Electronic Nonlinearity of Full-Bridge PWM Inverter for Zero-Power PEMS System

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ABSTRACT The permanent-electro magnetic suspension (PEMS) technology takes advantage of the attractive magnetic force between the magnet and the iron core and reduces the power consumption eventually to zero. However, the active current in the electromagnet of the zero-power PEMS system fluctuates around zero due to external disturbances and suffers from the electronic nonlinearity of the driving circuit, such as the full-bridge pulse-width-modulation (PWM) inverter. This work presents that the 2 µs turn-off delay (one electronic defect) of the integrated circuit L298N (one commercial full-bridge PWM inverter produced by STMicroelectronics) leads to the nonlinear current-duty cycle characteristic, which undermines the control stability and limits the PWM frequency of the zero-power PEMS system. Moreover, the nonlinear mechanism is experimentally and theoretically analyzed for the critical PWM frequency and the sensitivity transition with respect to the 2 µs turn-off delay. Besides, the critical PWM frequency is of great significance for the energy efficiency and the dynamic performance of the high-speed PEMS transportation system.

INDEX TERMS Electromagnetics, DC-DC power converters, nonlinear network analysis, zero-power maglev.

I. INTRODUCTION

The electromagnetic suspension (EMS) technology has played an essential role in the maglev transportation industry in Germany, Japan, and China since 1960s [1]. Besides, the active magnetic bearing [2]–[4] is another successful application of the EMS technology and is widely found in flywheel systems, high-speed drives, and turbomolecular pumps. However, the EMS technology consumes considerably large power to generate sufficient attractive force against the gravity [5] and requires high-performance electronics, such as the isolated gate bipolar transistor (IGBT) [6]–[8] and the LLC resonant converter [5], [9].

In 1980s, the Nd-Fe-B permanent magnet was invented as a special functional material with huge magnetic energy accumulation. Since then, the permanent-electro magnetic suspension (PEMS) technology [5], [7], [10], [11] is economically effective by using the Nd-Fe-B permanent magnet to produce the attractive force and reduce the power consumption eventually to zero [10], [12], whereas the electromagnet only plays a dynamic regulatory role [6]. Mizuno et al. [13] developed an active vibration isolation system with an infinite stiffness against disturbances based on the zero-power PEMS technology. Zhang et al. [14] numerically optimized the geometry for the zero-power PEMS system as undergraduate project kits in terms of the zero-power force and the controller-gain requirement. Yalcin and Erkan [10] proposed linear-matrix-inequality (LMI) structures based on $H_\infty$ and $H_2$ control approaches for 3-DoF micro vibration isolation with zero-power problem.

Due to its outstanding energy-saving and robust features, the zero-power PEMS technology also attracted great attention from the maglev transportation industry. Tzeng and Wang [15] presented a rigorous dynamic analysis for the high-speed maglev transportation system with the robust zero-power PEMS strategy. Zhang et al. [6] focused on the optimal structural design of the PEMS magnet and proposed optimized parameters with better carrying capability.
and lower suspension power loss. Cho et al. [16] reported a successful quadruple PEMS system with the improved zero-power control algorithm.

Importantly, unlike the non-zero-mean current \( (I) \) in the EMS technology [17], [18], the current for the zero-power PEMS technology fluctuates about zero due to external disturbances [8], [16]. Though the full-bridge pulse-width-modulation (PWM) inverter [2] (also known as the power converter [19] or the chopper [5]) can drive the zero-mean current in the electromagnet by the PWM duty cycle \( (D) \), there are few works analyzing its electronic nonlinearity that could be of great significance for the zero-power PEMS technology.

Moreover, the PWM frequency \( (f_{PWM}) \) is crucial for the dynamic performance of the high-speed PEMS transportation system, e.g., the disturbance rejection and the gap tracking [20]. In the literature, \( f_{PWM} \) is usually set from 10 kHz to 20 kHz [5], [8], [11], [17], [18]. Generally, in order to enhance the dynamic performance of the high-speed PEMS transportation system, higher \( f_{PWM} \) is under huge demand but may amplify the energy loss as well as the electronic nonlinearity of the full-bridge PWM inverter, which is to be discussed in the present work.

This work aims to analyze the electronic nonlinearity of the full-bridge PWM inverter for the zero-power PEMS system. The L298N full-bridge PWM inverter is chosen as the research target due to its wide application in low-power electronics, whereas the IGBT [6], [8] is mainly used for high-power usage. The paper starts with the mathematical model and the nonlinearity of the \( I - D \) characteristic of the L298N full-bridge PWM inverter. Then, the nonlinear mechanism of the \( I - D \) characteristic is further analyzed experimentally and theoretically, especially for the critical PWM frequency \( (f_{cr}) \) and the sensitivity transition with respect to (w.r.t.) the 2 \( \mu \)s turn-off delay. Nevertheless, concluding remarks are addressed.

## II. CURRENT-DUTY CYCLE CHARACTERISTIC

### A. MATHEMATICAL MODEL

Fig. 1 shows the electric circuit of the L298N full-bridge PWM inverter produced by STMicroelectronics. The inverter converts the PWM signal into the current in the electromagnet that is modelled as a serial combination of an inductance \( (L) \) and a resistance \( (R) \). \( In1 \) and \( In2 \) are two TTL logic-level input ports and control the voltages at Nodes M and N, denoted as \( U_M \) and \( U_N \), respectively. For example, \( U_M \approx +U \) when \( In1 = \text{HIGH} \), and \( U_M \approx 0 \) V when \( In1 = \text{LOW} \). Denote the current in the electromagnet as \( i_{MN} \) and the voltage across as \( u_{MN} = U_M - U_N \). Hence, the relationship between \( i_{MN} \) and \( u_{MN} \) can be expressed as,

\[
u_{MN} = R i_{MN} + L d i_{MN} / dt, \quad (1)
\]

where \( u_{MN} \) is driven by the full-bridge PWM inverter and \( i_{MN} \) varies continuously due to the inductance. Relating physical properties are summarized in Table 1.

### TABLE 1. Physical properties.

| Symbol | Quantity | Value | Unit |
|--------|----------|-------|------|
| \( R \) | Electromagnet resistance | 14.5 | \( \Omega \) |
| \( L \) | Electromagnet inductance | 23.5 | \( \text{mH} \) |
| \( \tau \) | Electromagnet time constant | 1.62 | \( \text{ms} \) |
| \( U \) | Power supply voltage | 12 | \( \text{V} \) |
| \( f_{PWM} \) | PWM frequency | 0.1-0.10 | \( \text{kHz} \) |
| \( T \) | PWM period | 0.01-0.10 | \( \text{ms} \) |

Moreover, Table 2 lists out three modes of the full-bridge PWM inverter with the electromagnet. When \( In1 \neq In2 \), the electromagnet is either positively charged \((+C)\) or negatively charged \((-C)\) depending on the sign of \( u_{MN} \): when \( In1 = In2 \), the electromagnet is effectively short-circuited with \( u_{MN} \approx 0 \) V and its magnetic energy is dissipated during the discharging (DC) mode.

### TABLE 2. Three modes of full-bridge PWM inverter with electromagnet.

| Symbol | \( In1 = \text{HIGH} \) | \( In1 = \text{LOW} \) |
|--------|-----------------|-----------------|
| \( In2 = \text{HIGH} \) | \( \text{DC}: u_{MN} = 0 \) V. | \( -C: u_{MN} = -U \). |
| \( In2 = \text{LOW} \) | \( +C: u_{MN} = +U \). | \( \text{DC}: u_{MN} = 0 \) V. |

Meanwhile, in Table 3, four quadrants [2] are defined according to \( u_{MN} \) and \( i_{MN} \). Generally, the energy flows from the power supply into the electromagnet \((|u_{MN}| < U)\) during the forwards charging (FC) mode and flows from the electromagnet into the power supply \((|u_{MN}| > U)\) during the backwards charging (BC) mode.

### TABLE 3. Four quadrants of full-bridge PWM inverter with electromagnet.

| \( u_{MN} \) | \( i_{MN} \) | \( u_{MN} \) | \( i_{MN} \) |
|-------------|-------------|-------------|-------------|
| \( u_{MN} = -U \) | \( i_{MN} > 0 \) | \( u_{MN} = +U \) | \( i_{MN} < 0 \) |

In the present work, \( In1 \) is the PWM signal and \( In2 \) is obtained by the NOT gate as,

\[
In1 = \text{PWM} \quad \text{and} \quad In2 = \overline{\text{PWM}}, \quad (2)
\]

where the PWM signal can be explicitly expressed as,

\[
PWM = \begin{cases} 
\text{HIGH}, & nT < t < (n + D)T \\
\text{LOW}, & (n + D)T < t < (n + 1)T 
\end{cases}, \quad (3)
\]

FIGURE 1. Electric circuit of L298N full-bridge PWM inverter.

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where \( T (= 1/f_{PWM}) \) denotes the PWM period, \( D \in [0, 1] \) denotes the PWM duty cycle, and \( n \) is any arbitrary integer.

### B. IDEAL SCENARIO

Assuming the ideal full-bridge PWM inverter, \( u_{MN} \) can be expressed by referring to Table 2, (2) and (3) as,

\[
u_{MN} = \begin{cases} 
+U, & nT < t < (n + D)T \\
-U, & (n + D)T < t < (n + 1)T 
\end{cases}
\]

Moreover, denote the equilibrium current at \( t = nT \) as \( i_1 \) and that at \( t = (n + D)T \) as \( i_2 \). Solving (1) together with (4) gives the two equilibrium currents as,

\[
\begin{align*}
I_1 &= \frac{U}{R} \left( e^{-T/\tau} + 1 - 2e^{-((1-D)T)/\tau} \right) / \left( e^{-T/\tau} - 1 \right) \\
I_2 &= \frac{U}{R} \left( 2e^{-DT/\tau} - e^{-T/\tau} - 1 \right) / \left( e^{-T/\tau} - 1 \right)
\end{align*}
\]

where \( \tau = L/R \) denotes the time constant of the electromagnet. Hence, (5) indicates that \( i_{MN} \) fluctuates between \( I_1 \) and \( I_2 \).

Assuming \( \tau \) to be much larger than \( T \), i.e., \( T/\tau \ll 1 \), (5) can be approximated as,

\[
i_{MN} \approx i_1 \approx i_2 \approx \frac{U}{R} (2D - 1) \in [-U/R, U/R] ,
\]

which indicates that ideal \( i_{MN} \) in the electromagnet is linear with the duty cycle for a high \( f_{PWM} \).

### C. NONLINEARITY

Fig. 2 shows the experiment rig for the \( i_{MN} - D \) characteristic of the L298N full-bridge PWM inverter. \( i_{MN} \) is measured by the ammeter, \( u_{MN} \) is measured by the oscilloscope, whereas the PWM signal is generated by the pulse generator.

**FIGURE 2.** Experiment rig for the \( i_{MN} - D \) characteristic of L298N full-bridge PWM inverter.

Fig. 3(a) visualizes the variations of \( i_{MN} \) w.r.t. \( D \) for \( f_{PWM} = 0.1, 1, 10, 100 \text{kHz} \). It is observed that the characteristic \( i_{MN} - D \) curve, e.g., \( i_{MN} = f_1 (D) \), is an odd function about Point (0.5, 0) as,

\[
f_1 (1 - D) = -f_1 (D) .
\]

Also, the characteristic \( i_{MN} - D \) curve deviates from the ideal scenario in (6), and the deviation increases with the increase of \( f_{PWM} \).

Moreover, define the sensitivity of \( i_{MN} \) w.r.t. \( D \) as,

\[
S = di_{MN} / dD.
\]

Fig. 3(b) visualizes the variations of \( S \) w.r.t. \( D \). It is observed that \( S \) at \( D = 0.5 \) decreases with the increase of \( f_{PWM} \) (as low as 70 mA for 100 kHz), whereas the four sensitivity curves converge to around 1700 mA for \( D < 0.3 \) and \( D > 0.7 \). In other words, with the increase of \( f_{PWM} \), the accumulation of the magnetic energy around \( D = 0.5 \) is weakened.

Furthermore, since \( F \) is linear with \( i_{MN} \), the large sensitivity change about the zero current (e.g., around 170/70 \( \approx 24.3 \) times for 100 kHz) will significantly undermine the control stability of the zero-power PEMS system.

**FIGURE 3.** Variations of (a) \( i_{MN} \) and (b) \( S \) w.r.t. \( D \) for \( f_{PWM} = 0.1, 1, 10, 100 \text{kHz} \).

### III. NONLINEAR ANALYSIS

This section analyzes the nonlinear mechanism of the \( i_{MN} - D \) characteristic for the electromagnet driven by the L298N full-bridge PWM inverter. Firstly, the voltage dynamics is investigated for two \( f_{PWM} \) (one ultra-high and one ultra-low) and \( D = 0.5 \). Secondly, DC, FC and BC modes are respectively characterized. Thirdly, the critical PWM frequency (\( f_{cr} \)) is modelled from the energy perspective. Fourthly, the sensitivity transition is analyzed w.r.t. \( f_{cr} \).

### A. VOLTAGE DYNAMICS

1) \( f_{PWM} = 100 \text{kHz (ULTRA-HIGH)} \)

Fig. 4 visualizes the whole-period variations of \( I_1, I_2, U_M, U_N \), and \( u_{MN} \) for \( f_{PWM} = 100 \text{kHz} \) and \( D = 0.5 \) w.r.t. the time \( t \). It is observed that the dynamic responses of \( I_1 \) and \( I_2 \) are so fast that the step change is completed in less than 0.1 \( \mu s \), the rising and falling step changes of \( U_M \) and \( U_N \) are completed in 0.2 \( \mu s \) and 1 \( \mu s \), respectively. However, by
comparing Figs. 4(a-b), there is a significant delay for around 2 µs between the falling step changes of \( U_M \) (or \( U_N \)) and \( \ln 1 \) (or \( \ln 2 \)), which is referred to as the turn-off delay in the datasheet of the L298N full-bridge PWM inverter produced by STMicroelectronics. Therefore, the 2 µs turn-off delay results in the DC mode as defined in Table 2 and highlighted in Fig. 4(c).

Technically, the energy-dissipation capability of the DC mode is finite in the 2 µs duration. Therefore, the immediate occurrence of the FC mode indicates that the magnetic energy is entirely dissipated by the DC mode for \( f_{PWM} = 20, 50, \) and 100 kHz, whereas the magnetic energy exceeds the energy-dissipation capability of the DC mode for \( f_{PWM} = 5 \) and 10 kHz.

2) CHARGING CHARACTERISTIC

Fig. 7 visualizes the step-change responses of \( u_{MN} \) in the initial 20 ms for \( f_{PWM} = 2, 10, 20, 50, \) and 200 Hz and \( D = 0.5 \). The six curves undergo the BC mode before the FC mode. Moreover, it is observed that the curves for \( f_{PWM} = 2, 10, \) and 20 Hz have a long BC mode for around 5 ms and a long FC mode for around 9 ms, and fluctuate around +10 V afterwards. On the contrary, the curves for \( f_{PWM} = 50, 100, \) and 200 Hz have shorter BC mode and falls out of the scope before reaching +10 V.

Technically, the charging characteristic is determined by the power supply (e.g., supply voltage and inner resistance) and the electromagnet (e.g., inductance and resistance). Therefore, the saturation at \( u_{MN} = \pm 10V \) indicates that the electromagnet is entirely charged by the FC mode for \( f_{PWM} = 2, 10, \) and 20 Hz but is partially charged due to the insufficient duration for \( f_{PWM} = 50, 100, \) and 200 Hz. Besides, \( u_{MN} = \pm 9.5V \) can be regarded as the boundary between the DC mode and the two charging modes for the present experimental setup.

C. CRITICAL PWM FREQUENCY

1) EXPERIMENTAL OBSERVATION

Fig. 8 visualizes the contour of \( u_{MN} \) w.r.t. \( \log_{10}(f_{PWM}) \) and the normalized time \( (t/T) \) for \( D = 0.5 \). It is observed that the BC regime expends for \( f_{PWM} \leq 100 \) Hz, shrinks for \( f_{PWM} > 100 \) Hz, and vanishes at \( f_{PWM} \approx 20 \) kHz, whereas the DC regime expends exponentially. Besides, \( u_{MN} \) reaches the saturation at ±10 V only for \( f_{PWM} < 10^{1.5} \text{Hz} \approx 31.6 \text{Hz} \) and is increasingly undercharged for \( f_{PWM} > 31.6 \text{Hz} \).
TABLE 4. Relative portions of three modes for various \( f_{PWM} \) and \( D = 0.5 \).

| \( f_{PWM} \) (kHz) | 1  | 2  | 5  | 10 | 20 | 50 | 100 |
|-------------------|----|----|----|----|----|----|-----|
| \( T_{FC}/T \) (%) | 11.23 | 9.33 | 8.08 | 6.85 | 0.00 | 0.00 | 0.00 |
| \( T_{BC}/T \) (%) | 0.40 | 0.80 | 2.00 | 4.00 | 8.00 | 20.00 | 40.00 |
| \( T_{DC}/T \) (%) | 88.37 | 89.87 | 89.92 | 89.15 | 92.00 | 80.00 | 60.00 |

2) THEORETICAL ANALYSIS

For \( D = 0.5 \), the magnetic energy \( (E) \) in the electromagnet is charged by the FC mode and dissipated by the DC and BC modes. And, the energy balance equation can be expressed as,

\[
E_{FC} = E_{DC} + E_{BC}, \tag{9}
\]

where \( E_{FC}, E_{DC}, \) and \( E_{BC} \) denote the charged energy by the FC mode and the dissipated energies by the DC and BC modes, respectively.

Moreover, assuming \( T \ll \tau, E_{FC}, E_{DC}, \) and \( E_{BC} \) can be linearized and approximated as,

\[
E_{FC} = e_{FC}T_{FC}, E_{DC} = e_{DC}T_{DC}, E_{BC} = e_{BC}T_{BC}, \tag{10}
\]

where \( e_{FC}, e_{DC}, \) and \( e_{BC} \) denote the energy-charging rate by the FC mode, and the energy-dissipating rates by the DC and BC modes, respectively.

Assuming \( T_{DC} > 0 \), substituting (10) into (9) gives,

\[
\frac{T_{FC}}{T_{DC}} = \left( \frac{e_{BC}}{e_{FC}} \right) \left( \frac{T_{BC}}{T_{DC}} \right) + \frac{e_{DC}}{e_{FC}}, \tag{11}
\]

which is in the first-order form, i.e., \( y = ax + b \), where \( y = \frac{T_{FC}}{T_{DC}}, x = \frac{T_{BC}}{T_{DC}}, a = \frac{e_{BC}}{e_{FC}}, \) and \( b = \frac{e_{DC}}{e_{FC}}. \)

By substituting \((x, y)\) associated with \( f_{PWM} = 1, 2, 5, \) and 10 kHz in Table 4 into (11), \((a, b)\) can be solved by the least square method as,

\[
a = \frac{e_{BC}}{e_{FC}} = 7.46 \quad \text{and} \quad b = \frac{e_{DC}}{e_{FC}} = 15.33, \tag{12}
\]

which indicates that the energy-dissipating rate by the DC mode is around twice as much as that by the BC mode. Therefore, the DC mode gradually replaces the BC mode with the increase of \( f_{PWM} \), which leads to the lower sensitivity as well as the larger energy loss for \( D = 0.5 \) as shown in Fig. 3(b).

Furthermore, \( f_{cr} \) is defined as the vanishing point of the BC regime \( (T_{BC} = 0) \) for \( D = 0.5 \) and can be expressed from (11) as,

\[
f_{cr} = \frac{1}{\frac{T_{FC}}{T_{DC}} + \frac{1}{1 + \frac{e_{DC}}{e_{FC}}}} \frac{1}{T_{DC}}, \tag{13}
\]

which gives \( f_{cr} = 15.28 \) kHz.

Table 5 lists out the estimated relative portions of the three modes by (12) and the associated errors for \( f_{PWM} = 1, 2, 5, \) and 10 kHz and \( D = 0.5 \). For example, the error \((\pm 0.18\%)\) for \( f_{PWM} = 1 \) kHz is the difference between \( T_{BC}/T = (11.23\%) \) and estimated \( T_{BC}/T = (11.05\%). \)

Therefore, the first-order estimation for the magnetic energy is generally acceptable with low errors.

TABLE 5. Estimated relative portions of three modes and associated errors for various \( f_{PWM} \) and \( D = 0.5 \).

| \( f_{PWM} \) (kHz) | 1  | 2  | 5  | 10 | \( f_{cr} = 15.28 \) | 50 |
|-------------------|----|----|----|----|---------------------|----|
| Estimated \( T_{BC}/T \) (%) | 11.05 | 10.28 | 7.96 | 4.10 | 0.00 | 0.00 |
| Estimated \( T_{BC}/T \) (%) | 0.40 | 0.80 | 2.00 | 4.00 | 6.13 | 20.00 |
| Estimated \( T_{BC}/T \) (%) | 88.55 | 89.92 | 90.04 | 91.90 | 93.87 | 80.00 |
| Error (%) | 0.18 | -0.95 | 0.12 | 2.75 | - | - |
D. SENSITIVITY TRANSITION

1) EXPERIMENTAL OBSERVATION

Figs. 9(a-c) visualize the phase diagrams of $u_{MN}$ w.r.t. $D$ and $t/T$ for $f_{PWM} = 1, 10,$ and $100$ kHz. It is observed that the boundaries between the FC and BC regimes are almost continuous for $f_{PWM} = 1$ kHz as a result of the negligible DC regime, whereas the BC regime is heavily squeezed by the DC regime for $f_{PWM} = 100$ kHz. Besides, the cyan dotted lines indicate the termination points of the FC regimes.

Moreover, Figs. 9(d-f) visualize the associated variations of $S$. The purple dotted lines indicate the transition points with $S = 1000$ mA. It is observed that the three sensitivity curves converge to the high sensitivity outside the purple dotted lines; the three curves stay near the respective minimum values between the cyan dotted lines; meanwhile, the sensitivity transitions take place between the purple and cyan dotted lines.

2) THEORETICAL ANALYSIS

Based on Fig. 9, Table 6 and Table 7 outline the evolution of the three modes w.r.t. $D \geq 0.5$ for $f_{PWM} < f_{cr}$ and $f_{PWM} > f_{cr}$, respectively. The red background indicates $i_{MN} > 0$, the blue background indicates $i_{MN} < 0$, and the green background indicates $i_{MN} \approx 0$.

By referring to Table 6 and Table 7, the variations of $S$ can be characterized into three zones from the energy perspective.
• **Low-sensitivity zone** (*Stages I, II, and III, between the cyan dotted lines*): Both +FC and −FC modes co-exist, which indicates that the magnetic energy stored during the DC mode (both positive and negative) is completely dissipated by the DC or BC mode. Hence, the low-sensitivity zone is dominated by the energy cancellation between the FC and BC modes for \( f_{PWM} < f_{cr} \) and between the FC and DC modes for \( f_{PWM} > f_{cr} \). Besides, since the energy-dissipating rate by the DC mode is larger than that by the BC mode in (12), the sensitivity for \( D = 0.5 \) decreases with the increase of \( f_{PWM} \).

• **Sensitivity-transition zone** (*Stage IV, between the purple and cyan dotted lines*): The 2 \( \mu s \) DC mode has a finite energy-dissipation capability and buffers the sensitivity transition until the capability is entirely consumed, as outlined by the purple and cyan dotted lines. Therefore, the sensitivity transition expends with the increase of \( f_{PWM} \).

• **High-sensitivity zone** (*Stage V, outside the purple dotted lines*): The expending of the FC regime and the shrinking of the BC regime occur simultaneously with the increase of the duty cycle, whereas the current in the electromagnet does not change the direction. Therefore, the magnetic energy accumulates more effectively and the corresponding sensitivity is larger in the high-sensitivity zone than those in the low-sensitivity zone.

### IV. CONCLUSION

The present work investigates into the electronic nonlinearity of the L298N full-bridge PWM inverter for the zero-power PEMS system. Based on the experimental observation and the theoretical analysis, we can draw the following conclusions:

- In micro level, it is observed that the L298N full-bridge PWM inverter possesses a 2 \( \mu s \) turn-off delay that leads to the passive DC mode; in macro level, such a tiny delay results in the nonlinear \( i_{MN} \sim D \) characteristic.
- With the increase of \( f_{PWM} \), the relative portion of the 2 \( \mu s \) DC mode is significantly amplified and the DC mode replaces the BC mode entirely at \( f_{cr} = 15.28 \text{ kHz} \) and \( D = 0.5 \), which is theoretically modelled from the energy perspective.
- Since the energy-dissipating rate by the DC mode is around twice as much as that by the BC mode, the sensitivity as well as the energy efficiency at \( D = 0.5 \) decrease with the increase of \( f_{PWM} \).
- The sensitivity transition results from the finite energy-dissipation capability of the 2 \( \mu s \) DC mode and expends with the increase of \( f_{PWM} \).
- A full-bridge PWM inverter with a lower turn-off delay will result in a higher \( f_{cr} \) and be more preferable, such as CM75TF-12H (IGBT produced by MITSUBISHI, with a 0.2 \( \mu s \) turn-off delay)[8].
- Though higher \( f_{PWM} \) can enhance the dynamic performance of the high-speed PEMS transportation system [20], a slightly smaller \( f_{PWM} \) than \( f_{cr} \) is more preferable due to the higher energy efficiency and the gentler sensitivity transition, whereas the suitable compensation measures are necessary to take care of the nonlinear \( i_{MN} \sim D \) characteristic, e.g., the bias current [3], [17] and the close-loop current control [2], [4], [7], [8], [10], [11].

### CONFLICTS OF INTEREST

The authors declare no conflict of interest.

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