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TMC603A – DATASHEET

Three phase motor driver with current sensing

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1 Features

The TMC603 is a three phase motor driver for highly compact and energy efficient drive solutions. It contains all power and analog circuitry required for a high performance BLDC motor system. The TMC603 is designed to provide the frontend for a microcontroller doing motor commutation and control algorithms. It directly drives 6 external N-channel MOSFETs for motor currents up to 30A and up to 50V and integrates shunt less current measurement, by using the MOSFETs channel resistance for sensing. Protection and diagnostic features as well as a step down switching regulator further reduce system cost and increase reliability.

Highlights

- Up to 30A motor current, up to 50V operating voltage
- 3.3V or 5V interface
- 8mm x 8mm QFN package
- Integrated dual range high precision current measurement amplifiers
- Supports shunt less current measurement using power MOS transistor RDSon
- Integrated break-before-make logic: No special microcontroller PWM hardware required
- EMV optimized current controlled gate drivers – up to 150mA possible
- Overcurrent / short to GND and undervoltage protection and diagnostics integrated
- Internal Q_{GD} protection: Supports latest generation of power MOSFETs
- Integrated supply concept: Step down switching regulator up to 500mA / 300kHz
- Common rail charge pump allows for 100% PWM duty cycle

Applications

- Motor driver for industrial applications
- Integrated miniaturized drives
- Robotics
- High-reliability drives (dual position sensor possible)
- Pump and blower applications

Motor type

- 3 phase BLDC, stepper, DC motor
- Sine or block commutation
- Rotor position feedback: encoder or hall sensor

*) note: The term TMC603 in this datasheet refers to the TMC603A and TMC603
The feature hallFX and related pins have been removed from this documentation

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3 System architecture using the TMC603

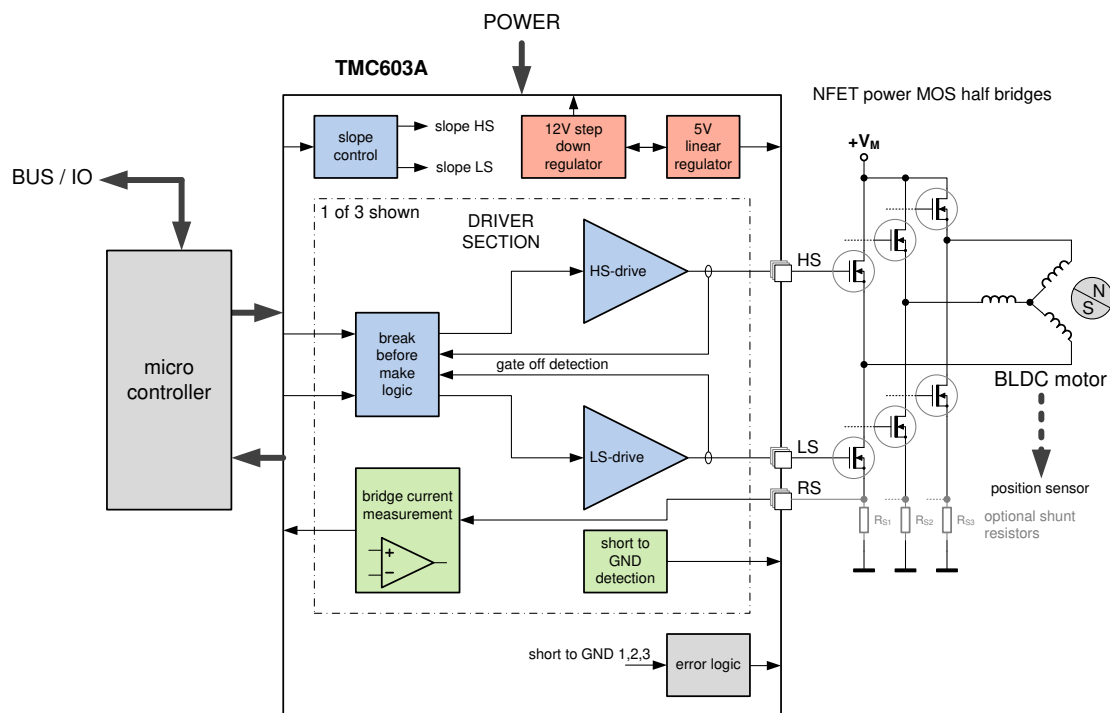


figure 1: application block diagram

The TMC603 is a BLDC driver IC using external power MOS transistors. Its unique feature set allows equipping inexpensive and small drive systems with a maximum of intelligence, protection and diagnostic features. Control algorithms previously only found in much more complex servo drives can now be realized with a minimum of external components. Depending on the desired commutation scheme and the bus interface requirements, the TMC603 forms a complete motor driver system in combination with an external 8 bit processor or with a more powerful 32 bit processor. A simple system can work with three standard PWM outputs even for sine commutation! The complete analog amplification and filtering frontend as well as the power driver controller are realized in the TMC603. Its integrated support for sine commutation saves cost and allows for maximum drive efficiency.

The external microcontroller realizes commutation and control algorithms. Based on the position information from an encoder or hall sensors, the microcontroller can do block commutation or sine commutation with or without space vector modulation and realizes control algorithms like a PID regulator for velocity or position or field oriented control based on the current signals from the TMC603. A sine commutated start-up minimizes motor vibrations during start up.

The TMC603 also supports control of three phase stepper motors as well as two phase stepper motors using two devices.

4 Pinout

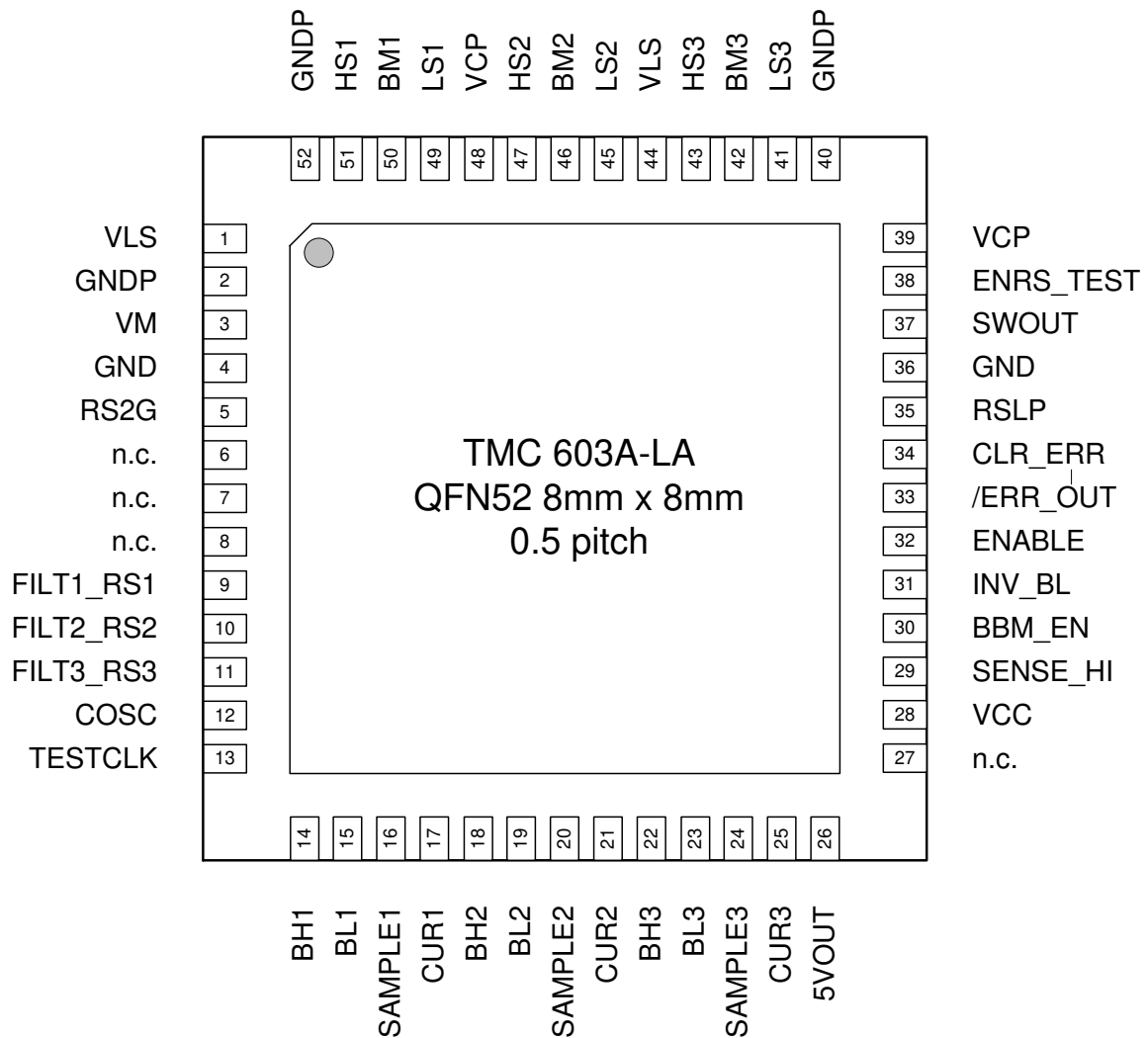


figure 2: pinning / QFN52 package (top view)

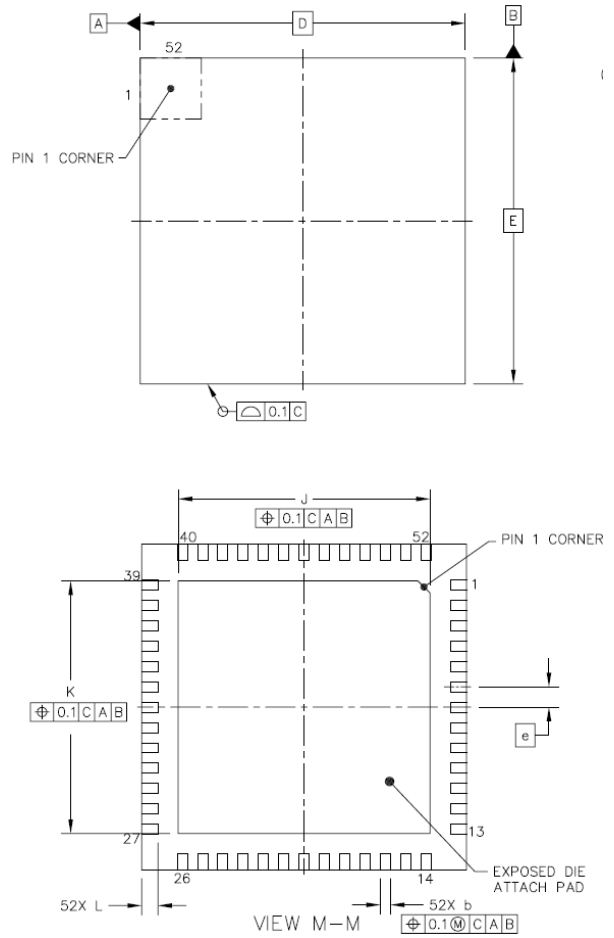
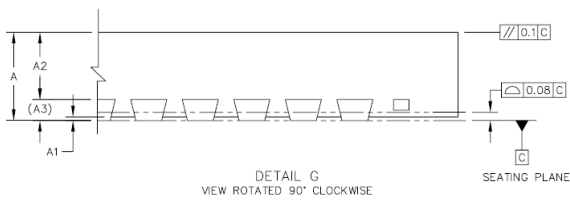
4.1 Package codes

Type	Package	Temperature range	Code/markings
TMC603A	QFN52 (ROHS)	-40°C ... +125°C	TMC603A-LA

4.2 Package dimensions QFN52

REF	MIN	NOM	MAX
A	0.80	0.85	0.90
A1	0.00	0.035	0.05
A2	-	0.65	0.67
A3		0.203	
b	0.2	0.25	0.3
D		8.0	
E		8.0	
e		0.5	
J	6.1	6.2	6.3
K	6.1	6.2	6.3
L	0.35	0.4	0.45

All dimensions are in mm.
Attention: Drawing not to scale.



5 TMC603 functional blocks

5.1 Block diagram and pin description

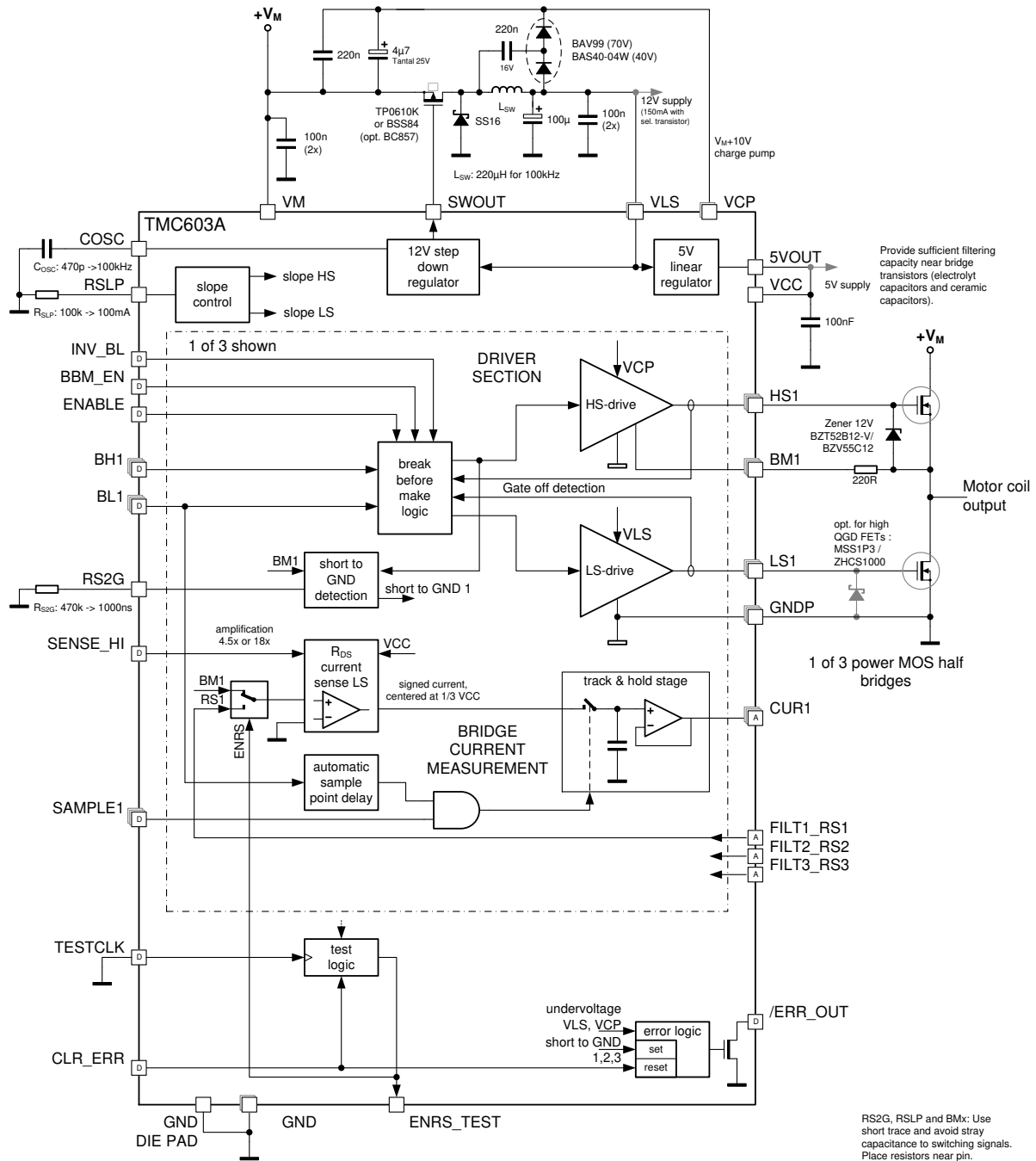


figure 3: application diagram

The application diagram shows the basic building blocks of the IC and the connections to the power bridge transistors, as well as the power supply. The connection of the digital and analog I/O lines to the microcontroller is highly specific to the microcontroller model used.

Pin	Number	Type	Function
VLS	1, 44		Low side driver supply voltage for driving low side gates
GNDP	2, 40, 52		Power GND for MOSFET drivers, connect directly to GND
VM	3		Motor and MOSFET bridge supply voltage
GND	4, 36		Digital and analog low power GND, connect directly to GND
RS2G	5	AI 5V	Short to GND control resistor. Controls delay time for short to GND test
n.c.	6, 7, 8		Do not externally connect these pins (unused outputs)
FILT _x RS _x	9, 10, 11	AI 5V AO 5V	Output of internal switched capacitor filter or input for external sense resistor (select using pin ENRS_TEST)
COSC	12	A 5V	Oscillator capacitor for step down regulator
TESTCLK	13	DI	Test mode input, connect to GND
BH _x	14, 18, 22	DI 5V	High side driver control signal: A positive level switches on the high side
BL _x	15, 19, 23	DI 5V	Low side driver control signal: Polarity can be reversed via INV_BL
SAMPLE _x	16, 20, 24	DI 5V	Optional external control for current measurement sample/hold stage. Set to positive level, if unused
CUR _x	17, 21, 25	AO 5V	Output of current measurement amplifier
5VOUT	26		Output of internal 5V linear regulator. Provided for VCC supply
n.c.	27		Do not externally connect this pin.
VCC	28		+5V supply input for digital I/Os and analog circuitry
SENSE_HI	29	DI 5V	Switches current amplifiers to high sensitivity
BBM_EN	30	DI 5V	Enables internal break-before-make circuitry
INV_BL	31	DI 5V	Allows inversion of BL _x input active level (low: BL _x is active high)
ENABLE	32	DI 5V	Enables the power drivers (low: all MOSFETs become actively switched off)
/ERR_OUT	33	DO 5V	Error output (open drain). Signals undervoltage or overcurrent. Tie to ENABLE for direct self protection of the driver
CLR_ERR	34	DI 5V	Reset of error flip-flop (active high). Clears error condition
RSLP	35	AI 5V	Slope control resistor. Sets output current for MOSFET drivers
SWOUT	37	O	Switch regulator transistor output
ENRS_TEST	38	DI 5V O 12V	Enables sense resistor inputs rather than R _{DS(on)} measurement. Test multiplexer output
VCP	39		Charge pump supply voltage. Provides high side driver supply
LS _x	41, 45, 49	O 12V	Low side MOSFET driver output
BM _x	42, 46, 50	I (VM)	Sensing input for bridge outputs. Used for MOSFET control and current measurement.
HS _x	43, 47, 51	O (VCP)	High side MOSFET driver output
Exposed die pad	-	GND	Connect the exposed die pad to a GND plane. It is used for cooling of the IC and may either be left open or be connected to GND.

5.2 MOSFET Driver Stage

The TMC603 provides three half bridge drivers, each capable of driving two MOSFET transistors, one for the high-side and one for the low-side. In order to provide a low on-resistance, the MOSFET gate driving voltage is about 10V to 12V.

The TMC603 bridge drivers provide a number of unique features for simple operation, explained in the following chapters:

- An integrated automatic break-before-make logic safely switches off one transistor before its counterpart can be switched on.
- Slope controlled operation allows adaptation of the driver strength to the desired slope and to the chosen transistors.
- The drivers protect the bridge actively against cross conduction (Q_{GD} protection)
- The bridge is protected against a short to GND

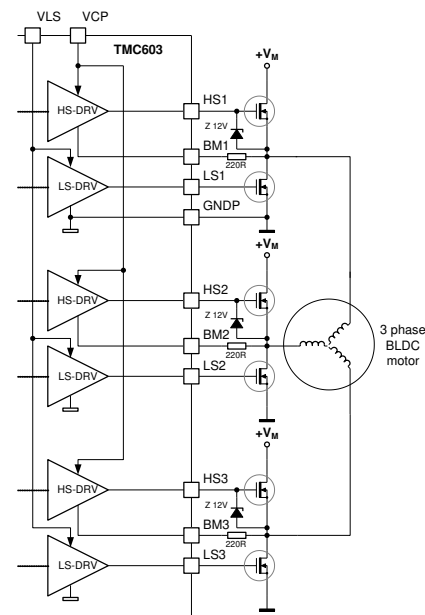


figure 4: three phase BLDC driver

5.2.1 Principle of operation

The low side gate driver voltage is supplied by the VLS pins. The low side driver supplies 0V to the MOSFET gate to close the MOSFET, and VLS to open it.

The TMC603 uses the following driver principle for driving of the high side (pat. fil.):

The high-side MOSFET gate voltage is referenced to its source at the center of the half bridge. Due to this, the TMC603 references the gate drive to the bridge center (BM) and has to be able to drive it to a voltage lying above the positive bridge power supply voltage V_M . This is realized by a charge pump voltage generated from the switching regulator via a Villard circuit. When closing the high-side MOSFET, the high-side driver drives it down to the actual BM potential, since an external induction current from the motor coil could force the output to stay at high potential. This is accomplished by a feedback loop and transistor TG1 (see figure). In order to avoid floating of the output BM, a low current is still fed into the HS output via transistor TG1a. The input BM helps the high side driver to track the bridge voltage. Since input pins of the TMC603 must not go below -0.7V, the input BM needs to be protected by an external resistor. The resistor limits the current into BM to a level, the ESD protection input diodes can accept.

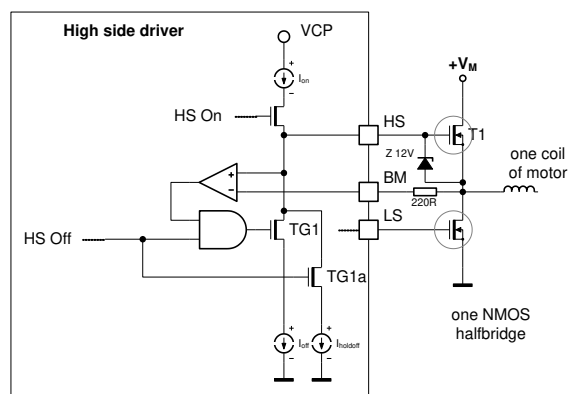


figure 5: principle of high-side driver (pat. fil.)

A zener diode at the gate (range 12V to 15V) protects the high-side MOSFET in case of a short to GND event: Should the bridge be shorted, the gate driver output is forced to stay at a maximum of the zener voltage above the source of the transistor. Further it prevents the gate voltage from dropping below source level.

The maximum permissible MOSFET driver current depends on the motor supply voltage:

Parameter	Symbol	Max	Unit
MOSFET driver current with $V_{VM} < 30V$	I_{HSX}, I_{LSX}	150	mA
MOSFET driver current with $30V < V_{VM} < 50V$		$150 - 2.5 * (V_{VM} - 30V)$	mA
MOSFET driver current with $V_{VM} = 50V$	I_{HSX}, I_{LSX}	100	mA

Pin	Comments
LSx	Low side MOSFET driver output. The driver current is set by resistor R_{SLP} . A Schottky protection diode to GND may be required for MOSFETs, where Q_{GD} is larger than Q_{GS} . Check that LSx voltage does not drop below GND by more than 0.5V.
HSx	High side MOSFET driver output. The driver current is set by resistor R_{SLP}
BMx	Bridge center used for current sensing and for control of the high side driver. For unused bridges, connect BMx pin to GND and leave the driver outputs unconnected. Place the external protection resistor near the IC pin.
RSLP	The resistor connected to this pin controls the MOSFET gate driver current. A 40 μ A current out of this pin (resistor value of 100k Ω to GND) results in the nominal maximum current at full supply range. Keep interconnection between IC and resistor short, to avoid stray capacitance to adjacent signal traces of modulating the set current. Resistor range: 60 k Ω to 500 k Ω
VLS	Low side driver supply voltage for driving low side gates
VCP	Charge pump supply voltage. Provides high side driver supply
GNDP	Power GND for MOSFET drivers, connect directly to GND
BHx	High side driver control signal: A positive level switches on the high side. For unused bridges, tie to GND.
BLx	Low side driver control signal: Polarity can be reversed via INV_BL
INV_BL	Allows inversion of BLx input active level (low: BLx is active high). When high, each BLx and BHx can be connected in parallel in order to use only 3 PWM outputs for bridge control. Be sure to switch on internal break-before-make logic (BBM_EN = Vcc) to avoid bridge short circuits in this case.

5.2.2 Break-before-make logic

Each half-bridge has to be protected against cross conduction during switching events. When switching off the low-side MOSFET, its gate first needs to be discharged, before the high side MOSFET is allowed to be switched on. The same goes when switching off the high-side MOSFET and switching on the low-side MOSFET. The time for charging and discharging of the MOSFET gates depends on the MOSFET gate charge and the driver current set by R_{SLP} . When the BBM logic is enabled, the TMC603 measures the gate voltage and automatically delays switching on of the opposite bridge transistor, until its counterpart is discharged. The BBM logic also prevents unintentional bridge short circuits, in case both, LSx and HSx, become switched on. The first active signal has priority.

Alternatively, the required time can be calculated and pre-compensated in the PWM block of the microcontroller driving the TMC603 (external BBM control).

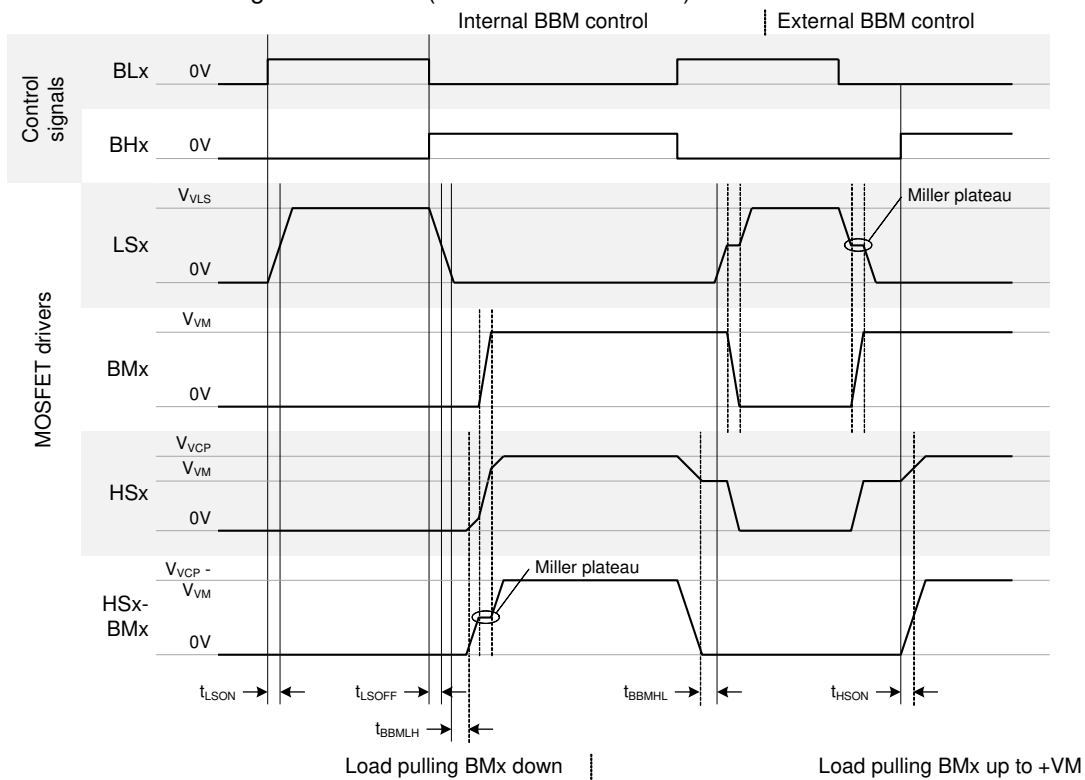


figure 6: bridge driver timing

Pin	Comments
BBM_EN	Enables internal break-before-make circuitry (high = enable)

5.2.3 PWM control via microcontroller

There are a number of different microcontrollers available, which provide specific BLDC commutation units. However, the TMC603 is designed in a way in order to allow BLDC control via standard microcontrollers, which have only a limited number of (free) PWM units. The following figure shows several possibilities to control the BLDC motor with different types of microcontrollers, and shall help to optimally adapt the TMC603 control interface to the features of your microcontroller. The hall signals and further signals, like CURx interconnection to an ADC input, are not shown.

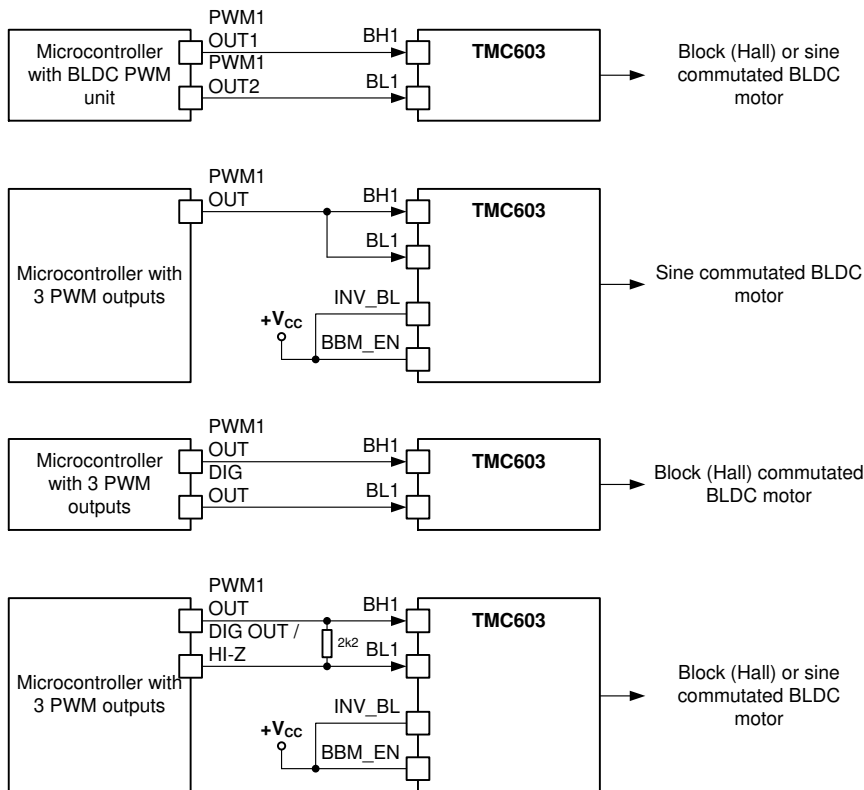


figure 7: examples for microcontroller PWM control

5.2.4 Slope control

The TMC603 driver stage provides a constant current output stage slope control. This allows to adapt driver strength to the drive requirements of the power MOSFET and to adjust the output slope by providing for a controlled gate charge and discharge. A slower slope causes less electromagnetic emission, but at the same time power dissipation of the power transistors rises. The duration of the complete switching event depends on the total gate charge. The voltage transition of the output takes place during the so called miller plateau (see figure 6). The miller plateau results from the gate to drain capacity of the MOSFET charging / discharging during the switching. From the datasheet of the transistor (see example in figure 8) it can be seen, that the miller plateau typically covers only a part (e.g. one quarter) of the complete charging event. The gate voltage level, where the miller plateau starts, depends on the gate threshold voltage of the transistor and on the actual load current.

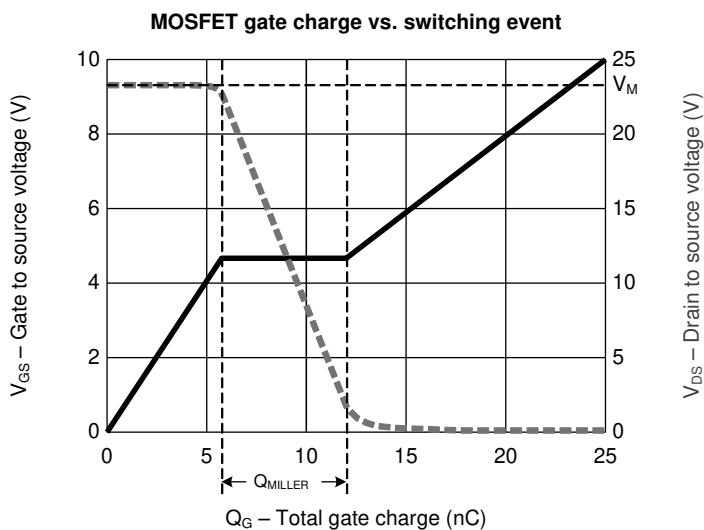


figure 8: MOSFET gate charge as available in device data sheet vs. switching event

The slope time t_{SLOPE} can be calculated as follows:

$$t_{SLOPE} = \frac{Q_{MILLER}}{I_{GATE}}$$

Whereas Q_{MILLER} is the charge the power transistor needs for the switching event, and I_{GATE} is the driver current setting of the TMC603.

Taking into account, that a slow switching event means high power dissipation during switching, and, on the other side a fast switching event can cause EMV problems, the desired slope will be in some ratio to the switching (chopper) frequency of the system. The chopper frequency is typically slightly outside the audible range, i.e. 18 kHz to 40 kHz. The lower limit for the slope is dictated by the reverse recovery time of the MOSFET internal diodes, unless additional Schottky diodes are used in parallel to the MOSFETs source-drain diode. Thus, for most applications a switching time between 100ns and 750ns is chosen.

The required slope control resistor R_{SLP} can be calculated as follows:

$$I_{GATE} = \frac{4V}{R_{SLP}} * \frac{100mA}{40\mu A} \leftrightarrow$$

$$R_{SLP} = \frac{10A * t_{SLOPE}}{Q_{MILLER}} k\Omega$$

Example:

A circuit using the transistor from the diagram above shall be designed for a slope time of 200ns. The miller charge of the transistor is about 6nC.

$$R_{SLP} = \frac{10A * 200ns}{6nC} k\Omega = 333k\Omega$$

The nearest available resistor value is 330 kΩ. It sets the gate driver current to roughly 30mA. This is well within the minimum and maximum R_{SLP} resistor limits.

5.2.5 Reverse capacity (QGD) protection

The principle of slope control often is realized by gate series resistors with competitor's products, but, as latest MOSFET generations have a fairly high gate-drain charge (Q_{GD}), this approach is critical for safe bridge operation. If the gate is not held in the off state with a low resistance, a sudden raise of the voltage at the drain (e.g. when switching on the complementary transistor) could cause the gate to be pulled high via the MOSFETs gate drain capacitance. This would switch on the transistor and lead to a bridge short circuit.

The TMC603 provides for safe and reliable slope controlled operation by switching on a low resistance gate protection transistor (see figure).

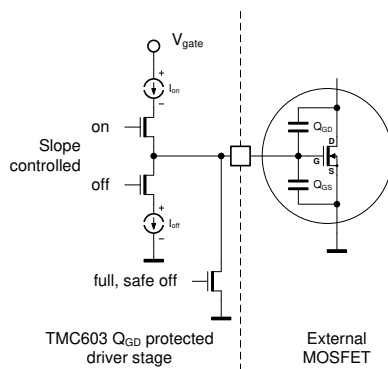


figure 9: QGD protected driver stage

5.2.6 Considerations for QGD protection

This chapter gives the background understanding to ensure a safe operation for MOSFETs with a gate-drain (Miller) charge Q_{GD} substantially larger than the gate-source charge Q_{GS} .

In order to guarantee a safe operation of the Q_{GD} protection, it is important to spend a few thoughts on the slope control setting. Please check your transistors' data sheet for the gate-source charge Q_{GS} and the gate-drain charge Q_{GD} (Miller charge). In order to turn on the MOSFET, first the gate-source charge needs to be charged to the transistor's gate. Now, the transistor conducts and switching starts. During the switching event, the additional Q_{GD} needs to be charged to the gate in order to complete the switching event. Wherever Q_{GD} is larger than Q_{GS} , a switching event of the complementary MOSFET may force the gate of the switched off MOSFET to a voltage above the gate threshold voltage. For these MOSFETs the Q_{GD} protection ensures a reliable operation, as long as the slopes are not set too fast.

Calculating the maximum slope setting for high Q_{GD} MOSFETs:

Taking into account effects of the MOSFET bulk diode (compare chapter 5.2.7), the maximum slope of a MOSFET bridge will be around the double slope as calculated from the Miller charge and the gate current. Based on this, we can estimate the current required to hold the MOSFET safely switched off:

During the bridge switching period, the driver must be able to discharge the difference of Q_{GD} and Q_{GS} while maintaining a gate voltage below the threshold voltage.

Therefore

$$I_{OFFQGD} * \frac{t_{SLOPE}}{2} > Q_{GD} - Q_{GS} \rightarrow$$

$$I_{OFFQGD} * \frac{\frac{Q_{GD}}{I_{ON}}}{2} > Q_{GD} - Q_{GS}$$

Thus the minimum value required for I_{OFFQGD} can be calculated:

$$I_{OFFQGD} = I_{ON} * \frac{Q_{GD} - Q_{GS}}{Q_{GD}} * 2$$

Where I_{ON} is the gate current set via R_{SLP} , and I_{OFFQGD} is the Q_{GD} protection gate current.

The low side driver has a lower Q_{GD} protection current capability than the high side driver, thus we need to check the low side. With its $R_{LSOFFQGD}$ of roughly 15 Ohm, the TMC603 can keep the gate voltage to a level of:

$$U_{GOFF} = I_{OFF} * R_{LSOFFQGD}$$

Now we just need to check U_{GOFF} against the MOSFETs output characteristics, to make sure, that no significant amount of drain current can flow.

Example:

A MOSFET, where Q_{GD} is 3 times larger than Q_{GS} is driven with 100mA gate current.

$$I_{OFF} = 100mA * \frac{3Q_{GS} - Q_{GS}}{3Q_{GS}} * 2 = 133mA$$

The TMC603 thus can keep the gate voltage level to a maximum voltage of $U_{GOFF} = 133mA * 15\Omega = 2V$

This is sufficient to keep the MOSFET safely off.

Note:

Do *not* add gate series resistors to your MOSFETs! This would eliminate the effect of the Q_{GD} protection. Gate series resistors of a few Ohms only may make sense, when paralleling multiple MOSFETs in order to avoid parasitic oscillations due to interconnection inductivities.

5.2.7 Effects of the MOSFET bulk diode

Whenever inductive loads are driven, the inductivity will try to sustain current when current becomes switched off. During bridge switching events, it is important to ensure break-before-make operation, e.g. one MOSFET becomes switches off, before the opposite MOSFET is switched on. Depending on the actual direction of the current, this results in a short moment of a few 100 nanoseconds, where the current flowing through the inductive load forces the bridge output below the lower supply rail or above the upper supply rail. The respective MOSFET bulk diode in this case takes over the current. The diode saturates at about -1.2V. But the bulk diode is not an optimum device. It typically has reverse recovery time of a few ten to several 100ns and a reverse recovery charge in the range of some 100nC or more. Assuming, that the bulk diode of the switching off MOSFET takes over the current, the complementary MOSFET sees the sum of the coil current and the instantaneous current needed to recover the bulk diode when trying to switch on. The reverse recovery current may even be higher than the coil current itself! As a result, a number of very quick oscillations on the output appear, whenever the bulk diode leaves the reverse recovery area, because up to the half load current becomes switched off in a short moment. The effect becomes visible as an oscillation due to the parasitic inductivities of the PCB traces and interconnections. While this is normal, it adds high current spikes, some amount of dynamic power dissipation and high frequency electromagnetic emission. Due to its high frequency, the ringing of this current can also be seen on the gate drives and thus can be easily mistaken as a gate driving problem. In order to reduce overshoot and ringing, a snubber element can be used, e.g. a capacitor with some nano Farad in series with a resistor in the range some 100mΩ on each motor output.

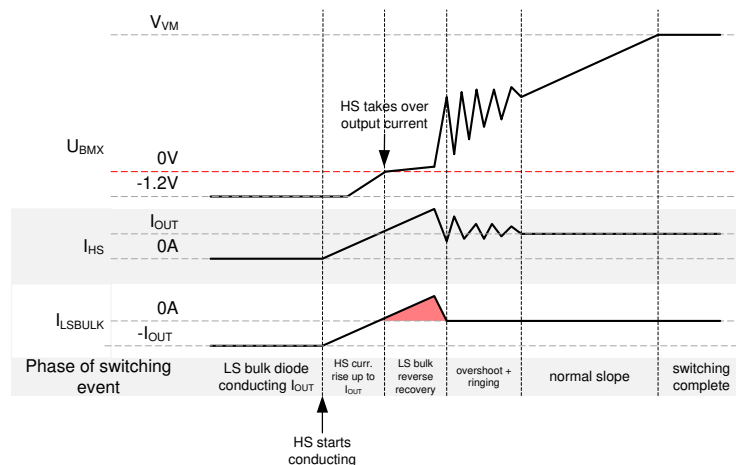


figure 10: effect of bulk diode recovery

A further conclusion from this discussion: Do *not* set the bridge slope time higher than or near to the reverse recovery time of the MOSFETs, as the parasitic current spikes will multiply the instantaneous current across the bridge. A plausible time is a factor of three or more for the slope time. If this cannot be tolerated please see the discussion on adding Schottky diodes.

5.2.8 Adding Schottky diodes across the MOSFET bulk diodes

In order to avoid effects of bulk diode reverse recovery, choose a fast recovery switching MOSFET. The MOSFET transistors can also be bridged by a Schottky diode, which has a substantially faster reverse recovery time. This Schottky diode needs to be chosen in a way that it can take over the full bridge current for a short moment of time only. During this time, the forward voltage needs to be lower than the MOSFETs forward voltage. A small 5A diode like the SK56 can take over a current of 20A at a forward voltage of roughly 0.8V. Even in this constellation, an optional snubber element at the output can reduce overshoot and ringing (see schematic).

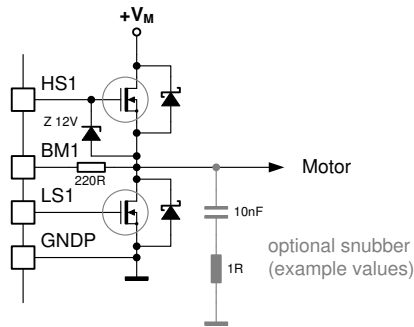


figure 11: parallel Schottky diode avoids current spikes due to bulk diode recovery, optional snubber reduces overshoot and ringing

5.2.9 Short to GND detection

An overload condition of the high side MOSFET (“short to GND”) is detected by the TMC603, by monitoring the BM voltage during high side on time. Under normal conditions, the high side power MOSFET reaches the bridge supply voltage minus a small voltage drop during on time. If the bridge is overloaded, the voltage cannot rise to the detection level within a limited time, defined by an external resistor. Upon detection of an error, the error output is activated. By directly tying it to the enable input, the chip becomes disabled upon detection of a short condition and the error flip flop becomes set.

A variation of the short to GND detection delay allows adaptation to the slope control, as well as modification of the sensitivity of the short to GND detection.

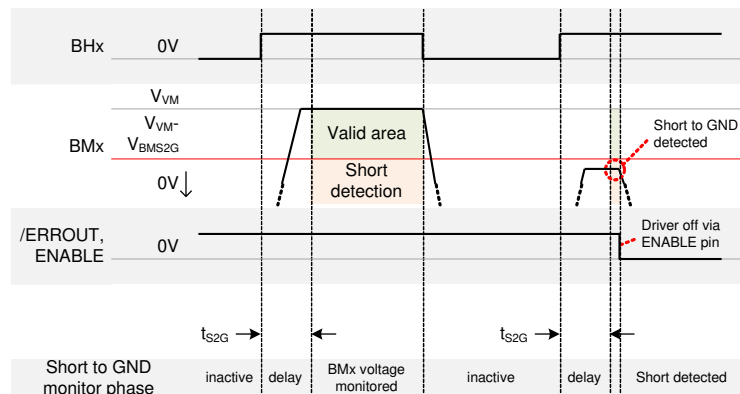


figure 12: timing of the short to GND detector

Pin	Comments
RS2G	The resistor connected to this pin controls the delay between switching on the high side MOSFET and the short to GND check. A 20µA current out of this pin (resistor value of 220 kΩ to GND) results in a 500ns delay, a lower current gives a longer delay. Disconnecting the pin disables the function. Keep interconnection between IC and resistor short, to avoid stray capacitance to adjacent signal traces of modulating the set current. Resistor range: 47 kΩ to 1 MΩ

5.2.10 Error logic

The TMC603 has three different sources for signaling an error:

- Undervoltage of the low side supply
- Undervoltage of the charge pump
- Short to GND detector

Upon any of these events the error output is pulled low. After a short to GND detector event, the error output remains active, until it becomes cleared by the CLR_ERR. By tying the error output to the

enable input, the TMC603 automatically switches off the bridges upon an error. The enable input then should be driven via an open collector input plus pull-up resistor, or via a resistor.

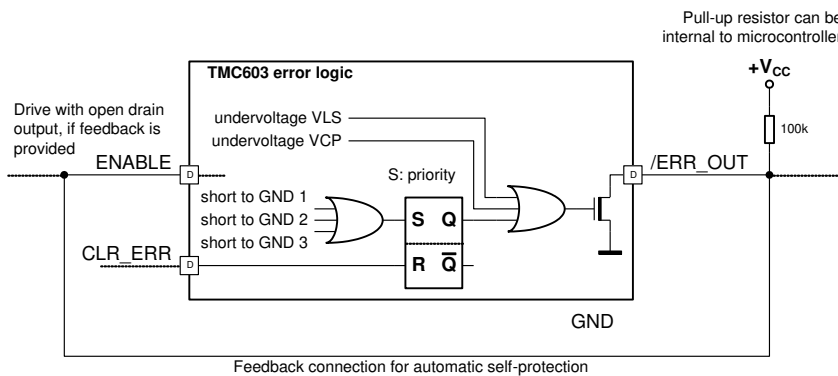


figure 13: error logic

Pin	Comments
/ERR_OUT	Error output (open drain). Signals undervoltage or overcurrent. Tie to ENABLE for direct self-protection of the driver. The internal error condition generator has a higher priority than the CLR_ERR input, i.e. the error condition cannot be cleared, as long as it is persistent.
CLR_ERR	Reset of error flip-flop (active high). Clears error condition. The error condition should at least be cleared once after IC power on.
ENABLE	Enables the power drivers (low: all MOSFETs become actively switched off)

5.2.11 Paralleling gate drivers for higher gate current

In order to double gate driver current in a BLDC application, two TMC603 can be switched in parallel to have the double gate driver current while taking advantage of all features. Therefore it is important to parallel the gate driver inputs and outputs of the second IC to the first IC, and to also parallel the ERR_OUT and ENABLE input. The driver strength of both ICs adds taking into account their respective slope control resistor. The switching regulator and charge pump of one device can supply both ICs!

5.3 Current measurement amplifiers

The TMC603 amplifies the voltage drop in the three lower MOSFET transistors in order to allow current measurement without the requirement for current sense (shunt) resistors. This saves cost and board space, as well as the additional power dissipation in the shunt resistors. Optional shunt resistors can be used, e.g. source resistors for each lower MOSFET or a common shunt resistor in the bridge foot point if a more precise measurement without the need for calibration and temperature compensation is desired. For the TMC603A, the FILTx pins in this mode are switched as inputs for the sensing of the shunt resistors. The internal amplifier conditions the signal for a standard microcontroller.

The TMC603 CURx outputs deliver a signal centered to 1/3 of the 5V VCC supply. This allows measurement of both, negative and positive signals, while staying compatible to a 3.3V microcontroller. The current amplifier is an inverting type.

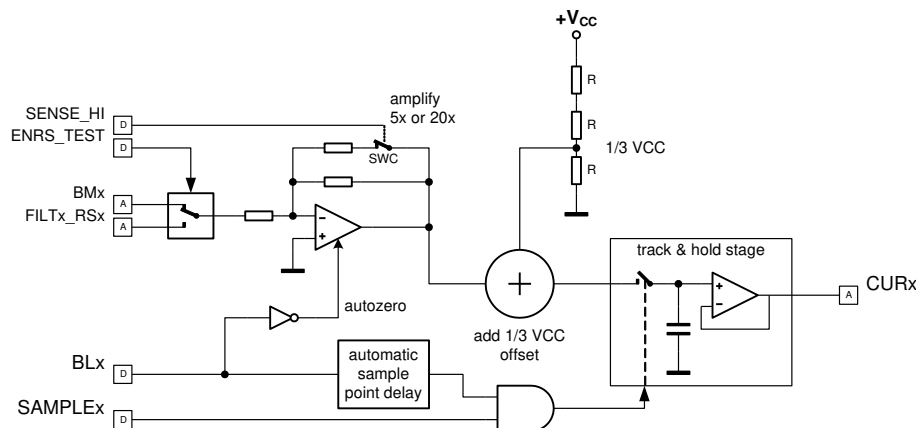


figure 14: schematic of current measurement amplifiers

Pin	Comments
CURx	Output of current measurement amplifier. The output signal is centered to 1/3 VCC.
SENSE_HI	Switches current amplifiers to high sensitivity (high level). Control by processor to get best sensitivity and resolution for measurement.
SAMPLEx	Optional external control for current measurement sample/hold stage. Set to positive level, if unused
FILTx_RSx	Input for optional external sense resistor. To enable, tie pin ENRS_TEST to VCC. This feature has been added in TMC603A.

The voltage drop over the MOSFET (or shunt resistor) is calculated as follows:

$$V_{DROP} = \frac{x_0 - x}{ADC_{MAX}} * V_{ADCREf} / A_{CUR}$$

whereas x is the ADC output value, x_0 is the ADC output value at zero current (e.g. 85 for an 8 bit ADC with 5V reference voltage), ADC_{MAX} is the range of the ADC (e.g. 256 for an 8 bit ADC), V_{ADCREf} is the reference voltage of the ADC and A_{CUR} is the currently selected amplification (absolute value) of the TMC603.

With this, the motor current can be calculated using the on resistance R_{DSON} (at 10V) of the MOSFET:

$$I_{MOSFET} = \frac{V_{DROP}}{R_{DSON}}$$

For a shunt resistor based measurement, the same formula is true:

$$I_{SHUNT} = \frac{V_{DROP}}{R_{SHUNT}}$$

For the shunt resistor measurement, care has to be taken not to exceed the voltage range which can be accepted by the measurement input, i.e. the shunt resistor should be selected in a way that the voltage drop is at maximum 0.3V at full motor current. This is the maximum voltage which can be measured. A lower sense resistor gives less power dissipation, but lower currents show with less resolution.

5.3.1 Current measurement timing

Current measurement is self-timed, in order to only provide valid output voltages. Sampling is active during the low side ON time. The sampling is delayed by an internal time delay, in order to avoid sampling of instable values during settling time of the bridge current and amplifiers. Thus, a minimum ON time is required in order to get a current measurement. The output CURx reflects the current during the measurement. The last value is held in a track and hold circuit as soon as the low side transistor switches off.

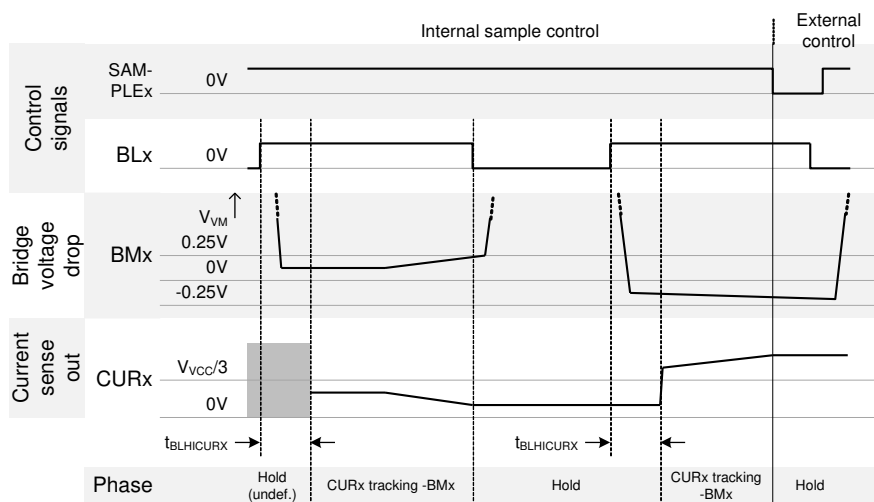


figure 15: timing of the current measurement

The SAMPLEx pins can be used to refresh the measurement during long on time periods, e.g. when the motor is in standstill, with the low side being continuously on, e.g. in a hall sensor based block commutation scheme with the chopper on the high side. In this application, all SAMPLEx pins can be tied together to one microprocessor output. For advanced applications, where a precise setting of the current sampling point is desired, e.g. centered to the on-time, SAMPLEx pins can be deactivated at the desired point of time, enabling the hold stage.

5.3.2 Auto zero cycle

The current measurement amplifiers do an automatic zero cycle during the OFF time of the low side MOSFETs. The zero offset is stored in internal capacitors. This requires switching off the low side at least once, before the first measurement is possible, and on a cyclic basis, to avoid drifting away of the zero reference. This normally is satisfied by the chopper cycle. If commutation becomes stopped, e.g. due to motor stand still, the respective phase current measurement could drift away. After the first switching off and on of the low side, the measurement becomes valid again. Therefore, you should integrate a timer in your commutation, which checks for the low side on time exceeding for example 10ms. If the on time of the respective low side reaches this time limit, you can either use the sample input SAMPLEx to refresh the current measurement, by switching it high for at least 1μs, or you switch off the low side for a short time of a few microseconds.

5.3.3 Measurement depending on chopper cycle

If the low side on-time on one phase $t_{BLHICURX}$ is too short, a current measurement is not possible. The TMC603 automatically does not sample the current if the minimum low side-on time requirement is not met. This condition can arise in normal operation, e.g. due to the commutation angle defined by a sine commutation chopper scheme. The respective CURx output then does not reflect the phase current. Thus, the CURx output of a phase should be ignored, if the on-time falls below the minimum low side on-time for current measurement (please refer to maximum limit). The correct current value can easily be calculated using the difference of the remaining two current measurements. This results from the fact that the sum of all three currents equals zero ($I_U + I_V + I_W = 0$). This way, all motor currents are always known from the measurement of two phase currents. It is important to know all three phase currents for a sine commutated motor. For block commutation, there is always one low side active and the full current can be seen at this low side.

5.3.4 Compensating for offset voltages

In order to measure low current values precisely, the “zero” value (x_0) of $1/3 V_{CC}$ should be measured via the ADC, rather than being hard coded into the measurement software. This is possible by doing a first current measurement during motor stand-still, with no current flowing in the motor coils, e.g. during a test phase of the unit. The resulting value can be stored and used as zero reference. However, the influence of offset voltages can be minimized, by using the high sensitivity setting of the amplifiers for low currents, and switching to low sensitivity for higher currents.

5.3.5 Getting a precise current value using MOSFET on-resistance

The on-resistance of a MOSFET has a temperature co-efficient, which should not be ignored. Thus, the temperature of the MOSFETs must be measured, e.g. using an NTC resistor, in order to compensate for the variation. Also, the initial $R_{DS(ON)}$ depends upon fabrication tolerance of the MOSFETs. If exact measurement is desired, an adjustment should be done during initial testing of each product. For applications, where an adjustment is not possible, external sense resistors can be used instead. A single resistor in the GND line often is sufficient for block commutation. For sine commutation, three sense resistors should be used.

5.4 Power supply

The TMC603 integrates a +12V switching regulator for the gate driver supply and a +5V linear regulator for supply of the low voltage circuitry. The switching regulator is designed in a way, that it provides the charge pump voltage by using a Villard voltage doubler circuit. It is able to provide enough current to supply a number of digital circuits by adding an additional 3.3V or 5V low voltage linear or switching regulator. If a +5V microcontroller with low current requirement is used, the +5V regulator is sufficient, to also supply the microcontroller.

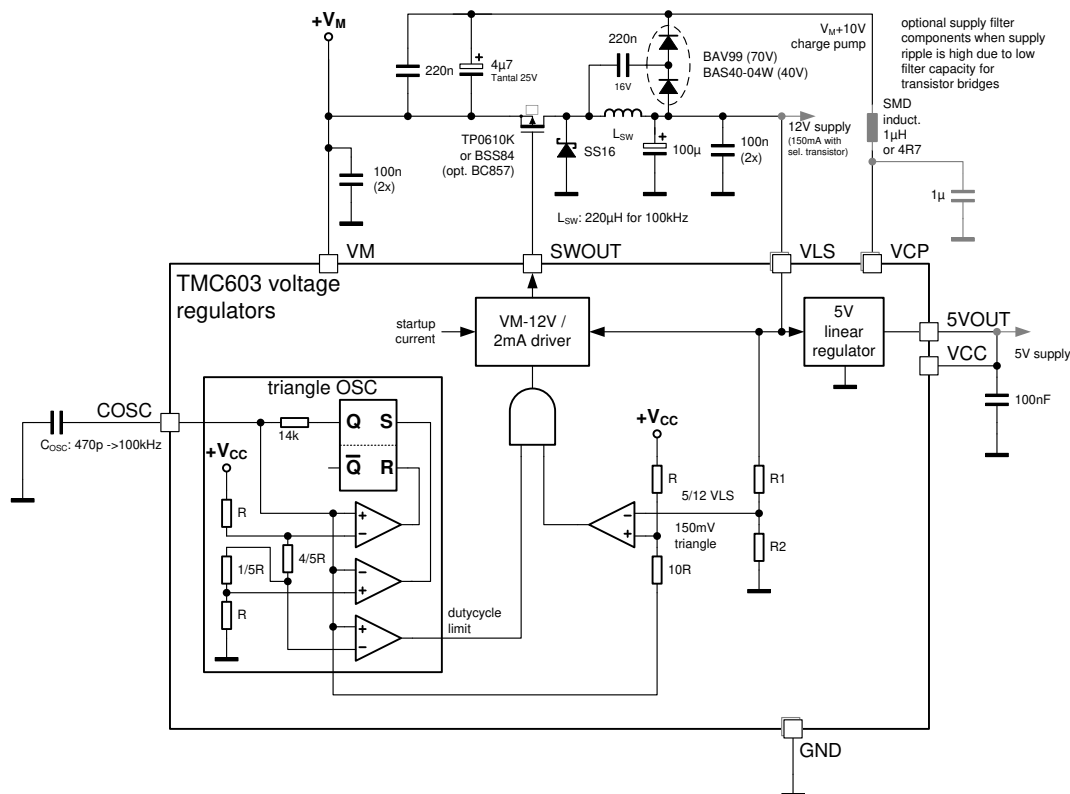


figure 16: power supply block with example values

Pin	Comments
COSC	Oscillator capacitor for step down regulator. A 470pF capacity gives 100kHz operation. Do not leave this pin unconnected. Tie to GND, if oscillator is not used.
SWOUT	Switch regulator transistor output. The output allows driving of a small signal P-channel MOSFETs as well as PNP small signal transistors
5VOUT	Output of internal 5V linear regulator. Provided for VCC supply

5.4.1 Switching regulator

The switching regulator has been designed for high stability. It provides an upper duty cycle limit, in order to ensure switching operation even at low supply voltage. This allows the combination with a Villard voltage doubler. The application schematic shows a number of standard values, however, the coil and oscillator frequency can be altered:

The choice of the external switching regulator transistor depends on the desired load current and the supply voltage. Especially for high switching frequencies, a low gate charge MOSFET is required. The following table shows an overview of available transistors and indicative operation limits. For a higher output current, two transistors can be used in parallel. In this case the switching frequency should be halved, because of the higher gate charge leading to slower switching slopes.

transistor type	manufacturer	gate charge (typ.)	max. frequency	max. voltage	max. load current
BC857	div.	- (bipolar)	100 kHz	40V	80 mA
BSS84	Fairchild, NXP	0.9 nC	300 kHz	50V	120 mA
TP0610K	Vishay	1.3 nC	230 kHz	60V	150 mA
NDS0605	Fairchild	1.8 nC	175 kHz	60V	150 mA
TP0202K	Vishay	1 nC	300 kHz	30V	350 mA

For the catching diode, use a Schottky type with sufficient voltage and current rating.

The choice of a high switching frequency allows the use of a smaller and less expensive inductor as well as a lower capacitance for the Villard circuit and the switching regulator output capacitor. However, the combination of inductor, transistor and switching frequency should be carefully selected and should be adapted to the load current, especially if a high load current is desired.

Choice of capacitor for the switching frequency (examples):

C _{osc}	frequency f _{osc}	inductivity example	Remark
470 pF	100 kHz	220 μH	
220 pF	175 kHz	150 μH	
100 pF	300 kHz	100 μH	Not recommended for V _{VM} < 14V

The switcher inductivity shall be chosen in a way, that it can sustain part of the load current between each two switching events. If the inductivity is too low, the current will drop to zero and higher frequency oscillations for the last part of each cycle will result (discontinuous mode). The required transistor peak current will rise and thus efficiency falls.

For a low load current, operation in discontinuous mode is possible. If a high output current is required, a good design value for continuous mode is to target a current ripple in the coil of +/-40%.

To give a coarse hint on the required inductor you can use the following formula for calculating the minimum inductivity required for continuous operation, based on a ripple current which is 100% of the load current:

$$L_{SW} = \frac{12V}{I_{OUT} * f_{OSC}} * \frac{V_{VM} - 12V}{V_{VM}}$$

V_{VM} is the supply voltage. For low voltage operation (15V or less), the output voltage sinks from 12V to 0.85*V_{VM}. The formula can be adapted accordingly.

I_{OUT} is the current draw at 12V.

For 40% current ripple, you can use roughly the double inductivity.

If ripple is not critical, you can use a much smaller inductivity, e.g. only 5% to 50% of the calculated value. But at the same time switching losses will rise and efficiency and current capability sink due to higher losses in the switching transistor. If the TMC603 does not supply additional external circuitry, current draw is very low, about 20mA in normal operation. This would lead to large inductivity values. In this case we recommend going for the values given in the table above in order to optimize coil cost.

Example:

$$f_{\text{OSC}} = 175 \text{ kHz}, I_{\text{OUT}} = 0.2 \text{ A}, V_{\text{VM}} = 48 \text{ V:}$$

$$L_{\text{SW}} = \frac{12\text{V}}{0.2\text{A} * 175000\text{Hz}} * \frac{48\text{V} - 12\text{V}}{48\text{V}} = 257\mu\text{H}$$

For continuous operation, a 330 μ H or 470 μ H coil would be required. The minimum inductivity would be around 100 μ H.

Note:

Use an inductor, which has a sufficient nominal current rating. Keep switching regulator wiring away from sensitive signals. When using an open core inductor, please pay special care to not disturbing sensitive signals.

5.4.2 Charge pump

The Villard voltage doubler circuit relies on the switching regulator generating a square wave at the switching transistor output with a height corresponding to the supply voltage. In order to work properly the load drawn at +12V needs to be higher than the load drawn at the charge pump voltage. This normally is satisfied when the IC is supplied by the step down regulator. For low voltage operation, the charge pump voltage needs to be as high as possible to guarantee a high gate drive voltage, thus, a dual Schottky diode should be used for the charge pump in low voltage applications.

5.4.3 Filter capacitors for switching regulator and charge pump

The filter capacitors in the switching regulator and the charge pump are required to provide current for the high current spikes which are caused by switching up to three MOSFETs at the same time. The required amount of charge can be estimated when looking at the MOSFETs gate charge. The gate voltage should not drop significantly due to the switching event, e.g. only 100mV. Additionally, the 12V filter capacitor provides charge for load spikes on the 12V net and filter switching ripple. In applications, where board space is critical, lower capacitance values can be used.

Choice of filter capacitors in the switching regulator for different current requirements (example):

12V load current	power MOSFET gate charge	12V filter capacitor (electrolytic/ceramic)	charge pump filter capacitor (tantalum / ceramic)
<20mA	<20nC	22 μ F (or 4.7 μ F ceramic)	1 μ F (e.g. ceramic)
<20mA	<50nC	22 μ F (or 10 μ F ceramic)	2.2 μ F (e.g. ceramic)
<50mA	>50nC	47 μ F (or 10 μ F ceramic)	4.7 μ F
100mA	>50nC	100 μ F (or 10 μ F ceramic)	4.7 μ F

5.4.4 Supply voltage filtering and layout considerations

As with most integrated circuits, ripple on the supply voltage should be minimized in order to guarantee a stable operation and to avoid feedback oscillations via the supply voltages. Therefore, use a ceramic capacitor of 100nF per supply voltage pin (VM to GND, VLS to GND and VCC to GND and VCP to VM). Please pay attention to also keep voltage ripple on VCC pin low, especially when the 5V output is used to supply additional external circuitry. It also is important to make sure, that the resistance of the power supply is low when compared to the load circuit. Especially high frequency voltage ripple >1MHz should be suppressed using filter capacitors near the power bridge or near the board power supply. The VM terminal is used, to detect short to GND situations, thus, it has to correspond to the bridge power supply. In high noise applications, it may make sense to filter VCP supply separately against ripple to GND. A large low ESR electrolytic capacitor across the bridge supply (VM to GND) should also be used, because it effectively suppresses high frequency ripple. This cannot be accomplished with ceramic capacitors. GND and GNDP pins should be tied to a common, massive GND plane. Pay attention to power routing: Use short and wide, straight traces. The PCB power supply should be placed near the driver bridge, where most current is consumed, to avoid current drop in the plane between critical components like TMC603 and microcontroller. This is especially important to get a precise current measurement.

5.4.5 Reverse polarity protection

Some applications need to be protected against a reversed biased power supply, i.e. for automotive applications. A highly efficient reverse polarity protection based on an N channel MOSFET can simply be added due to the availability of a charge pump voltage. This type of reverse polarity protection allows feeding back energy into the supply, and thus is preferable to a pure diode reverse polarity protection.

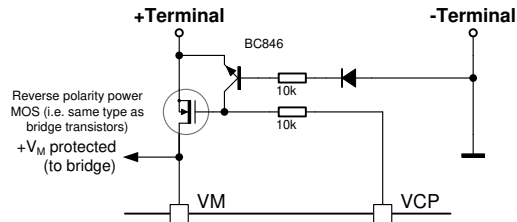


figure 17: adding a reverse polarity protection

5.4.6 Standby with zero power consumption

In battery powered applications, a standby function often is desired. It allows switching the unit on or off without the need for a mechanical high power switch. In principle, the bridge driver MOSFETs can switch off the motor completely, but the TMC603 and its switching regulator still need to be switched off in order to reduce current consumption to zero. Only a low energy standby power supply will remain on, in order to wake up the system controller. This standby power supply can be generated by a low current zener diode plus a resistor to the battery voltage, buffered by a capacitor. The example in the schematic uses a P channel MOSFET to switch off power for the TMC603 and any additional ICs which are directly supplied by the battery. Before entering standby mode, the motor shall be stopped and the TMC603 should be disabled.

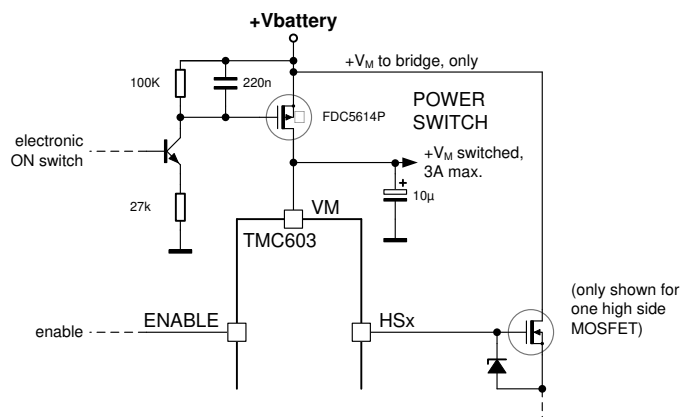


figure 18: low power standby

5.4.7 Low voltage operation down to 9V

In low voltage operation, it is important to keep the gate driving voltages as high as possible. The switching regulator for VLS thus is not needed and can be left out. Tie the pin COSC to GND. VLS becomes directly tied to +VM, which is possible as long as the supply voltage does not exceed 14V (16V peak). However, now a source for the Villard voltage doubler is missing. A simple solution is to use a CMOS 555 timer circuit (e.g. TLC555) oscillating at 250 kHz (square wave) to drive the voltage doubler.

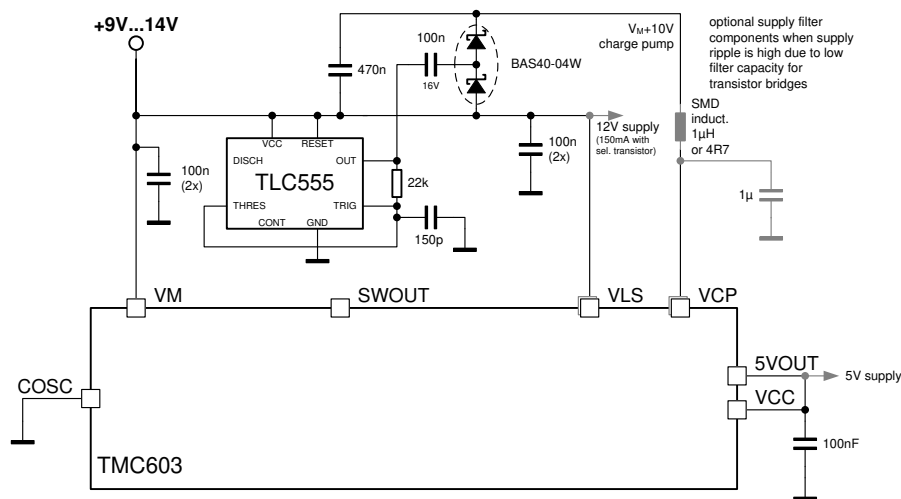


figure 19: low voltage operation

5.5 Test output

The test output is reserved for manufacturing test. It is used as an input for a normal application. Tie to GND or VCC in application.

Pin	Comments
ENRS_TEST	<p>Enable sense resistor input and output for test voltages. Output resistance 25kOhm \pm30%. Reset: ENABLE(=low); Clock: SCCLK (rising edge). Test voltage sequence: 0: 0V 1..3 / 4..6 / 7..9: Gate_HS_Off, Gate_LS_On, Gate_LS_Off (driver 1/2/3) 10..14: currently unused 15: 0V (no further counts: Reset for restart)</p>

5.6 ESD sensitive device

The TMC603 is an ESD sensitive CMOS device and also MOSFET transistors used in the application schematic are very sensitive to electrostatic discharge. Take special care to use adequate grounding of personnel and machines in manual handling. After soldering the devices to the board, ESD requirements are more relaxed. Failure to do so can result in defect or decreased reliability.

