{align*}
\]

where \( v_{o\text{max}} \) is the steady output voltage and \( i_{C\text{max}} \) and \( i_{C\text{min}} \) are the maximum value and the minimum value of a steady capacitor current. The value of \( v_{i\text{min}} \) is not 0 but 4.5 V; if the input voltage is less than 4.5 V, then the switch components shut off to avoid calculation error, and the capacitor then supplies power to the load.
The variables \(x_{1(\text{max})}\) and \(x_{1(\text{min})}\) are the maximum value and minimum value of the steady current error, \(x_{2(\text{max})}\) and \(x_{2(\text{min})}\) are the maximum value and minimum value of the steady voltage error, and \(x_{3(\text{max})}\) and \(x_{3(\text{min})}\) are the maximum value and minimum value of the steady integral of voltage error and current error. The designed parameters must satisfy equation (6).

The stability of a power converter controlled by double integral sliding could be achieved by guaranteeing that all the eigenvalues of the Jacobian matrix of the system have a negative real part. The stability condition could be derived by two steps: first, deduce the ideal sliding dynamics of the system; second, analyze the stability on its equilibrium [16].

According to the voltage of the inductor and the current of the capacitor of the system, the stability can be derived as:

\[
\begin{align*}
\frac{dI_L}{dt} &= \frac{v_i}{L} - \frac{v_o}{L} u \\
\frac{dv_o}{dt} &= \frac{i_L}{C} u - \frac{v_o}{r_I C}
\end{align*}
\]  

(7)

The ideal sliding dynamics could be obtained by replacing \(u\) with \(u_{eq}\); that is:

\[
\begin{align*}
\frac{dI_L}{dt} &= \frac{v_i}{L} - \frac{v_o}{L} u_{eq} \\
\frac{dv_o}{dt} &= \frac{i_L}{C} u_{eq} - \frac{v_o}{r_I C}
\end{align*}
\]  

(8)

Assume an equilibrium point exists in the sliding manifold; then, stability equation (9) could be derived from (8) when \(\frac{dI_L}{dt} = 0\) and \(\frac{dv_o}{dt} = 0\):

\[I_L = \frac{V_o^2}{V_i R_L}\]  

(9)

According to perturbation theory and referring to (5), (8) and (9), the small signal of ideal sliding dynamics could be linearized on the equilibrium point. Decomposing the signal into AC and DC (capital letters denote DC, lowercase letters with the superscript “\(\sim\)” denote AC), it could be derived as:

\[
\begin{align*}
\frac{d(I_L + \tilde{I}_L) + \tilde{V}_i}{dt} &= \frac{(I_L + \tilde{I}_L) + \tilde{V}_i}{L} - \frac{(V_o + \tilde{V}_o)}{L(V_o + \tilde{V}_o)} \left\{ K_1 C \frac{d(V_o + \tilde{V}_o)}{dt} \right. \\
&+ K_2(K + 1)[V_{ref} - \beta(V_o + \tilde{V}_o)] - K_3(I_L + \tilde{I}_L) + K_3 \int (K + 1)[V_{ref} - \beta(V_o + \tilde{V}_o)] dt - K_3 \int (I_L + \tilde{I}_L) dt \\
&+ \left. K_3 \int (V_o + \tilde{V}_o) dt \right\} \\
\frac{d(V_o + \tilde{V}_o)}{dt} &= \frac{(I_L + \tilde{I}_L)}{C(V_o + \tilde{V}_o)} \left\{ K_1 C \frac{d(V_o + \tilde{V}_o)}{dt} + K_2(K + 1)[V_{ref} - \beta(V_o + \tilde{V}_o)] dt \\
&+ K_3 \int (I_L + \tilde{I}_L) dt + (V_i + \tilde{V}_i) \right\} - \frac{(V_o + \tilde{V}_o)}{R_I C} \right. \\
&+ \left. K_3 \int (I_L + \tilde{I}_L) dt + (V_i + \tilde{V}_i) \right\}
\end{align*}
\]  

(10)

Ignoring the DC items of (10), the signal could be written as:
\begin{align*}
\frac{d\tilde{i}_L}{dt} &= a_{11}\tilde{i}_L + a_{12}\tilde{v}_o + a_{13}\int\tilde{i}_L dt + a_{14}\int\tilde{v}_o dt \\
\frac{dv_o}{dt} &= a_{21}\tilde{i}_L + a_{22}\tilde{v}_o + a_{23}\int\tilde{i}_L dt + a_{24}\int\tilde{v}_o dt \\
\frac{d(\int\tilde{i}_L dt)}{dt} &= a_{31}\tilde{i}_L + a_{32}\tilde{v}_o + a_{33}\int\tilde{i}_L dt + a_{34}\int\tilde{v}_o dt \\
\frac{d(\int\tilde{v}_o dt)}{dt} &= a_{41}\tilde{i}_L + a_{42}\tilde{v}_o + a_{43}\int\tilde{i}_L dt + a_{44}\int\tilde{v}_o dt
\end{align*}

\text{(11)}

where

\begin{align*}
\begin{cases}
a_{11} &= \frac{K_2 - K_1}{L} ; \quad a_{12} = \frac{K_1 C + (K + 1)K_2 \beta}{L} ; \quad a_{13} = \frac{K_3}{L} ; \quad a_{14} = \frac{(K + 1)K_3 \beta}{L} ; \\
a_{21} &= \frac{V_o(K_1 - K_2)}{CR_1 V_i} ; \quad a_{22} = \frac{V_o[K_1 C + (K + 1)K_2 \beta]}{CR_1 V_i} + \frac{1}{R_0 C} ; \quad a_{23} = \frac{V_o K_3}{CR_1 V_i} ; \\
a_{24} &= \frac{V_o (K + 1)K_3 \beta}{CR_1 V_i} ; \\
a_{31} &= 1 ; \quad a_{32} = 0 ; \quad a_{33} = 0 ; \quad a_{34} = 0 ; \\
a_{41} &= 0 ; \quad a_{42} = 1 ; \quad a_{43} = 0 ; \quad a_{44} = 0 ;
\end{cases}
\end{align*}

\text{(12)}

The characteristic equation derived from (11) and (12) is:

\begin{align*}
s^4 + bs^3 + cs^2 + ds + e &= 0
\end{align*}

\text{(13)}

where

\begin{align*}
\begin{cases}
b &= -a_{11} - a_{22} \\
c &= a_{11} a_{22} - a_{12} a_{21} - a_{13} - a_{24} \\
d &= a_{11} a_{24} + a_{13} a_{22} - a_{12} a_{23} - a_{14} a_{21} \\
e &= a_{13} a_{24} - a_{14} a_{23}
\end{cases}
\end{align*}

\text{(14)}

According the Routh-Hurwitz law, the judgment condition of stability could be written as follows:

\begin{align*}
s^4 &\quad 1 &\quad c &\quad e \\
s^3 &\quad b &\quad d \\
s^2 &\quad \frac{bc - d}{b} &\quad e \\
s^1 &\quad \frac{bcd - b^2 e - d^2}{bc - d} &\quad 0 \\
s^0 &\quad e
\end{align*}

\text{(15)}

The inequalities \( b > 0, \ bc - d > 0, \ bcd - b^2 e - d^2 > 0, \) and \( e > 0 \) must be satisfied to guarantee the stability of the sliding system.

The duty cycle \( D \) could be determined by \( u_{eq} \) plus a feedforward \( 1 - v_i/v_o \); the result is as follows:
\[ D = u_{eq} + (1 - v_i/v_o) = 1 + \frac{1}{v_o} \left( K_1 i_C + K_2 (V_{ref} - \beta v_o) + K_3 \int (V_{ref} - \beta v_o) dt \right) \\
+ K_2 [K(V_{ref} - \beta v_o) - i_L] + K_3 \int [K(V_{ref} - \beta v_o) - i_L] dt \]  

(16)

According to equation (16), the program of calculation of switch-on time could be written. \( v_{\text{ref}}, i, iL, v_i, v_o, K, K_1, K_2, K_3 \) denote \( V_{\text{ref}}, i_C, i_L, v_i, v_o, K, K_1, K_2, K_3 \). In equation (16), \( \beta \) could be set to 1, and the period is half of the switch period value of DSP. The program is as follows:

```c
v_err=v_ref-v_o;
I_err=K*v_err-il;
sigma_v_err+=v_err;
sigma_i_err+=i_err;
duty_delta=(K1*iC+K2*(v_err+i_err)+K3*(sigma_v_err+sigma_i_err))/v_o;
      duty=period+period*duty_delta;
```

3 Experiment and results

The prototype of totem pole PFC is shown in Fig. 2.

The waveforms of input voltage and input current at different input conditions are shown in Fig. 3; it can be concluded that the higher the input voltage and output power, the better the input current. When the input voltage is 230 V and the power is 2 kW, the RMS and peak-to-peak values of the current are 8.77 A and 29.2 A; when the input voltage is 115 V and the power is 1 kW, the RMS and peak-to-peak values of the current are 9.04 A and 32.4 A.
The test results for efficiency and power loss are shown in Fig. 4. As the switch frequency increases, the efficiency decreases, and power loss increases accordingly. At different switch frequencies, the peak efficiencies are 99.07% (50 kHz), 98.79% (100 kHz) and 98.48% (150 kHz).

As shown in Fig. 5, when the load is more than half, the THD is below 4% regardless of whether the input voltage is 115 V or 230 V, and the minimum values are 2.8% (115 V) and 2.7% (230 V), respectively. The test result of the current THD is good and meets the standard of IEC61000-3-12.
Conducted emissions have also been measured for this prototype using a LIN-115A LISN by Com-Power. The result, compared to EN55022A and EN55022B, is shown in Fig. 6. Test voltage and power are 230 V and 1200 W, respectively, and L and N are the output and input terminals of the EMI tester.

4 Conclusions
A 2.4 kW digital non-interleaved totem pole PFC with GaN HEMT is illustrated in this paper. Due to proposed advantages of the sliding mode control and GaN HEMT, the high-power prototype achieves better performances, such as efficiency, power loss, THD and EMI, than the traditional interleaved PFC based on the linear control and MOSFET. This means not only better performances but also lower cost and smaller volume. The cost of the totem pole PFC with GaN HEMT is not higher than the dual bridgeless PFC with MOSFET, owing to the fewer components and the lower filtering cost. It should be noted that GaN HEMT performs better than MOSFET only when operating in CCM, the advantages are not obvious when operating in either CRM or DCM.

Acknowledgments
This work was supported by the Science and Technology Projects of Guizhou Province of China (Grant No. 2014LH7050).