

Sensorless Brushless DC Motor Control with the Back-EMF Feedback Circuit Integrated into a Microcontroller

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-----ABSTRACT-----

Permanent-magnet brushless DC (BLDC) motors are compact, efficient and mechanically simple, and they are widely used in service robots, personal electric vehicles and multirotor unmanned aerial vehicles. Rotor position is normally obtained from Hall-effect sensors, but these sensors add cost and volume, complicate assembly and alignment, and reduce environmental robustness. Back-EMF sensing is the most mature sensorless alternative, yet its conventional analog implementation requires a low-pass filter and a comparator: the filter attenuates an already weak signal and introduces a phase delay that varies with speed, which is the dominant source of commutation error. This paper presents a sensorless BLDC drive in which the entire back-EMF feedback path is integrated into a Holtek HT32F52352 microcontroller. The terminal-voltage comparison method is combined with the PWM-switching estimation method so that the external low-pass filter is replaced by external-interrupt (EXTI) sampling synchronized to the PWM off-time, and the external comparator is replaced by software comparison of on-chip 12-bit analog-to-digital converter (ADC) readings. Speed regulation, torque compensation, locked-rotor detection and status reporting are implemented on the same device. Measurements at 600, 1200 and 1800 rpm show that the digital feedback obtained inside the microcontroller reproduces the modal shape of the analog waveform, that the commutation control signals show no appreciable delay or misalignment relative to the hardware comparator, and that the overall behavior matches Hall-sensor control. Applying PWM to the upper arm only, with the lower arm switched, is also shown to consume less power per unit commutation frequency than applying PWM to both arms.

KEYWORDS:- sensorless brushless DC motor; back electromotive force; zero-crossing point detection; terminal voltage sensing

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I. INTRODUCTION

Permanent-magnet brushless DC (BLDC) motors replace the mechanical brushes and commutator of a brushed machine with electronic commutation. The result is a simple, robust and compact machine with a long service life and high efficiency, and BLDC drives are now used in home appliances, health care, aerospace and robotics [1], in electric vehicles, robots and medical instruments [2], and in industrial equipment, fans, drones and automotive systems [3]. Service robots must run accurately for long periods, and the durability of a brushless machine supports continuous, energy-efficient operation; personal electric vehicles must respond quickly to road conditions and must do so without excessive thermal stress; and multirotor unmanned aerial vehicles depend directly on speed-control accuracy for flight stability. Pillay and Krishnan [4] compared permanent-magnet synchronous motors and BLDC motors at equal output power and reported that the BLDC machine is superior in copper loss, torque and current, while the synchronous machine is better in torque ripple, high-precision position control and field weakening; the former is normally equipped with Hall-effect devices and the latter with an optical encoder.

A conventional BLDC drive uses three Hall-effect sensors to detect rotor position in real time and to set the commutation instants. Those sensors, however, add cost and volume, complicate mounting and alignment, and are sensitive to temperature, humidity and vibration, all of which reduce reliability. Sensorless control, in which the machine's own back electromotive force (back EMF) replaces the physical sensors, is therefore an important line of research. Its principle is straightforward: under six-step 120° conduction only two phases carry current at any instant while the third is open, so the terminal voltage of the open phase is its back EMF and the zero-crossing point (ZCP) of that voltage can be detected directly. Because the ZCP lags the commutation instant by 30 electrical degrees, delaying each detected ZCP by 30 electrical degrees gives the correct commutation instant.

In the conventional analog implementation, however, the measurement requires a low-pass filter to suppress the high-frequency noise produced by PWM switching and a comparator to extract the ZCP. The filter

attenuates a signal that is already weak, and the phase delay it introduces varies with speed, which makes it the dominant source of commutation error. As microcontroller clock rates have risen and peripheral integration has improved, performing these analog functions digitally inside the microcontroller has become a practical alternative that also reduces circuit cost, board area and failure risk.

The back-EMF method is the most mature and most widely used technique for sensorless rotor-position estimation, and its analog implementations fall into three groups. Iizuka et al. [5] cascade a high-pass filter, a low-pass filter and a phase shifter to obtain the commutation signal from the three terminal voltages and the motor neutral point. Damodharan and Vasudevan [6] build a virtual neutral point to serve as the comparison reference for the terminal voltages. Shao et al. [7] sample the undriven phase during the PWM off-time and thereby avoid measuring the neutral-point voltage altogether. The circuits, limitations and operating ranges of these three methods are described in Section II, and they form the starting point for the present design. Recent work has concentrated instead on improving the accuracy and stability of back-EMF sensing under non-ideal conditions; the five studies most relevant to this paper are reviewed below.

Sagar et al. [8] observe that sensorless control of a three-phase permanent-magnet BLDC motor normally estimates three virtual position signals from the line-to-line terminal voltages and updates the commutation sequence three times per electrical cycle. They show that for steady-state and moderate transient operation a single virtual rotor-position estimate per cycle is sufficient, the remaining commutation instants being taken from a fixed predefined commutation sequence, so that the sequence need not be updated at every step and the computational burden falls. The same arrangement allows the virtual Hall signals to be generated with only one filter circuit, which reduces the hardware count and improves drive reliability. The authors also note that the phase delay of the low-pass filter varies with speed and must be compensated.

Li and Zhou [2] address the difficulty of detecting the ZCP at low speed, where the back EMF is very small. Their method computes the line-to-line back EMF from the mathematical model of the machine, passes it through a low-pass filter whose cutoff frequency is altered with speed in order to reduce disturbance, and detects the ZCP of the line-to-line back EMF using a G function. To correct the commutation delay introduced by the filter they combine an open-loop compensation derived from the filter cutoff frequency with a closed-loop PI compensation based on the integral of the three-phase back EMF, and they add a speed-calculation algorithm that raises the speed-update rate and so reduces the speed-feedback delay at low speed. Simulation and experiment both confirm the stability and reliability of the scheme.

Park et al. [3] examine the unbalanced ZCP detection that arises when the ZCP is obtained by comparing the terminal voltage with half of the dc-link voltage. A dc-link voltage drop, a dc-link offset voltage and asymmetric machine back EMFs all cause the intervals between successive ZCPs to lengthen and shorten repeatedly even at constant speed, which increases current and torque ripple and can make sensorless operation unstable. Because BLDC drives are typically low-cost, small-capacity systems, any compensation must be simple; the authors therefore propose an algorithm that uses only the existing ZCP signals, requires no change to the hardware configuration, and restores balanced commutation instants. The method is verified on a prototype motor mounted in an automotive fan module.

Li et al. [1] extend back-EMF sensing to a drive that alternates between two-phase and three-phase conduction. Working from the mathematical model and the conduction modes of the machine, they identify the conduction mode from the rate of change of the non-commutating phase current and the conduction state of the corresponding power device from the zero crossing of the line back EMF, which locates the rotor over the full speed range. A double closed loop with an inner current loop and an outer speed loop keeps operation smooth under no load and under load and at both low and high speed, and the simulation results agree with the theoretical analysis.

Jia et al. [9] treat the circular winding brushless dc (CWBLdc) motor, whose windings are connected in a circle inside the machine so that there is no neutral point and the phase currents cannot conveniently be measured. Existing sensorless algorithms are aimed at three-phase star-connected or open-winding machines, and the conventional back-EMF method yields only the zero-crossing point rather than the continuous, precise rotor position such a machine requires, so none of them applies. The authors propose a sensorless algorithm based on phase-voltage and dc-current detection that estimates the fundamental harmonic amplitude of the phase current from the dc current, which removes the need for direct phase-current measurement and simplifies the circuit; the error of the algorithm is derived and its accuracy analyzed. Experiments on a 25-phase prototype show performance equivalent to control with a position sensor.

Three observations follow from this body of work. First, the dominant error in back-EMF sensing is the phase delay of the filter, which varies with speed and must be compensated [2], [8]. Second, at low speed the back EMF is small and the relative noise large, so the ZCP is hard to detect [2]. Third, dc-link fluctuation and asymmetric back EMFs unbalance the commutation instants and amplify current and torque ripple [3]. The prevailing approach is to retain the analog filter and comparator and to compensate their side effects algorithmically [1]–[3], [8], [9]. This paper takes a different route: the analog filter and comparator are replaced

by the external interrupt and the on-chip ADC of the microcontroller, which removes the source of the phase delay in hardware while preserving the simple circuit and the low-speed capability of the terminal-voltage comparison method.

Concretely, the hardware low-pass filter is replaced by external-interrupt (EXTI) sampling synchronized to the PWM off-time, and the hardware comparator by software comparison of on-chip 12-bit ADC readings. Both substitutions are verified experimentally at 600, 1200 and 1800 rpm against the analog circuits they replace, and the integrated method is then compared with Hall-sensor control at the same speeds to confirm that the two produce equivalent back-EMF feedback waveforms without any physical position sensor or external analog feedback circuit. The effect of the upper- and lower-arm PWM strategy on the feedback signal and on power consumption is quantified as well, and the measurements support applying PWM to the upper arm only while the lower arm is switched. The remainder of this paper is organized as follows. Section II reviews the principles of brushless DC motor control and compares the three analog back-EMF estimation methods. Section III describes the microcontroller, the drive and feedback circuits, and the software architecture. Section IV presents the experimental results, and Section V concludes.

II. PRINCIPLES OF BRUSHLESS DC MOTOR CONTROL

2.1 Structure and Governing Relations

A brushless DC motor consists of the machine itself and a drive system, as shown in Fig. 1. The rotor is formed by permanent magnets and the stator by three sets of wound coils. The drive system is built from power transistors and integrated circuits; it accepts external start, stop, direction and speed commands, switches the power transistors according to rotor position so as to produce continuous torque, feeds the speed back, and provides interrupt and status-display functions.

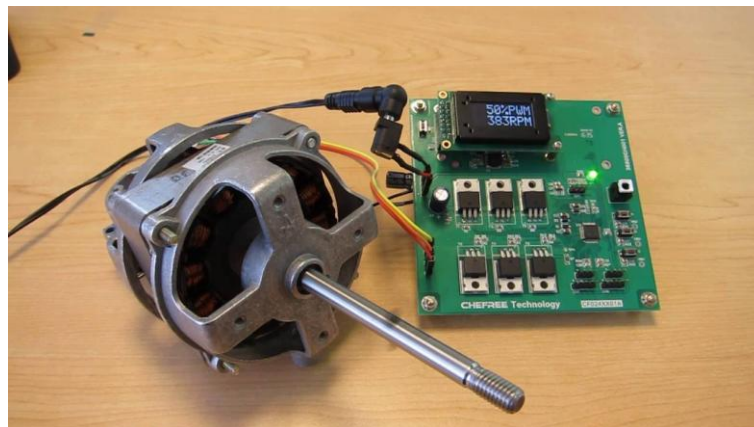


Fig. 1. Brushless DC motor drive system.

If the angular rate of change of the current is ω_I and the number of pole pairs is p , the rotor angular velocity is

$$\omega = \omega_I / p \quad (1)$$

The torque, the back EMF and the output power of the machine are

$$T = I \cdot k_I \quad (2)$$

$$e = \omega \cdot k_e \quad (3)$$

$$P = T \cdot \omega \quad (4)$$

where I is the stator current, k_I is the torque constant and k_e is the back-EMF constant. Equations (2) and (3) show that current is proportional to torque and that the back EMF, and hence the terminal voltage, is proportional to speed. The second relation is precisely what allows the back-EMF method to recover rotor position and speed, and it also explains why the signal becomes weak and the ZCP hard to detect at low speed.

2.2 Drive Method and Commutation Instants

Square-wave drive is available in a 120° and a 180° conduction form. Under 120° conduction, two channels are switched on at each change of the position signal, commutation occurs every 60 electrical degrees, six commutations make up one electrical cycle, and every switch is off for an average of 60 degrees, so shoot-through of an upper and lower arm cannot occur. Under 180° conduction, three channels are switched on at a time and consecutive switches overlap by 60 degrees, so a dead time must be designed in. Because sinusoidal drive is considerably more complex to control and 180° conduction requires a dead time, this work adopts 120° conduction with the commutation sequence $AC \rightarrow BC \rightarrow BA \rightarrow CA \rightarrow CB \rightarrow AB$, illustrated in Fig. 2.

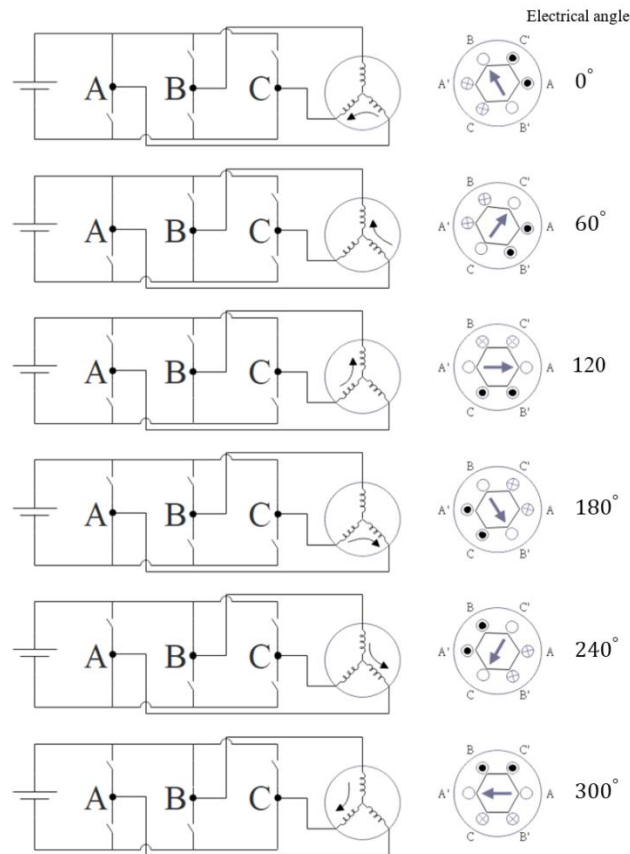


Fig. 2. Square-wave 120° conduction drive scheme.

Electronic commutation can only be performed accurately if the rotor position is known, and the phase current must be kept in step with the back EMF if the power consumption is to be minimized. The back EMF is a direct function of rotor position, as shown in Fig. 3, so switching the current in step with the back-EMF waveform is what aligns the phase current with the back EMF, as shown in Fig. 4.

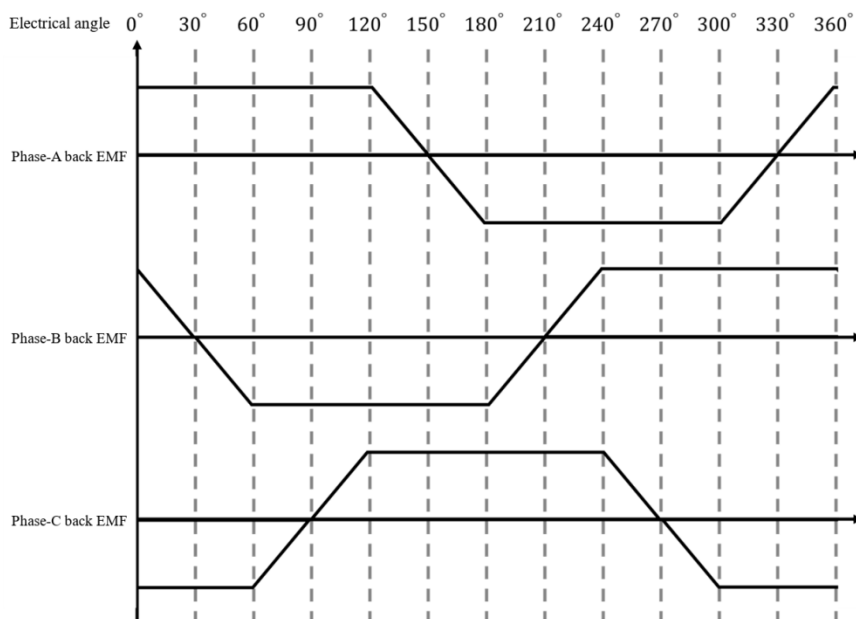


Fig. 3. Relationship between rotor angle and back EMF.

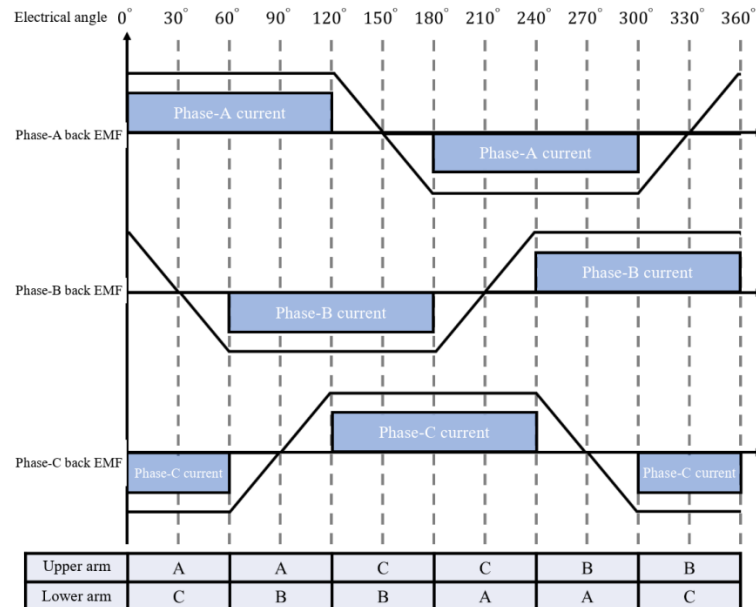


Fig. 4. Alignment of the back-EMF waveform with the phase-current waveform.

Rotor position can be obtained either with Hall-effect sensors or without them. In the sensed arrangement, three Hall devices inside the machine divide one rotor revolution into six equal parts and so identify six rotor positions, which, combined with 120° conduction, allows the power transistors to be switched so that the phase current stays aligned with the back EMF. In the sensorless arrangement the devices are removed, which widens the range of environments in which the machine can be used and lowers maintenance cost, but the rotor position must then be inferred by other means.

2.3 Terminal-Voltage Detection Method

Iizuka et al. [5] obtain the commutation signal from the three terminal voltages and the motor neutral point through a cascade of filters and a phase shifter. The circuit has three parts: a high-pass filter removes the dc bias from the terminal voltage, a low-pass filter removes the high-frequency noise caused by PWM, and a phase shifter then yields the correct commutation signal. Its drawback is that the phase response of the multistage filter is sensitive to speed, so the estimated signal carries a floating phase error that delays commutation.

2.4 Terminal-Voltage Comparison Method

Because only two phases carry current in any step, the third phase can be used to measure the back EMF. In principle it is enough to monitor that phase terminal and the motor neutral point and to compare them in order to find the zero crossing of the back EMF. In most machines, however, the neutral point is not brought out, so the usual practice is to synthesize a virtual neutral point from a resistor network; in theory its potential equals that of the neutral point of a Y-connected winding.

The detection scheme of Damodharan and Vasudevan [6] is very simple and has been in use for some time, but it has drawbacks. When PWM is used to regulate speed or torque and current, the virtual neutral point fluctuates at the PWM frequency and carries a very high voltage and considerable high-frequency noise, so a voltage divider and a low-pass filter are needed to reduce the voltage and suppress the noise. If the dc bus voltage is 300 V, for example, the neutral-point potential can vary anywhere between 0 and 300 V, whereas a comparator input is typically limited to about 5 V; the divider required to bridge that gap clearly reduces the sensitivity of the measurement, and it does so precisely when sensitivity matters most, at start-up. The low-pass filter also introduces a fixed delay that is independent of rotor speed, so the fraction of the electrical cycle that the delay occupies grows as speed rises. That delay disturbs the alignment between current and back EMF and causes serious commutation problems at high speed, which is why the method tends to have a narrow speed range and poor starting behavior.

2.5 PWM-Switching Estimation Method

Shao et al. [7] showed that with an appropriate PWM strategy the back EMF can be extracted directly from the motor terminal voltages, so that the neutral-point voltage need not be measured at all. For the two driven phases the PWM signal can be arranged in three ways: applied to the upper-arm switch only, with the lower arm held on for the duration of the step; applied to the lower-arm switch only, with the upper arm held on; or applied

to both arms together. The scheme adopts the first arrangement. PWM is applied only to the upper arm, the lower-arm switch is toggled only to provide commutation, and the remaining phase is left undriven for measurement, the back EMF on that phase being sampled during the PWM off-time.

During the PWM off-time the motor inductance forces the current to find a freewheeling path. The driven phase voltage is then at or slightly below ground potential, and the neutral-point potential is a function of the phase back EMFs alone, unaffected by any external drive voltage. Suppose that in a given step phases A and B carry current and phase C is off, that the upper switch of phase A is pulse-width modulated to regulate speed or current, and that the lower switch of phase B is on throughout the step. When the upper power transistor of the phase-A half bridge turns off, the current freewheels through the diode, and the neutral-point voltage during that freewheeling interval determines the back EMF on phase C. The method offers three advantages over the conventional schemes. First, no voltage divider is used, so the measured back EMF is not attenuated and sensitivity is high. Second, the back EMF is sampled only during the PWM off-time, and this synchronous sampling removes the problems associated with switching noise. Third, the measurement is referenced to ground, which eliminates the difficulty of conditioning an attenuated signal to fit a voltage limit.

2.6 Comparison and Design Choice

The characteristics of the three methods are summarized in Table 1.

Table 1. Comparison of the three back-EMF estimation methods.

Estimation method	Neutral-point signal	Filter circuit	Phase shifter	Commutation signal at low speed
Terminal-voltage detection [5]	Required	Required	Required	Not available
Terminal-voltage comparison [6]	Not required	Required	Not required	Available
PWM-switching estimation [7]	Not required	Not required	Required	Not available

Table 1 shows that removing the filter circuit calls for the PWM-switching estimation method, whereas the terminal-voltage comparison method has the simpler circuit model and is the only one of the three that can estimate the commutation signal at low speed. This work therefore takes the terminal-voltage comparison method as its basis and imports from the PWM-switching method the idea of sampling synchronously during the PWM off-time. The combination makes it possible to dispense with both the external filter and the external comparator and to integrate the whole back-EMF feedback path into the microcontroller.

III. SYSTEM DESIGN AND IMPLEMENTATION

3.1 Microcontroller and Peripherals

The controller is a Holtek HT32F52352, a 32-bit high-performance, low-power device built around an ARM Cortex-M0+ core running at a system frequency of 48 MHz and integrating an ADC, I2C, USART, UART, SPI, I2S, GPTM, MCTM, SCI, CRC-16/32, RTC, WDT, PDMA, EBI and USB 2.0 FS [10]. The peripherals used in this implementation are as follows:

- External interrupt (EXTI). Sixteen edge detectors, each able to generate a wake-up event or an interrupt request independently, used here to detect the transitions of the three terminal voltages.
- Analog-to-digital converter (ADC). A 12-bit all-channel converter with twelve external analog channels and two internal channels, together with an analog watchdog that monitors whether an input voltage stays within a preset threshold, used to sample the terminal voltages for software comparison.
- Analog comparator (CMP). Two general-purpose comparators, each able to raise an interrupt to the NVIC through the EXTI wake-up event management unit.
- General-purpose I/O (GPIO). Pins shared with alternate functions, with dedicated control and configuration registers for external interrupts.
- Motor control timer (MCTM). A 16-bit up/down counter, four 16-bit capture/compare registers, a 16-bit counter reload register and an 8-bit repetition counter, able to produce compare-match, PWM and complementary PWM outputs and to support motor control and Hall-sensor control.
- General-purpose timer (GPTM). Used for timing, input pulse-width measurement and PWM waveform generation.
- Universal asynchronous receiver/transmitter (UART). Full-duplex data exchange with line-status interrupts, used for status reporting and parameter monitoring.

Because the electrical period of a BLDC motor is short, the PWM period is normally on the order of microseconds. Controlling the duty cycle from a software loop would not only complicate the program but occupy most of the processor time, so that control commands could not be issued on schedule and the drive would misbehave. A microcontroller with PWM hardware built in is therefore essential, and the HT32F52352 meets that requirement.

3.2 Drive Circuit

The drive circuit, shown in Fig. 5, uses IR2103S gate drivers with power MOSFETs in a high-side and low-side arrangement. Each IR2103S drives one phase leg and requires two MOSFETs, so three drivers and six MOSFETs are needed for the three phases, and a large part of the drive cost lies here. The MOSFET power rating required varies with the application, so the devices must be selected with cost-effectiveness in mind.

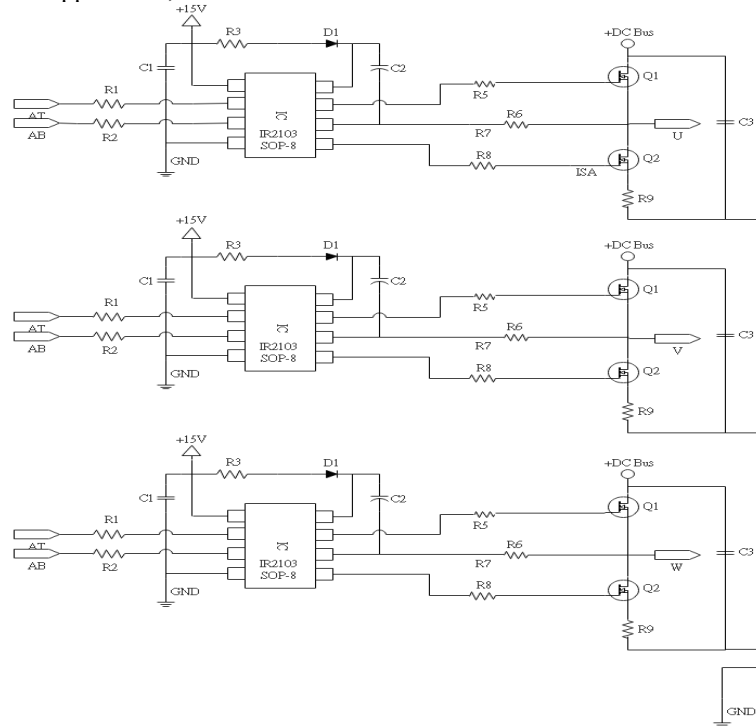


Fig. 5. Brushless DC motor drive circuit.

3.3 Integrating the Back-EMF Feedback Path into the Microcontroller

Figure 6 shows the circuit structure of the terminal-voltage comparison method. The average voltages of the three phases are obtained by low-pass filtering, and a comparator then yields directly a feedback signal that lags the back EMF by 30 electrical degrees, so that commutation can be performed without a phase shifter. The circuit works as follows: when the average voltage of phase A exceeds that of phase C, signal EA is obtained; when the average voltage of phase B exceeds that of phase A, signal EB is obtained; and when the average voltage of phase C exceeds that of phase B, signal EC is obtained. The estimation circuit needs neither the real nor a virtual neutral point, only a mutual comparison of the three average terminal voltages, which makes it simple and suitable for low-speed operation.

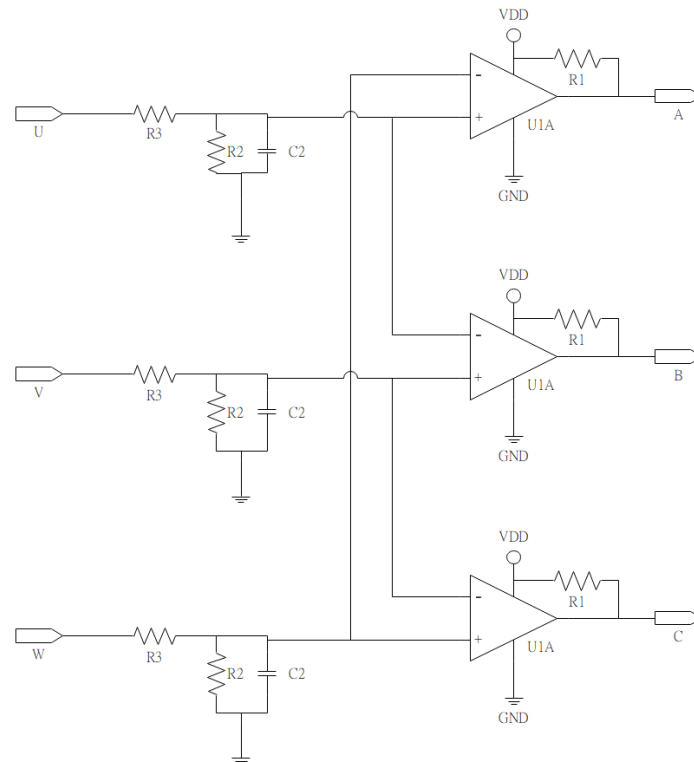


Fig. 6. Terminal-voltage comparison back-EMF feedback circuit.

The central idea of this work is to move both analog stages of Fig. 6 inside the microcontroller. First, the low-pass filter is replaced by ADC sampling triggered from the external interrupt. Because the sampling instant is synchronized with the PWM switching, the terminal voltage is captured during the PWM off-time and is therefore free of switching noise, so a usable signal is obtained without analog filtering and the phase delay of the filter is avoided altogether. Second, the hardware comparator is replaced by software comparison of the on-chip 12-bit ADC readings: the program compares the three terminal voltages directly and produces commutation signals equivalent to EA, EB and EC. The feasibility of both substitutions is verified experimentally in Section IV.

3.4 Software Architecture

The system is written in C and developed under Keil C. The software comprises a main program, a commutation subroutine, a timer interrupt routine, a back-EMF sampling interrupt routine, and speed compensation and locked-rotor subroutines.

The main program handles initialization of the speed control. Speed is regulated by PWM, so the CRR and CCR values must be set during initialization: CRR determines the period, and therefore the frequency, of the output waveform, and CCR determines its duty cycle. The system frequency of 48 MHz divided by 1000 gives the 48 kHz PWM frequency used here. The flow is shown in Fig. 7: initialization of the I/O pins, registers and functions; rotor alignment, in which a drive signal of a specific phase is applied before starting so that the back EMF can be read and converted into the relative rotor position; PI control, which adjusts the output from the external control signal once the machine is running; a one-second flag, on which all buffered data are output; and stop, which disables all outputs when a stop signal is received.

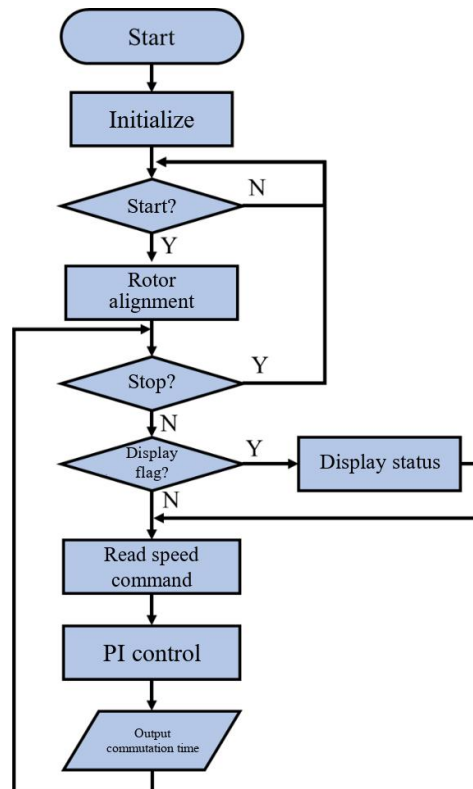


Fig. 7. Flowchart of the main program.

Commutation is driven from a timer interrupt in the order $AC \rightarrow BC \rightarrow BA \rightarrow CA \rightarrow CB \rightarrow AB$. A timer is started immediately after each commutation so that the response time of the machine can be measured and checked against the expected commutation; if commutation has not occurred as expected, the locked-rotor mechanism is triggered. When the subroutine is called it first examines the Hall flag and skips the step if the state is unchanged from the previous call. The flow is shown in Fig. 8: the commutation state is read, and a state of 0 stops the output, while states 1 to 6 switch on the upper and lower arms of A/C, B/C, B/A, C/A, C/B and A/B respectively and start the timer.

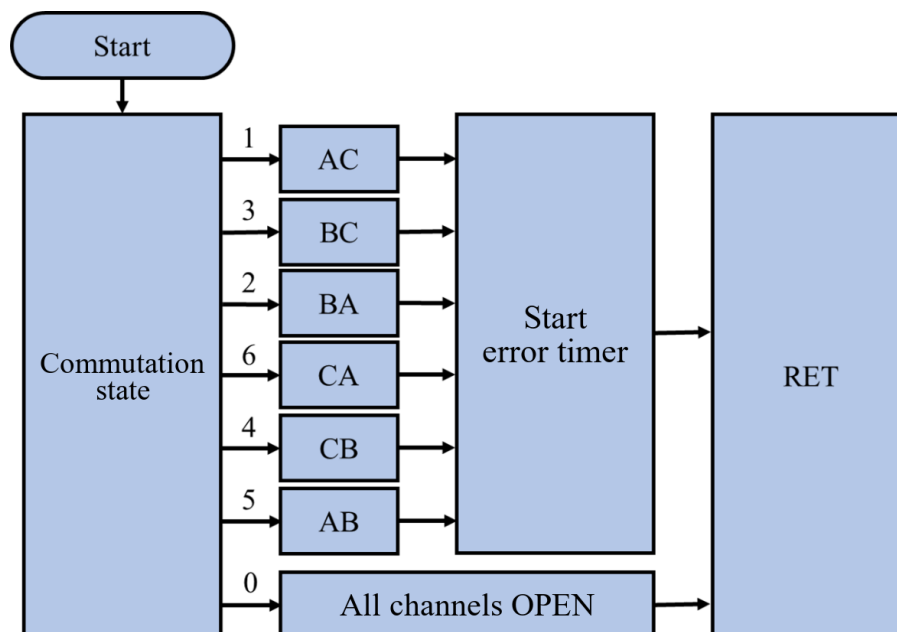


Fig. 8. Flowchart of the commutation subroutine.

All time-related interrupt flags in the system are generated by a single 24-bit counter clocked at 48 MHz with its reload value set to 1000, which gives a 48 kHz working frequency. The commutation time derived from the speed command updates the timer of every call to the commutation subroutine; the flag for the speed compensation subroutine is raised at twice the commutation time; the locked-rotor flag is re-evaluated at every commutation cycle; and an output flag is raised once per second.

When the external interrupt pin returns a feedback signal, the routine first samples that signal with the 12-bit ADC and then determines the rotor position by software comparison. If the position matches the previous commutation position the reading is treated as a false trigger and the loop is left. If it differs, the routine checks whether it matches the target commutation position; if not, a misalignment record is returned. If it does match, the timer started at commutation is sampled, and a misalignment record is likewise returned and the cycle ended if the error is too large. The flow is shown in Fig. 9.

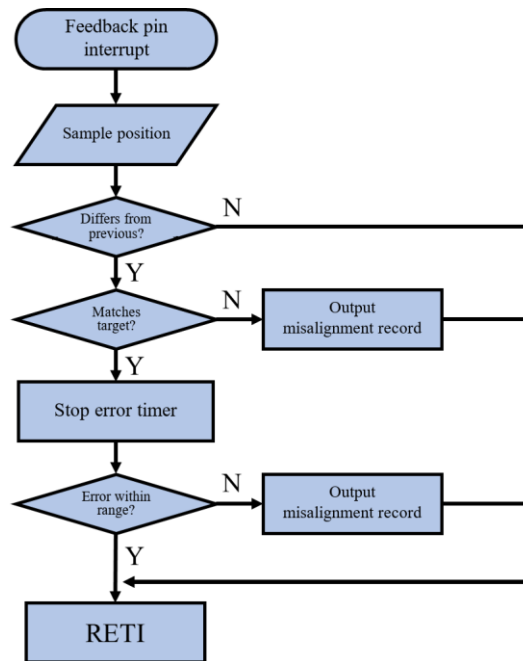


Fig. 9. Flowchart of the back-EMF sampling interrupt subroutine.

The speed compensation subroutine reads the misalignment record to decide whether commutation was early or late. A late commutation increases the PWM duty cycle to raise the torque and an early one reduces it, the correction being computed by a PI controller, and the misalignment record is then cleared. The locked-rotor subroutine outputs a commutation state of 0 and calls the commutation subroutine so that all control pins are opened, and then halts the main program. Both flows are shown in Fig. 10.

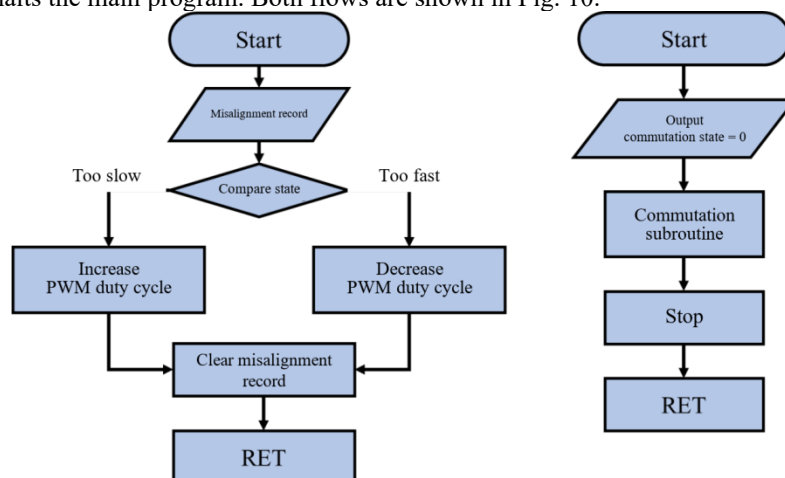


Fig. 10. Flowcharts of the speed compensation and locked-rotor subroutines.

IV. EXPERIMENTAL RESULTS AND DISCUSSION

Waveforms were captured with an oscilloscope and combined with statistical analysis to verify the proposed method step by step. The experiments cover, in order, the choice of lower-arm drive mode, the replacement of the hardware filter circuit by external-interrupt sampling, the replacement of the hardware comparator by on-chip ADC comparison, and a comparison of the integrated method with Hall-sensor control. All measurements were made with the machine unloaded.

4.1 Choice of Lower-Arm Drive Mode

This section compares the two lower-arm drive modes under Hall-sensor feedback control. Figure 11(a) was captured with the lower arm switched only and the upper-arm PWM duty cycle at 20%, and Fig. 11(b) with PWM applied to both arms at the same 20% duty cycle. In both figures channels 1 to 3 are the back-EMF voltages of phases A, B and C, and channel 4 is the Hall-sensor phase feedback signal.

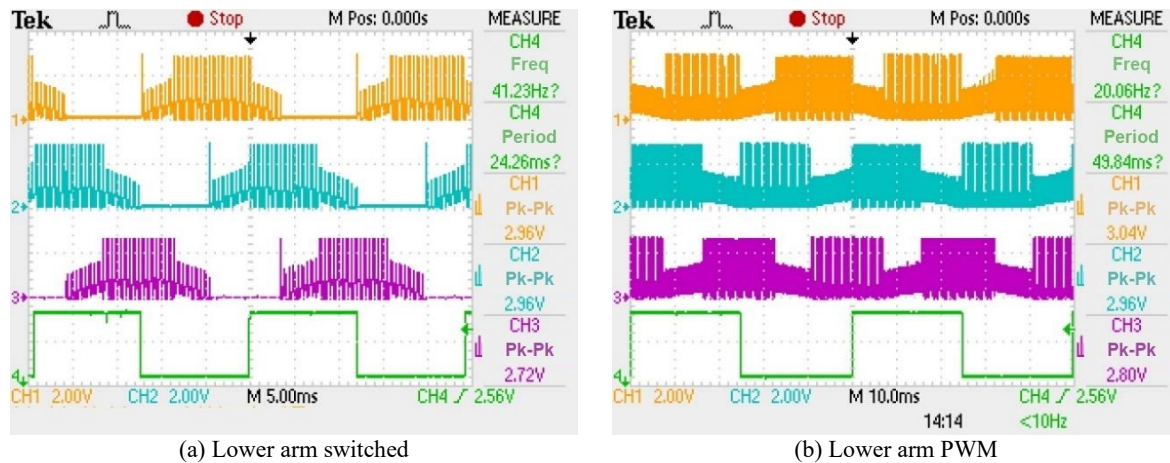


Fig. 11. Effect of the lower-arm drive mode on the back-EMF feedback signal.

In Fig. 11(a) the back-EMF feedback falls to zero whenever the lower arm goes to zero, which narrows the range over which the feedback torque can be controlled and makes the program harder to write. In Fig. 11(b) the back-EMF feedback does not fall to zero when the lower arm does, so a wider control range is available, but the signal is more complex and equally difficult for the program to interpret.

Table 2 gives the measured parameters for the two modes at duty cycles from 20% to 50%. Driving the upper arm with PWM while merely switching the lower arm gives a lower power consumption per unit commutation frequency at every duty cycle, and the difference is largest at low duty cycle, where it is 0.037 against 0.062 W/Hz. Applying PWM to the lower arm therefore widens the torque control range but greatly increases the complexity of the signal, makes the program harder to write, and consumes more power in low-power modes. This work consequently switches the lower arm rather than modulating it.

Table 2. Measured parameters for the two lower-arm drive modes at duty cycles from 20% to 50%.

Lower-arm mode	PWM duty cycle	Frequency (Hz)	Current (A)	Power (W)	Consumption (W/Hz)
Switched	20%	40.65	0.06	1.50	0.037
Switched	30%	63.20	0.09	2.25	0.036
Switched	40%	80.00	0.12	3.00	0.038
Switched	50%	93.40	0.13	3.25	0.035
PWM	20%	20.06	0.05	1.25	0.062
PWM	30%	48.22	0.08	2.00	0.041
PWM	40%	72.68	0.11	2.75	0.038
PWM	50%	87.69	0.13	3.25	0.037

4.2 Replacing the Hardware Filter with External-Interrupt Sampling

This section examines what replacing the hardware filter circuit by microcontroller external-interrupt sampling does to the back-EMF waveform at different speeds. In Fig. 12, panels (a), (c) and (e) are the analog signals captured with the oscilloscope after hardware filtering, and panels (b), (d) and (f) are the digital signals

obtained by external-interrupt sampling. Yellow is the phase-U back EMF, blue phase V and purple phase W. The green trace in the analog panels is the microcontroller's simulated Hall signal, derived from the green timer signal of the digital panels, and is included so that all waveforms can be aligned.

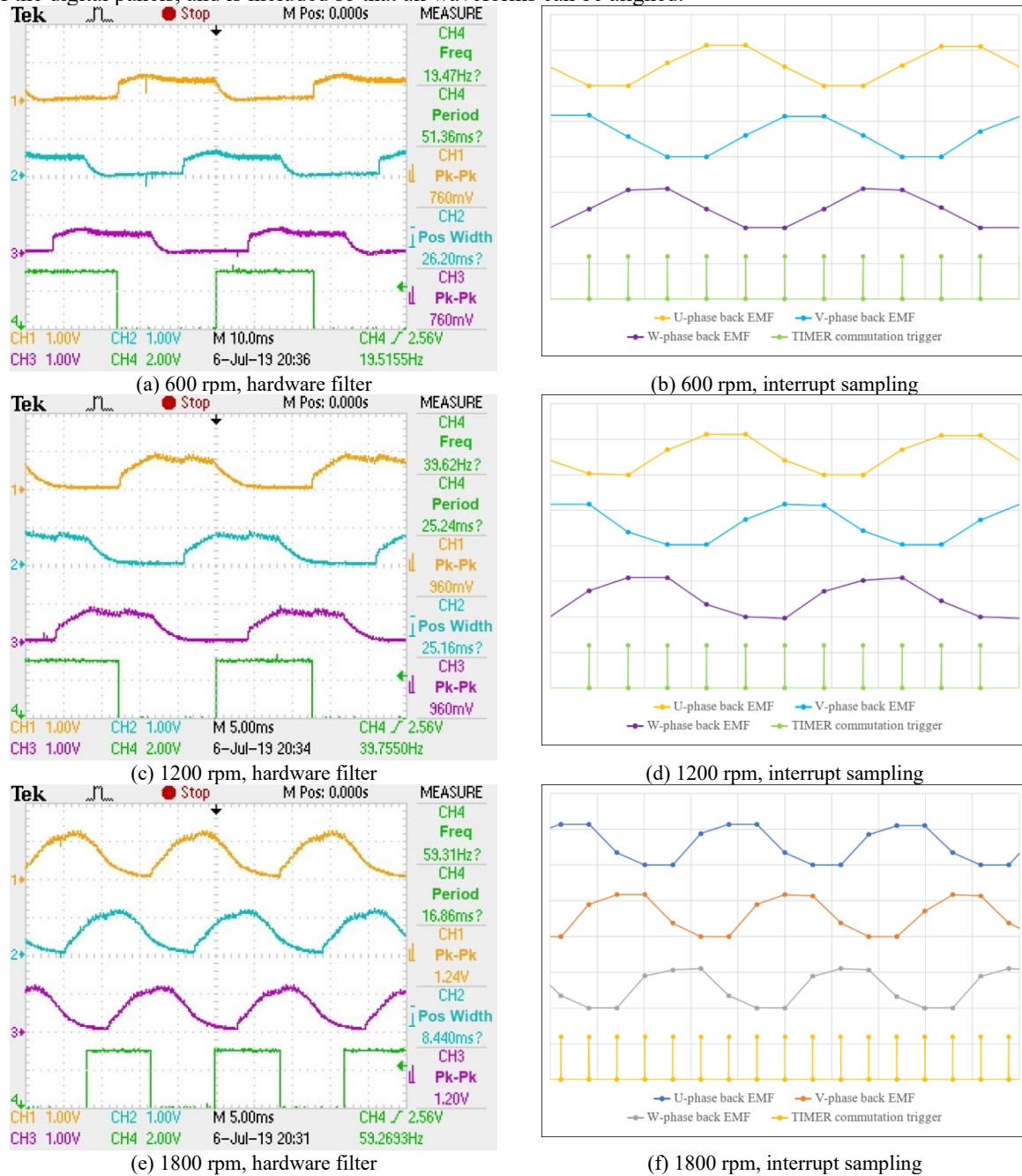


Fig. 12. Back-EMF feedback waveforms obtained with the hardware filter circuit and with microcontroller external-interrupt sampling.

At 600 rpm, Fig. 12(a) and (b), the sinusoidal character of the analog feedback is not pronounced because the speed is low, yet the digital signal obtained by external-interrupt sampling still shows it clearly, which confirms that the comparison values used later remain trustworthy. At 1200 and 1800 rpm, Fig. 12(c) to (f), the analog feedback shows a clear sinusoidal shape and so does the digital signal. The digital signal obtained by external-interrupt sampling therefore has the same modal shape as the analog signal captured after hardware filtering, and external-interrupt sampling can replace the hardware filter circuit.

4.3 Replacing the Hardware Comparator with On-Chip ADC Comparison

This section examines what replacing the hardware comparator by software comparison of on-chip ADC readings does to the control signal at different speeds. In Fig. 13, panels (a), (c) and (e) are the analog signals

captured after hardware comparison and panels (b), (d) and (f) the digital signals obtained by ADC comparison. Yellow is the W-U control signal, blue U-V and purple V-W, and the green trace again serves to align the waveforms.

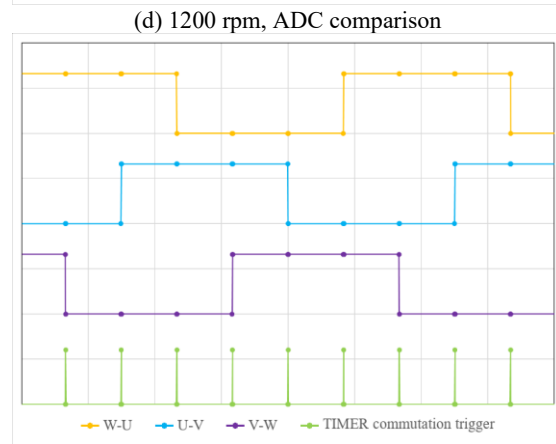
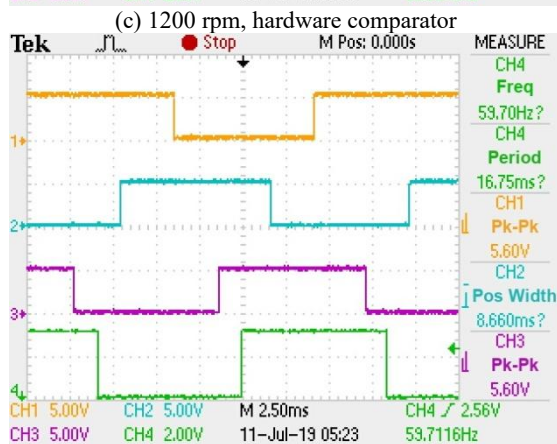
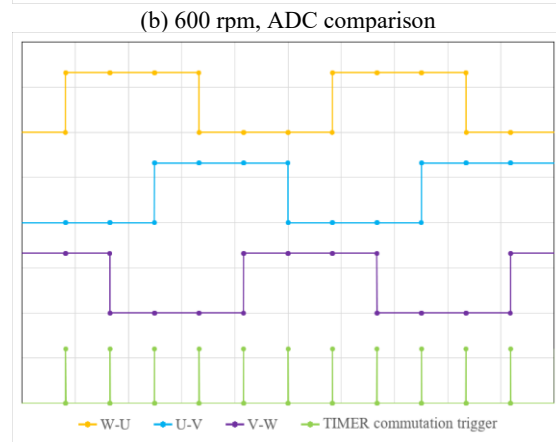
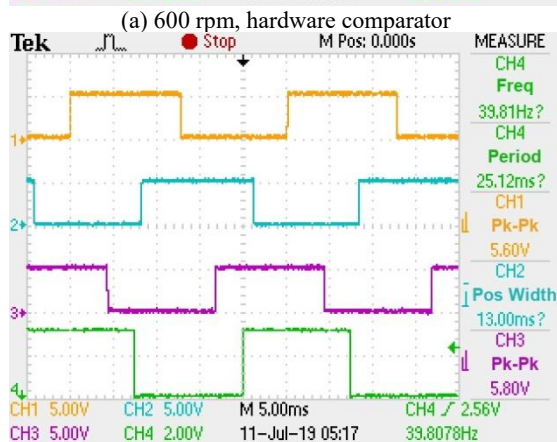
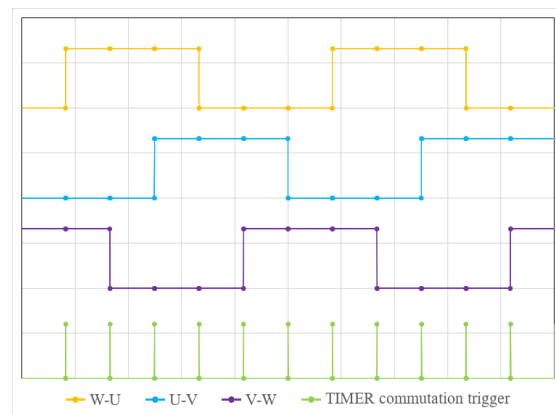
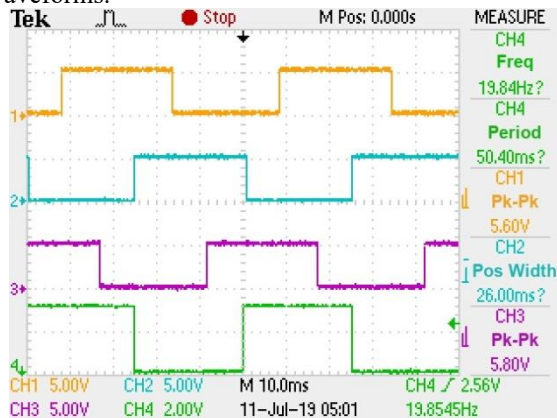


Fig. 13. Commutation control signals obtained with the hardware comparator circuit and with on-chip ADC comparison.

At 600, 1200 and 1800 rpm the control signals produced by on-chip ADC comparison are almost indistinguishable in shape from the hardware comparator output, and accurate real-time feedback is achieved. The digital signal obtained by ADC comparison therefore has the same characteristics as the analog signal captured after hardware comparison, and ADC comparison can replace the hardware comparator circuit.

4.4 Comparison with Hall-Sensor Control

This section compares the microcontroller control method integrated in the two preceding sections with the original Hall-signal control method, and examines the back-EMF feedback waveform at different speeds. In Fig. 14, panels (a), (c) and (e) use Hall-signal control and panels (b), (d) and (f) the proposed microcontroller

control. Yellow is the phase-U back EMF, blue phase V and purple phase W; the green trace is the Hall signal in the former and the simulated Hall signal in the latter, and it is used to align all the waveforms.

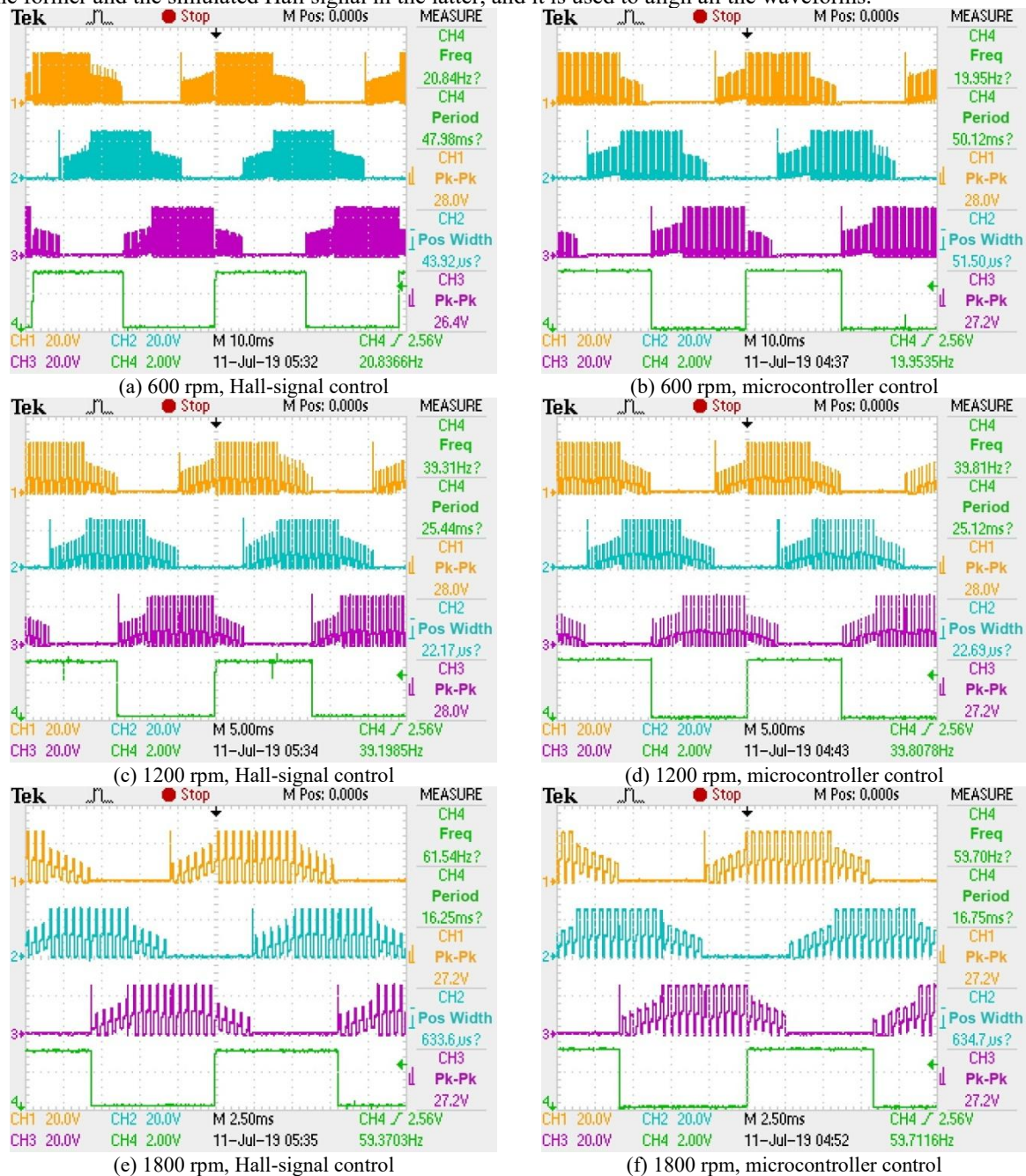


Fig. 14. Back-EMF feedback waveforms under Hall-sensor control and under the proposed microcontroller control.

At 600, 1200 and 1800 rpm the two methods give back-EMF feedback waveforms of almost identical shape, which confirms that the program runs correctly at each speed. Under no load and across this range of speeds, the microcontroller control method therefore achieves the same result as Hall-signal control without any physical position sensor and without an external analog feedback circuit.

V. CONCLUSION

This paper has integrated the back-EMF feedback path of a sensorless brushless DC motor drive into an HT32F52352 microcontroller. The feedback path divides into a filter stage and a comparison stage, and each was replaced by a microcontroller function and verified against measured waveforms:

- Comparing the back-EMF waveforms obtained by external-interrupt sampling with those from the hardware filter circuit showed the same characteristics at 600, 1200 and 1800 rpm, confirming that external-interrupt sampling can remove the filter circuit.
- Comparing the control signals obtained by ADC comparison with those from the hardware comparator showed no appreciable delay or misalignment, confirming that software ADC comparison can remove the comparator circuit.
- Comparing the integrated microcontroller control method with the original Hall-signal method showed no loss of waveform characteristics, achieving a cheaper, more accurate and more easily controlled drive.
- Quantifying the upper- and lower-arm PWM strategy confirmed that applying PWM to the upper arm alone, with the lower arm switched, gives a lower power consumption per unit commutation frequency at duty cycles from 20% to 50%.

Compared with the prevailing approach of retaining the analog circuit and compensating its phase delay algorithmically [1]–[3], [8], [9], the method presented here removes the source of the delay in hardware, and it obtains feedback quality equivalent to Hall-sensor control without adding computational burden, while reducing circuit cost, board area and failure risk.

Future work can extend the method to other machines that provide a back-EMF feedback signal, although a wider range of motor characteristics will have to be studied in order to optimize the control parameters. The complete control method could then be modeled so that the controller determines the machine characteristics and the decision thresholds for the required feedback signal automatically on installation, giving a control chip that adapts itself to any motor. The experiments reported here were all performed with the machine unloaded; the effects of load variation, of starting at speeds well below 600 rpm, and of dc-link voltage fluctuation on the balance of the zero-crossing points [3] remain to be verified.

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