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Full Text

Preamble

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The Application of Permanent Magnet Synchronous Motor with Small Electrical Time Constant in Fiber Positioner

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Abstract

With the development of cutting-edge multi-object spectrographs, fiber positioners located in the focal plane are being scaled down in size, and miniature hollow-cup Permanent Magnet motors are now being considered as a suitable replacement for Faulhaber Precistep stepper motors. However, the small electrical time constant of such coreless motors poses a challenge, as the problem of severe commutation torque ripple in a fiber positioner running a position loop has been tricky. To overcome this challenge, it is advised to increase the Pulse Width Modulation (PWM) frequency as much as possible to mitigate the effects of the current fluctuation. This must be done while ensuring adequate resolution of the PWM generator. By employing a voltage open-loop field-oriented control based on a modulation frequency of 1 MHz, the drive current only costs 25 mA under a 3.3 V power supply. The sine degree of phase current is immaculate, and the repeat positioning accuracy can reach 2 μ m. Moreover, it is possible to further shrink the bill of devices and the layout area of the Printed Circuit Board, especially in size-sensitive applications. This device has been developed under the new generation of The Large Sky Area Multi-Object Fiber Spectroscopic Telescope.

Key words: instrumentation: adaptive optics -instrumentation: spectrographs -proper motions -reference systems

1. Introduction

The powerful Multi-Object Spectroscopy survey is a fascinating area of study for astronomers worldwide. It offers the unique capability of simultaneously observing and studying the formation and evolution of galaxies [?], the search for dark matter and dark energy [?], and the characterization of exoplanets. This impressive feat is the result of decades of tireless efforts by researchers who have focused on large-scale spectral measurements. The establishment of many MOS (Multi-Object Spectroscopy) telescope projects has been crucial to this field. Among these projects are MOONS (The Multi-Object Optical and Near-

infrared Spectrograph) [?], LAMOST (The Large Sky Area Multi-Object Fiber Spectroscopic Telescope) [?], and DESI (The Dark Energy Spectroscopic Instrument) [?]. These projects have significantly improved the parallel observation capability of MOS and achieved large-scale spectral observations.

To further enhance the parallel observation capability of MOS, a large number of SCARA (Selective Compliance Assembly Robot Arm) robotic arms have been arranged on the focal plane of the above telescope system. These robotic arms are used to quickly and accurately place the spectral receiving fibers to collect faint spectra from stars and galaxies [?]. The parallel working SCARA robotic arms have become the most critical component in the fiber positioning system, and they are called fiber positioners. Figure 1 [Figure 1: see original paper] explains the schematic of the double rotary structure of the classic fiber positioner, also known as the RR planar kinematics [?].

In the pursuit of more efficient and parallel observation of celestial targets, modern multi-spectral sky surveys require a greater quantity and smaller size of fiber positioners. To this end, the volume of motors used to drive rotating shafts is developing toward miniaturization, according to research [?, ?]. Despite this trend, Faulhaber Precistep stepper motors remain one of the preferred solutions for many fiber positioner designs, such as those in MEGARA [?], LAMOST [?], and MOONS [?]. However, the smallest stepper motor currently available is only f6 mm (ref Faulhaber), and stepper motors also possess a passive holding torque. This means that when the power is turned off, the stepper motor must stop at a certain full step, requiring a high reduction ratio.

Nowadays, with the low efficiency of the stepper motor working mode and its larger physical size, smaller-sized miniature brushless DC motors have emerged as a more suitable option for future downscaling of the design. Brushless DC motors, unlike stepper motors, do not have the full-step stop problem. However, it should be noted that some brushless DC motors do possess cogging torque. Araujo et al. [?] have observed this phenomenon and noted that when powering off the motor coils, the axis might slightly move during their tests. The solution to this issue is to apply a small current into the motor coil to keep the position, causing additional thermal dissipation. In order to address the issue of cogging torque, Hörler et al. [?] chose a hollow-cup Permanent Magnet (PM) motor without cogging torque for the DESI project. Their motor drive module can be directly turned off to enter low-power mode during the focal plane exposure, which is essential to develop astronomical instruments. This is because unnecessary power consumption can impact the signal-to-noise ratio if important temperature gradients are created, producing turbulent air flows near the focal plate [?].

The DESI has successfully deployed and implemented fiber positioners driven by f4 mm hollow-cup brushless DC motors [?]. This innovation has allowed for greater efficiency and smaller size compared to the original stepper motors that were used. The new generation of LAMOST is also considering utilizing more efficient and smaller-sized brushless DC motors to replace the original stepper

motors. However, the adoption of miniature hollow-cup motors also presents new challenges in terms of positioning control technology. Accurately placing the fibers at the target point is crucial for improving the spectral reception capacity of the telescope system. Therefore, the development of advanced positioning control technology for miniature hollow-cup motors is an urgent issue that needs to be addressed.

The hollow-cup PM motor is a novel type of permanent magnet brushless DC motor (BLDCM) that has recently become popular in the field of fiber positioner. The distinctive feature of this motor is that it lacks iron laminations in its armature, which completely eliminates the stator slots and magnetic saturation issues that are common in traditional iron-core motors [?, ?]. This reduction in stator slots and magnetic saturation leads to a significant decrease in cogging torque ripple and reluctance [?, ?], which makes it possible for the fiber positioner to cut off the power to the motor coils during focal plane exposure without any slight movement of the fiber. This measure is critical because it ensures good seeing of the telescope system [?] and prevents the creation of seeing-degrading thermal plumes in the telescope field of view [?]. Therefore, the hollow-cup PM motor is expected to become more widely adopted in fiber positioner in the near future.

As we embark on the path of understanding miniature-sized hollow-cup PM motors, we encounter a perplexing dilemma. The coreless windings with a low inductance pose a formidable obstacle in drive control, leading to severe torque ripple that can be difficult to tame. It is worth noting that the larger-sized hollow-cup PM motors seem to be better equipped to handle such a situation, as research indicates [?, ?]. However, the miniature-sized hollow-cup PM motors are constrained by their compact size, and their windings are fashioned from fine copper wire, which imbues them with higher impedance. This high impedance, in turn, causes the electrical time constant to shrink significantly, giving rise to a host of challenges in attaining a smooth spin of the motor. Therefore, in this article, we aim to explore this issue further and uncover potential solutions by working with a 4 mm hollow-cup PM (supplied by NAMIKI Precision Jewel Co., Japan).

We organize the paper as follows. In Section 2, we demonstrate the typical characteristics of miniature hollow-cup PM used in a fiber positioner, namely a very small electrical time constant, and analyze the impact of such characteristic on commutation current ripple from the perspective of drive control. The expression of current ripple in time domain is deduced in Section 3. In Section 4, we verify the idea with relevant hardware and figure out that upper limitation of the resolution of Pulse Width Modulation (PWM) generator would also play an important role in practical applications, i.e., power bridge's signal-level delay and timer's upper frequency. A high carrier frequency control method based on open-loop field-oriented control (FOC) is deployed in fiber positioner for validation in Section 5 and performance and test results of fiber positioner are discussed in Section 6. We conclude in Section 7.

2. Characteristics of Miniature-sized Hollow-cup PM

Generally, BLDCM has three kinds of torque ripple [?]:

1. The reluctance torque caused by the interaction of the stator magnetomotive forces with the angular variation of the rotor magnetic reluctance.
2. The cogging torque resulting from the interaction of the magnet and the stator tooth.
3. The current ripple due to PWM-controlled technology, as a consequence of periodic commutation and limitation of the resolution of PWM generator [?].

From Figure 2 [Figure 2: see original paper], it is clear that miniature-sized hollow-cup PM' s inner rotor is composed of a whole permanent magnet due to physical limitation, i.e., there is no cogging structure and corresponding torque ripple. The reluctance torque is presented in Equation (1) [?]:

$$t_r = \frac{3}{2}p(L_d - L_q)i_d i_q \sin 2\delta$$

Taking the rotor coordinates (d-q axes) of the motor as reference coordinates [?], L_d , L_q are respectively d- and q-axis inductance, δ represents angle between stator flux and rotor, and ψ_s is the flux generated by stator. Reluctance torque t_r is apparently correlated to L_d , L_q .

Equation (2) describes definition of inductance:

$$L = \frac{T_{ph}\Phi}{I} = \frac{T_{ph}F}{IR_m}$$

where L is single phase inductance, T_{ph} stator winding turns, I phase current, Φ magnetic flux, F magnetomotive force (MMF), R_m magnetic resistance of the magnetic circuit. Based on this, the inductance is inversely proportional to the resistance of the magnetic circuit:

$$R_m = \frac{l}{\mu_0\mu_r A}$$

where l is magnetic circuit length, μ_0 is permeability of vacuum, μ_r permeability of permanent magnet, A cross-sectional area of magnetic circuit. We obtain inductance ratio of d-axis to q-axis as follows:

$$\frac{L_d}{L_q} = \frac{l_d}{l_q}$$

According to Figure 2, we can conclude that l_d equals l_q without consideration of inhomogeneity of motor manufacture. Here, l_d and l_q represent the magnetic circuit length of d- and q-axis respectively. Basically, there is no concern of reluctance torque ripple because of $L_d = L_q$.

The concept of electrical time constant plays an essential role in understanding the performance of a f4 mm motor. The motor time constant of $2.49158 \mu s$ is calculated as $\tau = L/R$. This time constant is an indication of how fast the motor' s phase current can change in response to a change in voltage. In comparison, the mechanical time constant supplied by the motor datasheet is 9.15 ms, which is much larger than the electrical time constant. This implies that the adjustment of the motor' s phase current is much faster than the change of Back ElectroMotive Force (BEMF) generated by the rotor' s rotation. In other words, the former is only less than 1/3700 of the latter, namely, when phase current adjustment ends, BEMF almost maintains the same. Consequently, BEMF can be seen as constant during a switching period [?].

So the above circuit (Figure 3 [Figure 3: see original paper]) can be simplified as a classical LR series model. We choose the seven-segment Space Vector Pulse Width Modulation (SVPWM) method to generate a smoother sine wave in the windings [?]. The count and effective level distribution of the timer during switching period are shown in Figure 4 [Figure 4: see original paper]:

$$T = T_{on} + T_{off}$$

The phase current follows the first-order full-correspondence expression:

$$i(t) = i(\infty) + [i(0) - i(\infty)]e^{-t/\tau}$$

BEMF constant of motor K_e equals $0.059 \text{ mV rpm}^{-1}$, and motor rotates uniformly at a low speed in fiber positioner, so we ignore the fluctuation of BEMF here. Besides, SVPWM modulation results in that the maximum effective value of the phase voltage equals $V_{dc}/\sqrt{3}$, where V_{dc} remains constant. Then the current during a switching period T can be derived further into:

$$i(t) = \frac{V_{dc}/\sqrt{3}}{R}(1 - e^{-t/\tau})$$

During a switching period, the peak phase currents take place at t_1 and t_2 , then we get the maximum value of current ripple:

$$i_{ripple} = i(t_1) - i(t_2)$$

with $t_1 + t_2 = T$, consequently:

$$i_{ripple} = \frac{V_{dc}/\sqrt{3}}{R}(e^{-t_2/\tau} - e^{-t_1/\tau})$$

Determine i_{ripple} to express the severity of current fluctuation. Plot current waveforms in math analysis software, Figure 5 [Figure 5: see original paper] shows that i_{ripple} varies with PWM frequency. Here, i_{ripple} represents the maximum value during single PWM period. It can be clearly seen that:

1. As the PWM carrier frequency increases from 5 kHz to 15,000 kHz by degree, the current fluctuation tends to be smoother.
2. The maximum value of i_{ripple} is 200%, which is easy to understand: during single PWM period, $i(t)$ reaches $i(\infty)$ and $-i(\infty)$, while switching period is much longer than τ .
3. i_{ripple} varies periodically and i_{ripple} max always occurs at a fixed position during 2π electrical angle.

Based on above conclusion (3), we take the assumption that i_{ripple} max always occurs at the same duty ratio over an electrical angle cycle of 2π . More precisely, i_{ripple} max $\ll i(\infty)$, regardless of PWM frequency. In that case, Equation (7) can be rewritten as:

$$i_{ripple} \approx \frac{V_{dc}/\sqrt{3}}{R} \cdot \frac{T}{\tau} \cdot f(\beta)$$

where $f(\beta)$ is a function of duty ratio β ($-1 < \beta < 1$), and define e_{ripple} as normalized ripple:

$$e_{ripple} = \frac{i_{ripple}}{i(\infty)} = \frac{T}{\tau} \cdot f(\beta)$$

Apparently, when the PWM frequency increases gradually, switching period T shall decrease, $x = T/\tau$ decreases. Therefore, when the PWM frequency goes down, x increases, finally e_{ripple} increases. It is necessary to analyze the monotonicity of i_{ripple} max to find out detailed mathematical relationship between i_{ripple} max and x , which will make above conclusions (1) and (2) more convincing mathematically.

Consider $g(x) = \frac{x}{1-e^{-x}}$, the part of $g(x)$ monotonically increasing, i.e., $g'(x) > 0$ when $0 < x < 1$. When $x > 1$, $[a - 1 + (3a - 1)x^2]$ is apparently monotonically increasing, and $g(x)$ monotonically decreases. When a equals 0.5, $g'(x) < 0$ when $x > 1$, $g(x)$ varying likewise. When $0 < x < 1$, $[a - 1 + (3a - 1)x^2]$ is monotonically decreasing, whose value ranges from plus to minus, causing $g(x)$ to first increase and then decrease. In conclusion, $g'(x) < 0$ is always right when $x > 0.5$, so i_{ripple} max monotonically decreases along with x increasing. With PWM frequency increasing, i_{ripple} max is on the decrease. These are consistent with the above conclusions (1) and (2) from Figure 5.

4. The Experiments of Upscaling PWM Frequency and Adding Inductance

Relevant hardware platform is designed to validate conclusions obtained in the theoretical derivation of Section 3. Figure 6 [Figure 6: see original paper] reveals the details.

4.1. Upscaling PWM Frequency

In Figure 7 [Figure 7: see original paper], it is evident that the phase current ripple diminishes as the PWM carrier frequency increases up to 1 MHz. However, as the switching frequency surpasses 1.5 MHz, the degree of the phase current sine worsens, and the phase current nearly becomes six-step commutation with a PWM carrier frequency of 15 MHz. This is quite different from the ideal analysis results obtained in Section 2. The reason that leads to such a difference is that the MOSFET is not an absolutely perfect device, but rather has a rising and falling delay when generating signal-level toggle. Besides, MCU' s timer has an upper frequency limit. Taking the STM32G4 as an example, its advanced timer has a maximum frequency of 170 MHz. Hence, when the PWM frequency is 15 MHz, the count of one cycle is only 11, and various SVPWM voltage vectors switch to the same due to the output of rounding in the MCU, resulting in severe current distortion. Both of these key factors combine to determine the resolution of the PWM generator.

4.2. Seriesing Inductor

Based on Equation (8), it can be observed that by keeping the PWM frequency constant, if we add a series inductor in the motor circuit, it would increase the value of τ . Consequently, the value of x would also increase, leading to a decrease in $i_{ripple} \max$. To implement this, we purchase lots of inductor 4590R-684K (manufactured by Coilcraft Inc., The USA) with 10% inductance tolerance, and pick out three inductors whose resistance and inductance are nearly identical by Direct Current Inductance Capacitance Resistance (LCR) meter, then series them in motor circuit. This ensured that the time constant τ increased, and the electrical characteristics of the three-phase circuit of the motor remained consistent.

The inductors we selected are all near 672 μH and 520 $\text{m}\Omega$, which make τ increase from 2.49158 μs to 46.129 μs . The experiment conditions are consistent with Section 4.2, and results about the fluctuation of the phase current are obtained in Figure 7. Apparently, current ripple is significantly suppressed at each PWM frequency spot below 1 MHz, after which the resolution of PWM generator becomes the main limitation. More precisely, it is the upper limit frequency of MCU' s timer that leads to such a dilemma.

4.3. Explanation in Frequency Domain

The fundamental concept of PWM control is transforming direct current into alternating current, whereby the utilization of PWM as a carrier serves to engender an alternating current signal commonly referred to as modulation wave. Considering the circuit model of single phase of the three-phase motor is a typical RL series configuration, i.e., a classic low-pass filter as shown in Figure 8 [Figure 8: see original paper]. In order for the filter to operate in a desirable fashion, it is critical to ensure that the cut-off frequency is greater than the frequency of the modulating wave while still being less than the frequency of the carrier wave, thereby facilitating the efficient modulation of the alternating current signal.

As for modulation wave:

$$V_{out} = V_{in} \cdot \frac{1}{\sqrt{1 + (\omega/\omega_{3dB})^2}}$$

where V_{out} represents output of low-pass filter, and V_{in} input of that. When $2\pi f = \omega_{3dB} = 1/\tau$, $f = 63.877$ kHz, output power of the carrier wave is subject to an abrupt attenuation of 50% of the input power. When the carrier frequency goes on ascending, the carrier wave mixed in the low-frequency current signal exhibits a proclivity to gradually dissipate, thereby resulting in the effective elimination of current ripple. In essence, the salient aim of increasing the carrier frequency or adding inductance primarily rests on the endeavor to further diminish the power of the PWM carrier while concurrently abating the pernicious influence of the PWM carrier on the low-frequency modulation signal [?].

5. High PWM Frequency Control Method based on FOC

It has been observed that the inductor possessing a larger inductance value and a relatively minute direct current resistance tends to have a larger physical size, which renders the incorporation of a series inductor incompatible with certain compact fiber positioners or applications that require size sensitivity. We choose to increase the PWM carrier frequency to control motor. Motor in fiber positioner works in low-speed and making phase current smooth is critical. FOC [?], a classic sine wave control method, turns into first-preference. Since the absence of an encoder commensurate with the f4 mm motor, it is necessary to pursue a sensorless drive method, whereby the primary focus is on the sensorless closed-loop control, which operates based on the underlying principles of the following two cardinal tenets:

1. Detect the variations of the windings inductance which is caused by the change of the magnetic flux while rotor spins.
2. On the basis of BEMF detection.

Because the stator of the hollow-cup PM has no iron core, this engenders a state of affairs wherein the spin of the rotor does not effectuate any alteration to the magnetic resistance of the inductance magnetic loop. Thus the value of phase inductance remains stable. Additionally, six-step commutation based on BEMF is not applicable at low speed because of hollow-cup PM's small BEMF constant $0.059 \text{ mV rpm}^{-1}$, not to mention the fact that square wave commutation elicits a problematic vibration that can engender deleterious effects on the wearing and precision of the mechanic [?]. Therefore, this paper proposes an open-loop FOC method to realize the low-speed positioning control with motor that exhibits a small electrical time constant.

Equation (10) puts the d-q axis voltage equation [?]:

$$\begin{cases} V_d = R_s I_d - \omega_e L_q I_q \\ V_q = R_s I_q + \omega_e L_d I_d + \omega_e \psi_f \end{cases}$$

where V_d , V_q represent d- and q-axis stator voltage respectively, I_d , I_q are d- and q-axis armature current, and ω_e is rotor electric angular speed. Control expectation is that $I_d \rightarrow 0$. Then Equation (10) is simplified as:

$$\begin{cases} V_d = -\omega_e L_q I_q \\ V_q = R_s I_q + \omega_e \psi_f \end{cases}$$

I_d is desired to be zero, hence, V_d^* equals zero, and V_q^* is used to tune I_q . The open-loop control aims to make V_q^* to tune I_q , in that proper driving torque will be found out. Figure 9 [Figure 9: see original paper] illustrates the detailed logic of open-loop control.

According to the results of Section 4.2, we decided to adopt 1 MHz PWM switching frequency with V_{bus} equaling 3.3 V. The period for a voltage vector in the open-loop control is $2 \mu\text{s}$, while the calculation time of a voltage vector in MCU consumes $0.9 \mu\text{s}$, which is fair adequate. The 2π electrical angle is expressed in Q15 ($-32768 \sim 32767$, 1 bit represents the sign position, and 15 bits represent the variable in a 16-bit stored data), and count is incremented by 2 in timer, i.e., the electrical angle increment is $2\pi/32768$. The voltage vector undergoes a staggering 32,768 commutations over the course of a single rotation of the motor. The increased number of electronic commutations makes the motor phase current change more continuously and smoothly. Besides, the motor rotates at a constant speed of 1024 rps. The commutation time of a single voltage vector amounts to a mere $2 \mu\text{s}$. The number of commutations per revolution serves as a key determinant of both the rotation speed of the motor as well as the overall smoothness of the phase current. It stands to reason that a higher number of commutations per revolution would lead to a slower motor rotation speed while concurrently resulting in a smoother phase current.

In the above case, the peak phase current we test with the hall current probe is 25 mA, so $I_q = 25$ mA. $V_d = -L_q I_q \omega_e = 37\mu\text{H} \times 25 \text{ mA} \times 2\pi \times 15625/1024 = -88.7\mu\text{V}$, the maximum value of V_d in program is 32,767, practically corresponding to $-88.7\mu\text{V}$. In programming, V_d value = 0. On the other hand, $V_q - j_f \omega_e = r I_q$, $j_f \omega_e$ remains constant, so $V_q \propto r I_q$.

However, there is a phase difference ignored in Equation (11). Here is the supplementary explanation why such neglect does not matter: the model of motor is a RL series circuit, the voltage across the inductor leads the current through it, this phase difference is also reflected in the differential term L_q of Equation (10). Also we obtain $\vartheta = 0.0137^\circ$, hereby the phase difference can basically be neglected.

Under the equipment conditions shown in Figure 6, we regulate directly the setting value of $V_{q_}\{\text{value}\}$ to tune the torque of positioning control. When $V_{q_}\{\text{value}\} = 10000$, the fiber positioner can run smoothly. Simultaneously, the peak current measured in the oscilloscope is merely 25 mA, which is quite a small current that could be effectively driven directly by the MOSFET inside the MCU. As such, we could optimize the hardware configuration further, which might conceivably involve eliminating the extraneous driver STSPIN233 in Figure 6. Figure 10 [Figure 10: see original paper] exhibits the fluctuations of the phase current while utilizing MCU directly drive the motor. The experimental conditions remained constant throughout, there is no visible difference about the peak value of the phase current, but the sine degree of current has certain improvement. When the MCU is deployed to drive the motor directly, since the rising and falling edges of the MOSFET are merely determined by the internal power bridges, the ramping time of chopper is reduced, the current waveform is destined to be improved in accordance with expectation.

6.1. Positioning Experiment Results

The digital universal microscope 19JC is a high-precision optical measuring instrument with high-resolution image measuring system. We utilize ceramic head on fiber as microscope's optical measurement target, i.e., the center of the ceramic head, then drive the central rotary axe of SCARA to implement a back and forth movement of a full circle, record ceramic head's central coordinate after every positioning operation. Figure 11 [Figure 11: see original paper] shows the results of the positioning experiment.

We discover that the center coordinates of ceramic head do not coincide at one position, but swing back and forth in two primary positions, which indicates that there is a certain backlash in the mechanical structure of the fiber positioner. After a full circle of central axe's motion, the mechanical clearance brings a $160 \mu\text{m}$ error to the fiber ceramic head. The histogram of positioning errors in Figure 11 gives the repeatable positioning error of fiber positioner at two fixed coordinates, both of which are merely several μm . Compared to mechanical clearance, repeatable positioning error is quite light. Results of positioning

experiment validate the engineering feasibility of the open-loop positioning control method based on FOC proposed in this paper. Additionally, the proposed method only takes about 25 mA phase current at most, and the internal MOSFETs of the MCU can completely afford to drive it, which not only reduces the cost of the hardware case, but also saves valuable physical space for fiber positioner.

6.2. Discussion: Switching Losses

When driving a MOSFET with a carrier frequency as high as 1 MHz, an immediate concern might be: will the switching losses of the power device increase dramatically and adversely affect the system? To figure it out, this section provides a detailed assessment of this issue.

MOSFET (Metal-Oxide-Semiconductor Field-Effect Transistor), is the most important component in low-voltage and low-power motor drive systems. During the MOSFET turn-on process, as the V_{DS} voltage decreases, the I_D current gradually increases, and there will be an overlap between voltage and current, resulting in power losses. When the MOSFET is fully conducting, the V_{DS} voltage is not equal to zero due to the conducting resistance between the DS terminals, resulting in voltage drop and power loss. The turn-off process of MOSFET is similar to the turn-on process, so the turn-off process also generates losses.

When making recommendation of MHz PWM frequency for the motor, we must consider the impact of switching losses. Generally speaking, the principle of switch loss in the operation of a MOSFET is shown in Figure 12 [Figure 12: see original paper]. The relationship graph between the drain-source voltage $V_{DS}(t)$ and drain-source current $I_D(t)$ can well explain the switch loss during the transition process. As shown in the upper waveform, there is a transient voltage and current during the $t_{SW(ON)}$ and $t_{SW(OFF)}$ periods, and the MOSFET capacitance is charged and discharged. Before the $V_{DS}(t)$ drops to the final conduction state ($V_{DS}(t) = I_D \times R_{DS(ON)}$), the full load current I_D flows through the MOSFET. Conversely, during turn-off, $V_{DS}(t)$ gradually rises to the final value of the turn-off state before the MOSFET current drops to zero. The overlapping part of voltage and current during the switching process is the source of switching loss.

The main power loss during the entire process is reflected in the “conduction process,” “turn-off process,” and “conduction state,” while the loss in the turn-off state is very small and can be neglected.

Switching losses are caused by the non-ideal behavior of the switch, and it is difficult to accurately calculate the switching losses of MOSFETs and diodes. Typically, the switching losses can be estimated based on the triangular waveform in Figure 12, using the following formula:

$$P_{SW} = \frac{V_D I_D}{2} (t_{SW(ON)} + t_{SW(OFF)}) f_{PWM}$$

Here V_D is the drain-source voltage when the MOSFET is turned off, I_D is the channel current when the MOSFET is turned on, $t_{SW(ON)}$ and $t_{SW(OFF)}$ are the turn-on and turn-off times, respectively. These parameters can be obtained from the MOSFET datasheet. $t_{SW(ON)}$ is the sum of turn-on delay time $t_{D(ON)}$ and rise time t_{rise} , while $t_{SW(OFF)}$ is the sum of turn-off delay time $t_{D(OFF)}$ and fall time t_{fall} .

The datasheet of STSPIN233 only provides typical values of t_{rise} and t_{fall} . For the unknown parameters of $t_{D(ON)}$ and $t_{D(OFF)}$, t_{rise} and t_{fall} are used as a substitute, because the turn-on/turn-off delay times are usually not much different from the rise/fall times. I_D is assumed to be the peak phase current of 25 mA, V_D is assumed to be the bus voltage of 3.3 V, and the carrier frequency is 1 MHz. The switching power loss is calculated as 0.825 mW. The conduction loss is 0.25 mW ($3.3 \text{ V} \times 25 \text{ mA} \times R_{DS(ON)}$). Even at a carrier frequency of 1 MHz, the switching power loss of the MOSFET is only comparable to the conduction loss, and the average conduction time in a 1 μs period is much greater than 20 ns, so the conduction loss still dominates and switching loss is not significant enough to cause too much attention.

7. Conclusion

Because of compact structure, miniature hollow-cup PM is widely utilized in notebook fans, electric toothbrushes or certain space-constrained applications, which facilitates typical drive control of rotation without concern for detailed speed-tuning. As the actuator of moving optical fiber in telescopes system, fiber positioner is usually required to be manufactured as small as possible to obtain more excellent spectral coverage [?]. To such applications, miniature hollow-cup PM also plays an indispensable part in precision positioning control, especially for fiber positioner [?] with compact mechanical structures.

However, encoders are not originally included in the design of the compact positioner due to a lack of cost effective commercially available encoders at the very small sizes needed [?]. Smaller fiber positioners invariably encounter this problem during design process. If a viable (small and cheap) encoder can be found it would likely resolve the dilemma of low efficiency in open-loop control. Therefore, we have to adopt an open-loop control strategy similar to DESI [?] to fulfill the precise positioning requirement in MOS. Obviously, this is a way of sacrificing efficiency to achieve the goal.

Compared to the use of a stepper method to control fiber positioner (where phase current reaches as high as 66 mA [?]), deploying open-loop FOC control based on voltage only takes 25 mA phase current and 3.3 V bus voltage. Under the high carrier frequency control mode, with each full rotation of the motor, up to 32768 electronic commutations ensure a smooth and continuous motor

current with a smaller peak value. The current consumption is much smaller than the open-loop drive used in DESI, which can consume hundreds of mA [?]. Such a small phase current is sufficient to drive the motor using the MCU's built-in MOSFET bridge without additional voltage rails, saving on power devices and buck components, reducing costs, and also saving valuable PCB (the Printed Circuit Board) space. The voltage-open-loop FOC drive for fiber positioner can also keep up with the time consumption of the stepper method, meaning that the open-loop FOC is more efficient without the problem of passive holding torque that the stepper controller has. Taken together, the results of this paper can inform the present and future miniature hollow-cup PM control of fiber positioner or positioning control of precision instruments with compact design.

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