L99MH98 indirect current sense method

Introduction

The L99MH98 is a versatile octal half-bridge predriver able to control up to sixteen N-channel MOSFETs. The device is intended for DC motor control applications such as automotive power seat control, sunroof, window lift, and most generic DC motor drive applications. Each half-bridge driver independently monitors its external MOSFET drain-source voltage to provide overload and short circuit protection.

A new and innovative method, different from the traditional voltage drop on the shunt resistors, has been designed for L99MH98 to support indirect current measurement of external MOSFETs. This method can calculate the current value by reading any external MOSFET V_{ds}, delivered via the CSO pins and estimating $R_{ds(on)}$ by measuring the temperature of the same MOSFET.

To achieve these measurements, the sensing of an external diode or diodes chains is provided by the L99MH98. These diodes, placed as close as possible to the external switches, may also help sense and monitor the temperature of specific PCB areas on top of indirect current sensing.

1 Indirect current sense method

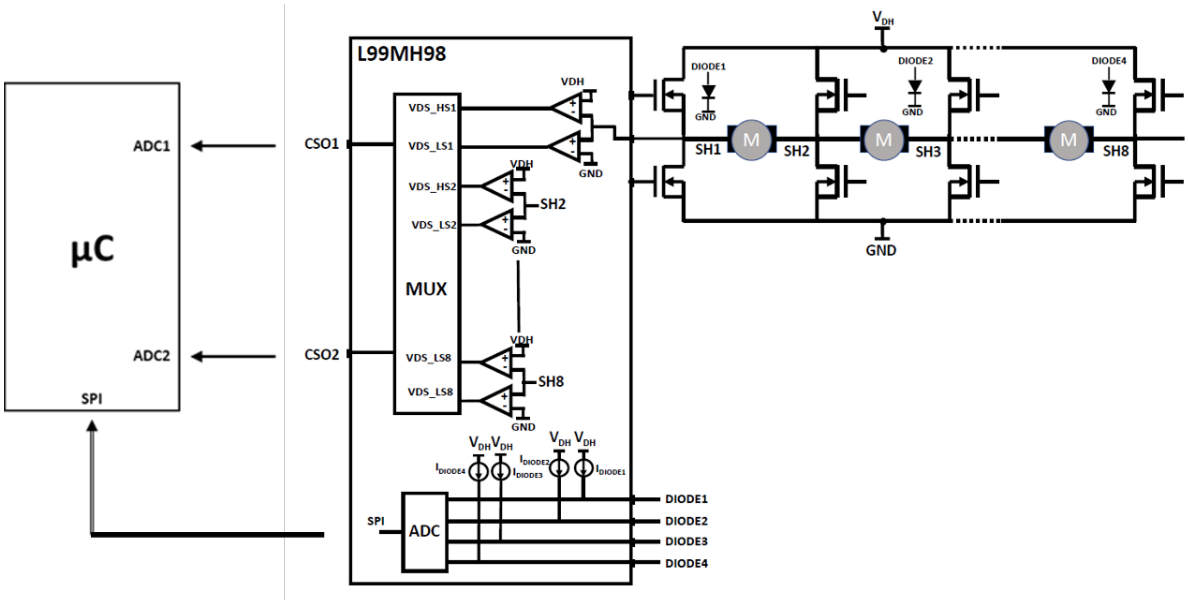
The proposed indirect method is based on the calculation of the current flowing through a MOSFET as a ratio between the $\mathbf{V_{ds}}$ across it and deriving its $\mathbf{R_{ds(on)}}$ from the temperature of the same MOSFET.

This feature demonstrates the advantage of not using any external sensing resistors, allowing a lower cost of ownership and no voltage drop on the sensing shunt, hence reducing power dissipation.

The new current sense method consists of the following different phases:

1. **$\mathbf{CSO_X}$ setting** according to the adopted external MOSFET and the targeted load current (to be performed once, before working condition)
2. **$\mathbf{R_{ds(on)}}$ calibration** (and temperature) at application level (to be performed once, before working condition)
3. **Real-time $\mathbf{V_{ds}}$ measurement** via CSO pins
4. **Temperature measurement for $\mathbf{R_{ds(on)}}$ estimation** (periodically provided by IC)
5. **Current calculation** (by MCU)

Figure 1. L99MH98 indirect current sense schematic



The schematic above shows the main circuits employed to perform this method:

1. The V_{ds} are read by the uC, provided by the IC as amplified Current Sense Amplifiers' outputs and available to the uC's ADC channels.
2. The diodes, placed close to the driving MOSFETs, provide the information on the temperature values through periodic voltage signals that are read and sampled by the ADC of the device and then sent to the MCU via SPI.

1.1 CSOx setting

This preliminary step of the indirect procedure regards the V_{ds} measurement system setting. To read the V_{ds} voltage across the external transistors, the device uses internal current sense amplifier circuits. In this way, the L99MH98 can reflect the drain to source voltage across each of the external MOSFETs with a gain on one of the two CSO_X pins. This applies to both low- and high-side MOSFETs using an analog multiplexer.

The total gain of the CSO_X consists of the contribution of two different configurable gain stages. The choice of the gain of each stage is very important and must be performed according to:

- The current range involving the load, and depending on the piloting MOSFET's family, the V_{ds} values generated across it (input of the current sense amplifier).
- The desired output voltage range of the CSO_X pins.

To cover these configurations, the specifications of the device have been defined in different compatible ranges. The setting of the first gain stage of the current sense amplifiers has been in fact conceptually associated with two different **input voltage V_{ds} ranges**:

- CSO_{xInR_A} = **10 mV < V_{ds} < 140 mV** (indirect current sense input voltage **Range A**)
- CSO_{xInR_B} = **120 mV < V_{ds} < 450 mV** (indirect current sense input voltage **Range B**)

Below are the settings of the **gains of both CSOx stages**:

1. The gain of the first stage is related to the expected input voltage range and can be set by defining the **drain-source monitoring thresholds** of the used MOSFETs (configured by `VDS_CONFx` bits, as can be seen in the datasheet of the device). In fact, by setting the drain-source monitoring threshold to 75 mV (`VDS_CONFx` = `0000`) or 150 mV (`VDS_CONFx` = `0001`), the gain of the first stage will be set to **10 (V/V).** This high gain is expected to be used with low V_{ds} voltages, compatible with values in Range A.
2. For drain-source monitoring thresholds from 200 mV (`VDS_CONFx` = `0010`) to 600 mV (`VDS_CONFx` = `0111`), or even 2 V (`VDS_CONFx` = `1xxx`), however, the gain of the first stage is set to **2.5 (V/V)**. This lower gain is expected to be used with higher V_{ds} voltages, compatible with values in Range B.
3. The second stage gain, on the other hand, is controlled by a bit for each CSO channel, `CSO_GAIN_SELx` registers. When `CSO_GAIN_SELx` = `0` the second gain set is **1.5 (V/V)**, while when `CSO_GAIN_SELx` = `1` is **3 (V/V)**.

The total gain is therefore given by the product of the two selected gains. The results of the possible combinations of the two gains are: **3.75** and **7.5** or **15** and **30**. The 15 (V/V) and 30 (V/V) gains are used with Range A inputs. The 7.5 (V/V) is the one adopted with Range B, while the 3.75 (V/V) combination is generally not used.

Moreover, from datasheet specification, for each input range (and associated gain combinations) a compatible output voltage range is delivered by the CSO_x channels:

- CSO_{xOutR_A} = **0.1 V < CSO_x < V_{DD} -0.3 V** (indirect current sense output voltage **Range A**)
- CSO_{xOutR_B} = **0.3 V < CSO_x < V_{DD} -0.3 V** (indirect current sense output voltage **Range B**)

Since the validity of CSO_x voltage is granted if its value is in the output range specified, it is suggested not only to set the gains compatible with load's produced V_{ds} (input) but also to choose these gains to obtain a compatible CSO_x output value (approximately in the middle of its range, depending also by V_{DD}).

Example: CSO settings

All these considerations can be evaluated in detail with an example. In fact, selecting the MOSFET STD12NF06LAG from STMicroelectronics, it can be noted that its typical $R_{ds(on)}$ value is 70 mΩ @ room temperature. If these MOSFETs are adopted to pilot a motor with an average current of 2 A, the average V_{ds} generated by this current flowing through the $R_{ds(on)}$ of the transistor will be:

$$V_{ds\_AVG} = I_{AVG} \cdot R_{ds(on)typ} = 2A \cdot 0.07\Omega = 140mV$$

This value seems to be suited for both voltage input ranges, but additional consideration must be made. It can be added at first that the value of $R_{ds(on)}$ can double in temperature (0.14 Ω), and by that, rising the expected V_{ds}. Moreover, since the motor can be piloted by a PWM signal, the periodic peak current of the motor will be higher than its average current, even double.

This reasoning can be applied both to set the gain of the first stage and the drain-source monitoring thresholds. It is suggested to consider this higher current value with a proper additional margin to evaluate and set a safer and more correct monitoring threshold. In this specific case, a threshold of 600 mV or even 2 V could be a good choice, since:

$$\mathrm{V_{ds\_MAX}} = \mathrm{I_{MAX}} \cdot \mathrm{R_{ds(on)max}} = 4\mathrm{A} \cdot 0.14\Omega = 560\mathrm{mV}$$

Now it is clear that the best expected input voltage range is Range B (120 mV < V_{ds} < 450 mV). So, by setting the monitoring threshold mentioned above also the first stage gain is set (to 2.5 (V/V)).

Then, the gain of the second stage must be chosen to have the output value of the CSO_x centered around the average value of its accepted output range (for Range B, 0.3 V < CSO_x < V_{DD}-0.3 V). In this specific case, a second stage gain of 3 (V/V) is chosen and set. So, the average output voltage will be given by:

$$\mathrm{CSO_x} = \mathrm{V_{ds\_AVG}} \cdot 1^\circ\,\mathrm{gain} \cdot 2^\circ\,\mathrm{gain} = 140\mathrm{mV} \cdot 2.5\left[\frac{\mathrm{V}}{\mathrm{V}}\right] \cdot 3\left[\frac{\mathrm{V}}{\mathrm{V}}\right] = 1.05\mathrm{V}$$

Considering the maximum values calculated above:

$$\mathrm{CSO_x} = \mathrm{V_{ds\_max}} \cdot 1^\circ\,\mathrm{gain} \cdot 2^\circ\,\mathrm{gain} = 560\mathrm{mV} \cdot 7.5\left[\frac{\mathrm{V}}{\mathrm{V}}\right] = 4.2\mathrm{V}$$

Following the specifications of the device, this value could be read by the system only if the supplied V_{DD} is high enough, for example V_{DD} = 5 V.

With these elements, the following table shows some examples of gain settings of the two stages of the CSO_x for the STD12NF06LAG MOSFET, depending on the current flowing through it.

Table 1. CSO gain settings for STD12NF06LAG

Peak Current (A)	V_{ds} (mV)		Drain-source threshold (mV)	1° stage gain (V/V)	2° stage gain (V/V)	CSO output voltage (mV)	
	$R_{ds(on)typ(25°C)}$ = 70 mΩ	$R_{ds(on)max(150°C)}$ = 140 mΩ				Typ	Max
0.4	28	56	75	10	3	840	1680
0.8	56	112	150	10	1.5	840	1680
1	70	140	200	2.5	3	525	1050
2	140	280	300	2.5	1.5	525	1050
2.5	175	350	400	2.5	1.5	656	1312
3	210	420	500	2.5	1.5	787	1574
4	280	560	600	2.5	1.5	1050	2100

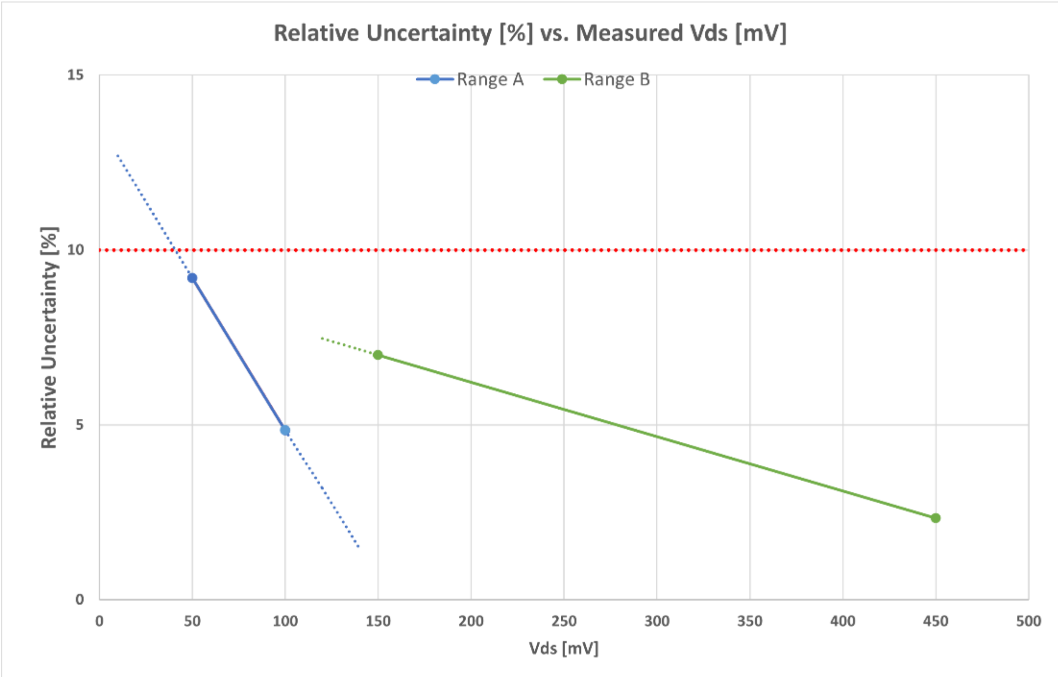
Obviously, for the same current load, changing MOSFET family means changing the $R_{ds(on)}$ and therefore the V_{ds} across it. With a different V_{ds}, the user is still able both to adapt, through SPI, the gain of the two stages of the CSO_x and to select and monitor the drain to source voltage on any MOSFET, with $R_{ds(on)}$ very low or very high, for the purpose of load current estimation.

Some final considerations must be made to reduce the uncertainty linked to the measurement of the V_{ds} and to optimize later the calibration phase of the system. As can be seen in the datasheet of L99MH98, the absolute uncertainty linked to the V_{ds} measurement system has been evaluated in four different configurations ($CSO_{xTOT\_error\_CHx\_50mV}$, $CSO_{xTOT\_error\_CHx\_100mV}$, $CSO_{xTOT\_error\_CHx\_150mV}$, and $CSO_{xTOT\_error\_CHx\_450mV}$) for both voltage input ranges, A and B:

- $CSO_{xTOT\_error\_CHx\_50mV}$ = ±4.6 mV, (evaluated in Range A, with CSO_{xGain} = 15 and 30 (V/V))
- $CSO_{xTOT\_error\_CHx\_100mV}$ = ±4.85 mV, (evaluated in Range A, with CSO_{xGain} = 15 and 30 (V/V))
- $CSO_{xTOT\_error\_CHx\_150mV}$ = ±10.5 mV, (evaluated in Range B, with CSO_{xGain} = 7.5 (V/V))
- $CSO_{xTOT\_error\_CHx\_450mV}$ = ±10.5 mV (evaluated in Range B, with CSO_{xGain} = 7.5 (V/V))

These values can be read as the total error introduced by the overall V_{ds} measurement system (offset on the input, uncertainty on the gain stages). If this absolute uncertainty is then elaborated as a relative parameter ($CSO_{xTOT\_error\_CHx}/V_{ds}$), it can be indeed useful to visualize its trend and add important comments to guarantee more accurate measurements later.

Figure 2. Relative uncertainty vs measured Vds



The above graph shows the trend of relative uncertainty with respect to V_{ds}, evaluated for both possible input ranges. As can be seen, therefore, it is possible to state that the accuracy of this system improves, and the impact of the error lowers, by performing measurement increasing V_{ds} values (within the ranges).

To measure a V_{ds} with a low relative uncertainty (<±5%) in Range A, it is recommended to work with V_{ds} >100 mV, while for Range B, at least V_{ds} >275 mV. Moreover, this means that to improve the precision of the measuring system it is necessary to use MOSFETs whose V_{ds} reaches at least 75 mV (middle value of Range A) at ambient temperature.

Reconsidering the previous example, since MOSFET STD12NF06LAG has a $R_{ds(on)}$ typ of 70 mΩ, it is suggested to use it with a current higher than 1 A. If instead in the application are present MOSFETs with small $R_{ds(on)}$, 5 mΩ typ for example, it is suggested to use them with larger currents and produce a voltage drop similar to the values mentioned above.

Moreover, it can be added that the selected drain-source voltage is always reflected on the CSO_x pin whether the MOSFET is ON (fixed or piloted by a PWM) or OFF (pins are tri-stated).

For the indirect current measurement system, when a PWM signal is provided to drive a load, it is highly recommended to **map to the CSO** channel to the H-bridge's **MOSFET which will always remain ON**, and so taking into consideration its V_{ds} as the main voltage reference for next calculations.

1.2 $R_{ds(on)}$ calibration

After a proper setting of the CSO_x channels, the first phase of the indirect current measurement method is the calibration of the value of $R_{ds(on)}$, relative to the MOSFET adopted at application level to drive the motor. This step is fundamental, since during indirect measurements the variations in temperature caused by the current will also change the $R_{ds(on)}$ value itself, so a starting reference value is necessary to estimate its trend with a variable temperature (caused by a variable current). This value, which from now on will be referred as $R_{ds(on)CAL}$, obtained during calibration, must also be paired with a specific calibration temperature, for example 25°C (since the $R_{ds(on)}$ value is linked to temperature variations, its calibration value must be associated with this initial calibration temperature, T_{CAL}).

If a high accuracy measurement on $R_{ds(on)CAL}$ is not already provided externally, an easy way to evaluate this starting point is to measure this resistance experimentally by letting a fixed known current, I_{CAL}, flow in the switched-on MOSFET and let the L99MH98 perform a fast V_{ds} measurement (fast enough to not heat the MOSFET and raise the $R_{ds(on)}$).

This can be easily arranged by turning on the single targeted MOSFET with a static load connected, letting a measurable current flow through it and then reading the V_{ds} across the MOSFET (using CSO_x measurements or other laboratory instruments).

The $R_{ds(on)CAL}$ can be then estimated as:

$$R_{ds(on)CAL} = V_{dsCAL}/I_{CAL}$$

Where, if the V_{ds} has been read by a CSOx channel, is:

$$V_{dsCAL} = CSO_{xCAL}/CSO_{xGain}$$

Since this calibration value must be associated with a calibration temperature, this resistance can be for example evaluated at 25°C. So, in this case, $R_{ds(on)CAL}$ can also be declared as $R_{ds(on)@25°C}$. This procedure can also be operated differently and at different calibration temperatures, as long as it is performed with high accuracy. Although these values are usually provided in the MOSFET datasheets, these resistances are typical values and may differ from the actual value of the individual device in use. This difference could in fact impact on the final accuracy of the measurement system since the calibration value is one of the most important parameters to later evaluate the current.

Example: calibration

To test the overall indirect measurement system, a practical example can be carried out during the different phases of the method. L99MH98 has been equipped in an applicative scenario (a validation board) to pilot a H-bridge composed by STD12NF06LAG MOSFETs from STMicroelectronics. The aim is later to pilot a compatible DC motor in PWM and measure its current (with an average of 2 A).

Following the previous example and considerations, it has been already evaluated that for a MOSFET with 70mΩ typical $R_{ds(on)}$, the expected input V_{ds} of the measurement system is within Range B (140mV < V_{ds} < 450mV) and the selected CSO_{xGain} is 7.5 (V/V).

To evaluate the $R_{ds(on)CAL}$ of this setup, a constant 3.5 A current has been flowed through the HS of the piloting circuit adopting an active load, and its V_{ds} has been measured on a digital oscilloscope probing the CSO_x channel as soon as the MOSFET has been turned on by the device. Then its value has been calculated as a ratio between CSO_{xCAL} (the average value measured by the scope after the MOSFET turn-on) and its CSO_{xGAIN}. In this way $R_{ds(on)25°C}$ is obtained dividing V_{dsCAL} by I_{CAL}.

The following table shows the measurements obtained:

Table 2. **Calibration Values**

I_{CAL} (A)	T_{CAL} (°C)	CSO_{xGain} (V/V)	CSO_{xCAL} (V)	V_{dsCAL} (mV)	$R_{ds(on)25°C}$ (mΩ)
3.48	25	7.5	1.627	216.93	62.34

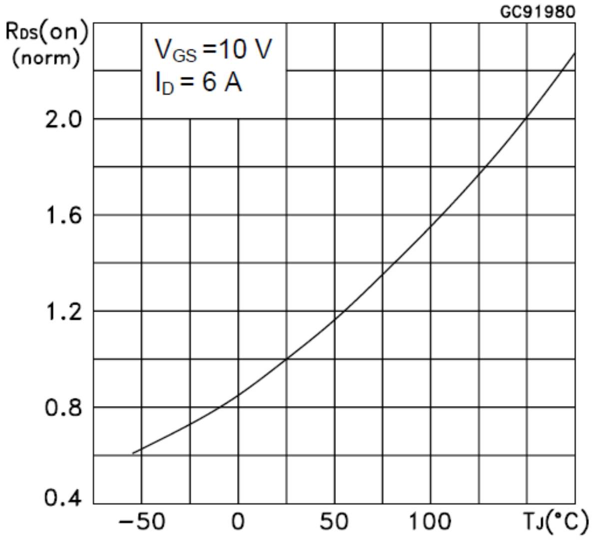
Where I_{CAL}, T_{CAL}, and CSO_{xCAL} have been directly measured, while V_{dsCAL} and $R_{ds(on)CAL}$ are calculated.

1.2.1 $R_{ds(on)}$ curve vs temperature

As already stated, the temperature has a strong effect on the $R_{ds(on)}$ of a MOSFET. In many models in fact, the ON resistance value approximately doubles from 25°C to 175°C. Assuming this trend, the temperature coefficient of $R_{ds(on)}$ is comparable to the slope of the curve obtained by the normalization of the $R_{ds(on)}$ on its $R_{ds(on)@25°C}$ value. This coefficient is always positive.

In Figure 3 the example of the obtained curve (from the datasheet of the device). It is possible to notice in this figure that this particular MOSFET already doubles its $R_{ds(on)@25°C}$ value at 150°C.

Figure 3. **STD12NF06LAG $R_{ds(on)}$ normalized v. temperature**



As said, the purpose of the indirect measurement system integrated in L99MH98 is to calculate the current of the load thanks to the strong dependence of the $R_{ds(on)}$ to temperature variations. As seen, this system needs at first an evaluation of the $R_{ds(on)}$ at room temperature (as a calibration phase); i.e., 25°C. Then it uses the temperature measurement system equipped (see next chapter), when the MOSFET is driving a load, in order to calculate the value of the $R_{ds(on)}$ at the working temperature (reached during piloting). This second step is necessary since the current flowing through the MOSFET heats it, and so varies its temperature and, consequently, its $R_{ds(on)}$.

To achieve this measure, it is necessary to follow the evolution of the $R_{ds(on)}$ when the temperature varies. The suggested approach is to integrate, inside the microcontroller memory, the on-resistance curve of the MOSFET that will be used in the application. This is possible by inserting the mathematical relationship which describes this curve and its trend.

For example, a preliminary evaluation of the normalized $R_{ds(on)}$ on $R_{ds(on)@25°C}$ curve can be elaborated as a simple first-degree equation (line passing through two points):

$$(y - y_1)/(y_2 - y_1) = (x - x_1)/(x_2 - x_1)$$

Knowing that when the temperature is 25°C, its normalized value is 1, and when the curve reaches 150°C, it is 2.

$$\left(R_{ds(on)norm} - 1\right)/\left(2 - 1\right) = \left(T - 25°\mathrm{C}\right)/\left(150°\mathrm{C} - 25°\mathrm{C}\right)$$

$$R_{ds(on)norm} = 1 + \left(T - 25°\mathrm{C}\right)/125°\mathrm{C}$$

$$R_{ds(on)}\left(T\right) = R_{ds(on)25°\mathrm{C}}\left[1 + \left(T - 25°\mathrm{C}\right)/125°\mathrm{C}\right]$$

The calculation of the $R_{ds(on)}$ value in this case depends on $R_{ds(on)@ 25°C}$. This value is known when the user performs a calibration at this specific temperature (25°C).

However, if the calibration is executed at a different temperature, the user should still provide this $R_{ds(on)@ 25°C}$ value, which is different though from the value retrieved during calibration. To avoid the dependence of the calculation on the value $R_{ds(on)@25°C}$ the relationship can be changed rewriting this value as:

$$R_{ds(on)25°\mathrm{C}} = R_{ds(on)CAL}/R_{ds(on)norm}$$

where $R_{ds(on)norm}$ = ($R_{ds(on)}$/$R_{ds(on)@25°C}$) ratio can be retrieved by the $R_{ds(on)}$ normalized curve (Figure 3).

For example, the user could calibrate the system at T=-40°C and knowing the $R_{ds(on)@-40°C}$ can replace the previous coefficient in this way:

$$R_{ds(on)}\left(T\right) = \left(R_{ds(on)\,-\,40°\mathrm{C}}/0.64\right)\cdot\left[1 + \left(T - 25°\mathrm{C}\right)/125°\mathrm{C}\right]$$

where 0.64 is the value of ($R_{ds(on)CAL}$/$R_{ds(on)@25°C}$) obtained by the curve shown in Figure 3.

Moreover, for an advanced profiling of the $R_{ds(on)}$ calibration curve, it is suggested to adopt a second order function (parabola):

$$R_{ds(on)}(T) = \left(R_{ds(on)CAL}/R_{ds(on)norm}\right) \cdot \left[aT^2 + bT + c\right]$$

Where the coefficients a, b and c are retrieved mathematically by identifying three points shown in Figure 3 (or even by performing an experimental 3-points calibration).

This mathematical equation, once characterized with the correct parameters can easily be implemented inside the microcontroller memory as a part of the measuring algorithm and will return the value of the $R_{ds(on)}$ at any temperature (given by the temperature measurement system).

Example: $R_{ds(on)}$ curve

To trace the $R_{ds(on)}$ evolution of the MOSFET selected for the proposed example, from Figure 3 these three points have been selected:

P_1 = (-25°C, 0.72), P_2 = (25°C, 1) and P_3 = (150°C, 2)

The coefficients calculated for a second order curve passing through these three points are:

a = 0.0000133, b = 0.0057, c = 0.850

Considering that these are coefficients for a normalized curve, to convert to the real $R_{ds(on)}$ value, these parameters must be multiplied by $R_{ds(on)CAL}$ (in this case equal to $R_{ds(on)@25°C}$), which is 62.34 mΩ:

a = 0.0008312, b = 0.3532, c = 52.987

The final relationship is:

$R_{ds(on)}$ = 0.0008312T2 + 0.3532T + 52.987

1.3 Temperature measurement system

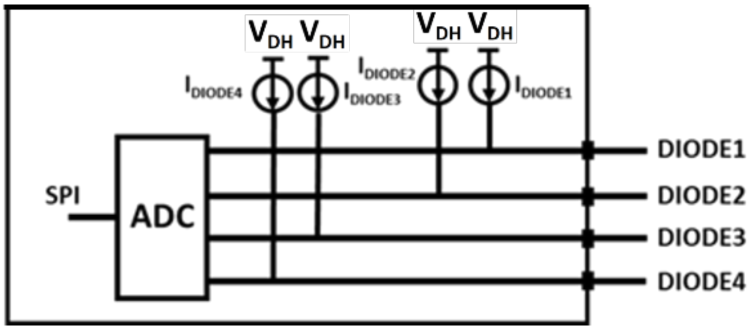
To retrieve information on the working temperature of the application and allow $R_{ds(on)}$ estimation, L99MH98 provides a temperature sensing mechanism, performing constant readings of the voltage across external diodes/ diode chains. Each diode/chain will be placed, in fact, as close as possible to one of the piloting MOSFETs of each H-bridge (in case to support the Indirect current sense method) or to a relevant temperature hot-spot (when the diodes are used as PCB temperature monitor). Monitoring the external diodes/chains also allows in fact to detect steady, not fast, temperature rising of specific PCB areas, so a temperature control of the module can be also implemented.

This sensing structure reads the forward voltage of the diode, $\mathbf{V_f}$, which also varies with temperature. In this specific case, in fact, the diode will experience a temperature which is strongly related to the junction temperature of the closest MOSFET (the diode fully behaves as a temperature sensor). Therefore, the diode's temperature coefficient is needed to get an accurate representation of its voltage variation during the load piloting phase.

Specifically, for this indirect system, the strategy to adopt diodes as thermal sensors has been preferred with respect to other sensing devices, such as NTC (Negative Temperature Coefficient) thermistors, for their better response time.

In this device, the control structure of the diodes is constituted by a current generator that injects a small current (programmable from 250 µA to 1 mA) in the anode of the diode, while the cathode will be connected to the ground of the application board. The voltage across the diode is then measured by an internal ADC which will make the value available externally (to a microcontroller): 4 dedicated registers, `DIODEx_READ`, x = 1..4, will report the read values to the microcontroller by SPI communication. The interval time between one temperature measurement and another is of 1 ms typ. If one or more diodes are not connected the information related to the connected diodes will always be available after 1 ms.

Figure 4. **Temperature measurement system**



`DIODEx_READ` register is 11 bits: since the maximum binary value is 11111111111_2 (2048_{10}), the maximum voltage value that can be read by the integrated ADC is 2.2 V. To read the voltage value across the diode, the following formula must be applied:

V_{diode} (V) = ($\mathtt{DIODEx\_READ}_{10}$) . ($2.2/2048_{10}$) ≈ $\mathtt{DIODEx\_READ}_{10}$. 0.00107421875

In the case of a chain of diodes, the final voltage value retrieved is then divided by the same number of diodes of the chain:

V_{diode} (V) = ($\mathtt{DIODEx\_READ}_{10}$. 0.00107421875)/n_{diodes}

This system can be also evaluated to pair the T_{CAL} measured during calibration.

Example: V_{diode} measurement

During calibration, when T=25°C, a chain of two diodes in series and polarized with 1 mA, returned a $\mathtt{DIODEx\_READ}_{10}$ register of 1101, corresponding to:

V_{diode} = (1101 * 0.00107421875)/2 = 0.591 V.

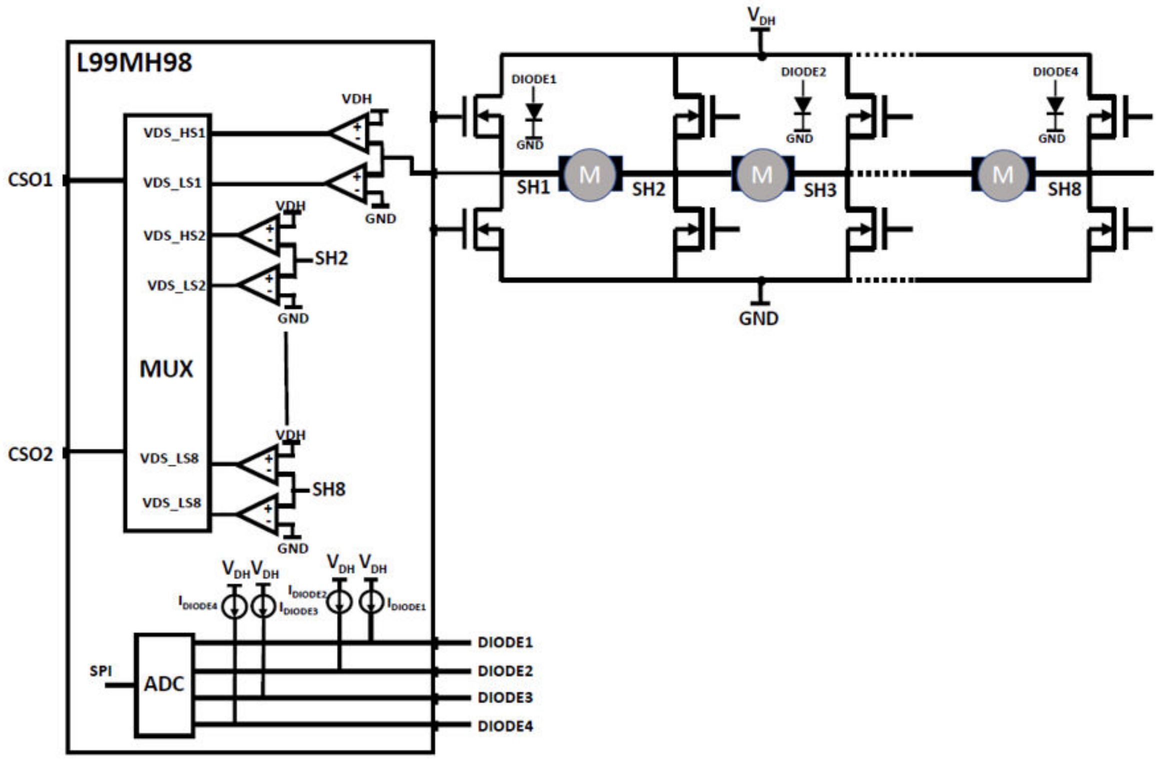
1.4 Real-time V_{ds} measurement

As previously illustrated, internally, L99MH98 is able to reflect the drain to source voltage across each of the external MOSFETs (through an analog multiplexer) on one of the two CSO_x pins, with a specific gain. This allows the user, via SPI, to set which MOSFET and Drain to Source voltage to monitor for the purpose of load current estimation.

The selected voltage is in fact reflected on the CSO_x pins when the MOSFET is fully ON. This is possible because the CSO_x pins are tri-stated when the selected MOSFET is OFF, and are enabled, after some dead time, when the MOSFET is ON. This allows a small capacitor on the CSO_x pins to sustain the amplified Drain to Source voltage for reliable conversion by the ADC of the microcontroller.

As previously suggested, when the load is piloted by a PWM signal, the CSO_x should be mapped to the MOSFET kept ON, in order to get the full trend of the current.

Figure 5. V_{ds} to CSO_x mapping



Once the calibration phase is achieved by setting the CSO_X gain and correlating the calibration temperature of the diode with the calculated $R_{ds(on)CAL}$ (relative to a specific calibration current), the indirect current sense method can be applied to test and identify the load current. This is possible mainly by recovering the V_{ds} and $R_{ds(on)}$ values during working conditions.

During this measuring phase, the V_{ds} value (voltage across the selected MOSFET, associated to one of the two CSO_x pin) is recovered knowing the CSO_x value through a direct reading of the pin (in real scenarios, this voltage will be then sampled by the MCU ADC) and by simply dividing next this value by the gain previously set.

Example: V_{ds} measurement

Since the calibration phase is finished, the testing phase can be now applied to calculate a real load current. Up to now, here the collected calibration values:

Table 3. Calibration values

I_{CAL} (A)	T_{CAL} (°C)	$V_{diodeCAL}$ (V)	CSO_{1Gain} (V/V)	CSO_{1CAL} (V)	V_{dsCAL} (mV)	$R_{ds(on)CAL}$ (mΩ)
3.48	25	0.591	7.5	1.627	216.93	62.34

And the trend of the $R_{ds(on)}$, varying the temperature, has been estimated as:

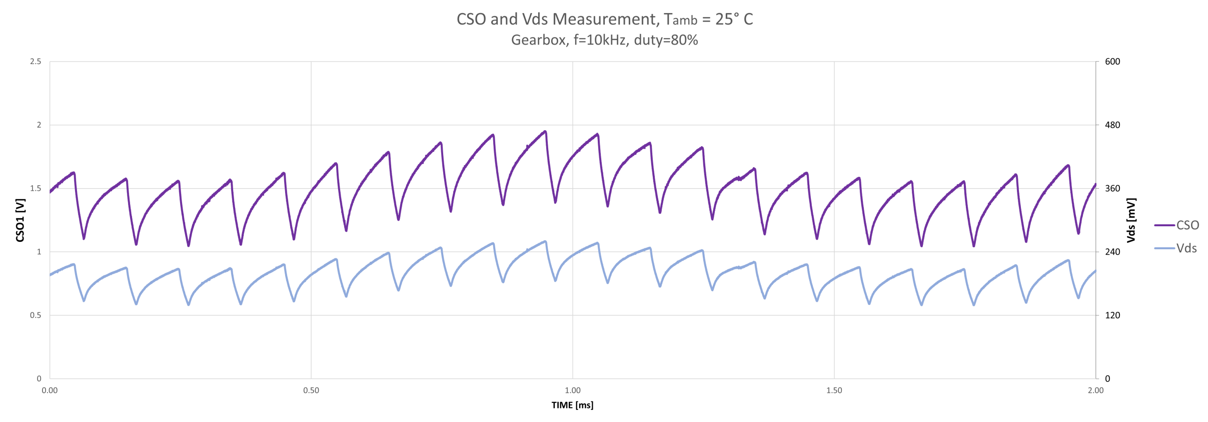
$R_{ds(on)}$ = 0.0008312T2 + 0.3532T + 52.987

Recalling now the applicative scenario here adopted as an example: an H-bridge with STD12NF06LAG MOSFETs has been driven by L99MH98, equipped in a real validation board. A brushed DC motor (automatic gearbox, 2 A average current) has been then connected between the HS and LS MOSFETs (of two different branches) of the bridge.

Piloting the LS with a switching PWM signal of 10 kHz and with an 80% of duty cycle, the V_{ds} of the HS switches following the trend of the current flowing through this MOSFET. This varying voltage is then reflected by the internal current sense amplifiers on the CSO_1 channel, with a 7.5 (V/V) gain factor.

Probing with a digital oscilloscope this pin (emulating an ADC sampling) and acquiring a 2 ms curve, it trend can be show and the V_{ds} trend, dividing each sampled point of CSO_1 by its gain:

Figure 6. **CSO and V_{ds} measurements**



In the above figure, the CSO_1 and V_{ds} values are shown (with two different scales, each with a specific reference vertical axis). It can be added that:

Table 4. **CSO_1 and V_{ds} measurements**

CSO_{1AVG} (V)	CSO_{1MAX} (V)	CSO_{1MIN} (V)	V_{dsAVG} (mV)	V_{dsMAX} (mV)	V_{dsMIN} (mV)
1.502V	1.952	1.044	200.3	260.2	139.22

1.5 Temperature measurement for $R_{ds(on)}$ estimation

The second term necessary to the identification of the current is the $R_{ds(on)}$ value.

To estimate this value at any compatible working condition, the indirect method relies on the calculation of the $R_{ds(on)}$ following the equation previously modeled during calibration phase. This means that a measure of the working temperature must be executed.

The proposed strategy, as seen, is to adopt a temperature measurement system equipped, in the final application, with an external diode/chain of diodes. L99MH98 will be able then to periodically read the forward voltage of the diode, V_f.

This voltage, inversely proportional to the temperature, is readable every 1ms through SPI. The IC provides in fact this value in the 11-bit register `DIODEx_READ`. This $V_{f,x}$, at an unknown temperature, can be then mathematically associated to an estimated temperature, knowing the $V_{f,amb}$ (at 25°C for example) and knowing the temperature coefficient of the diode (usually addressed as **α**). The microcontroller will then calculate this working temperature by applying the following formula:

$$T = T_{amb} + \left(V_{f,x} - V_{f,amb}\right)/\alpha$$

Where:

- T_{amb} (°C) is the reference temperature, usually 25°C
- $V_{f,amb}$ is evaluated experimentally at the same reference temperature (so $V_{f,25°C}$) and varies slightly from device to device
- α is the temperature coefficient of the forward voltage (typically in the range of -2 mV/°C to -2.5 mV/°C for silicon diodes). This characterization parameter can be provided by the datasheet of the equipped diode or easily recovered experimentally during calibration phase.

Example: working temperature measurement

In the example here reported, the validation board equips an external diode/diodes chain with a temperature coefficient α = -2 mV/°C.

This value has been easily evaluated by heating the diode at two different temperatures (for example, 25°C and 150°C) and then reading the different V_f measured. Then the coefficient has been calculated dividing the difference between the voltages (in mV) by the difference between the temperatures.

Considering that in the case presented, $V_{diodeCAL} = V_{f,amb}$ = 0.591V. Then, in working condition, the voltage read by the diodes chain closest to the piloting HS of the motor (with a $\mathtt{DIODE2\_READ}_{10}$ register of 1065), corresponds to a V_{diode} = 0.572 V.

Then the working temperature has been evaluated as:

T = 25°C + (572 mV – 591 mV)/-2 mV/°C ≈ 34.67°C

1.5.1 Further improvements for temperature measurement

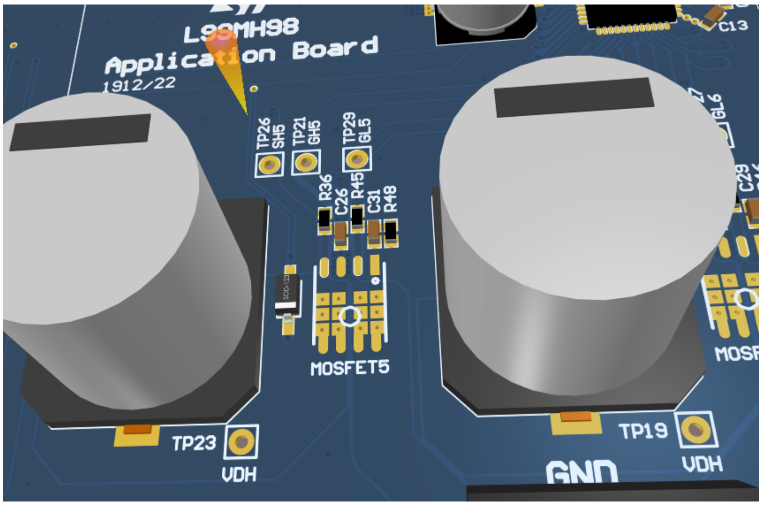
One of the most important parameters to be measured in the indirect current sensing system is the temperature reached by the MOSFET when the current flows, e.g. when a motor is running. The measurement of the temperature is made through one or more diodes which must be suitably placed. Thus, two of the most important choices to be made are where to place the diodes and how to correlate the temperature measured by the diode with that of the MOSFET.

Few rules must be followed in choosing the position of the diode(s):

1. Since generally a motor is piloted by a half-bridge and one of its MOSFET is piloted by a PWM signal, it is preferable to place the diode(s) near the H-bridge's MOSFET which will always remain on. This is suggested because this MOSFET will always have a current flowing through it, even when the PWM is off, implying a more stable temperature to measure.
2. However, if the motor is piloted in both directions, and the ON-MOSFET is switched during application with its equivalent of the other branch, the diode(s) could be placed right in the middle of the driving devices (e.g. between two high-sides).
3. To further improve the position of the diode and correctly link its temperature and the junction temperature of the MOSFET, thermal maps can be used.

The following figure shows the positioning of the diode used in the application board designed to drive a general MOSFET. It is possible to see that the diode is placed, in fact, very close to MOSFET itself.

Figure 7. **Diode positioning**



A suggested improvement of the measurement system can consist of a mathematical parameter able to link the temperature of the sensor (the diode/diodes chain) as close as possible to the real MOSFET's temperature, using a thermal map.

Even with a visual examination using a thermal camera, a difference can be noticed between the temperature at which the diode's case is brought respect to the temperature to which the MOSFET's case is brought. Figure 8 and Figure 9 show the thermal maps of the STD12NF06LAG MOSFET are shown (in this case a chain of two diodes is considered). It is possible to observe a temperature difference between the case of the MOSFET and the case of the diodes, respectively 43.4°C and 37.8°C.

Figure 8. **Thermal map of STD12NF06LAG MOSFET with 2 A load current, MOSFET case centered**

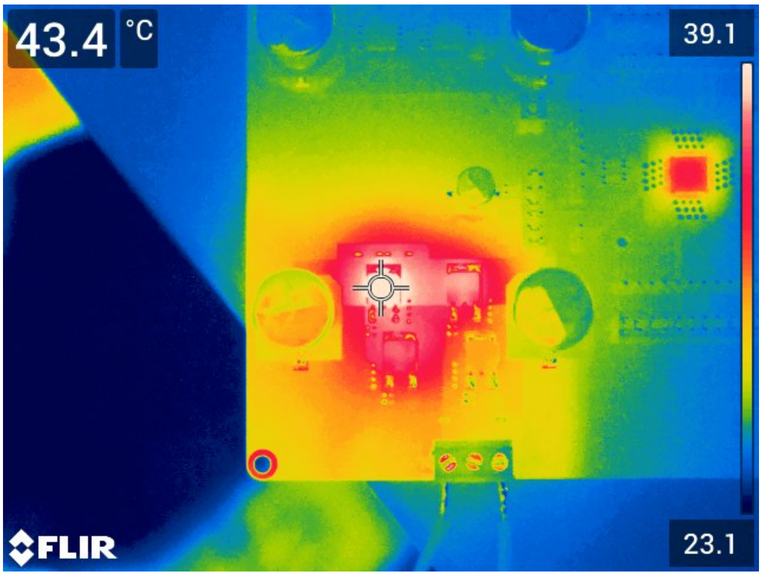
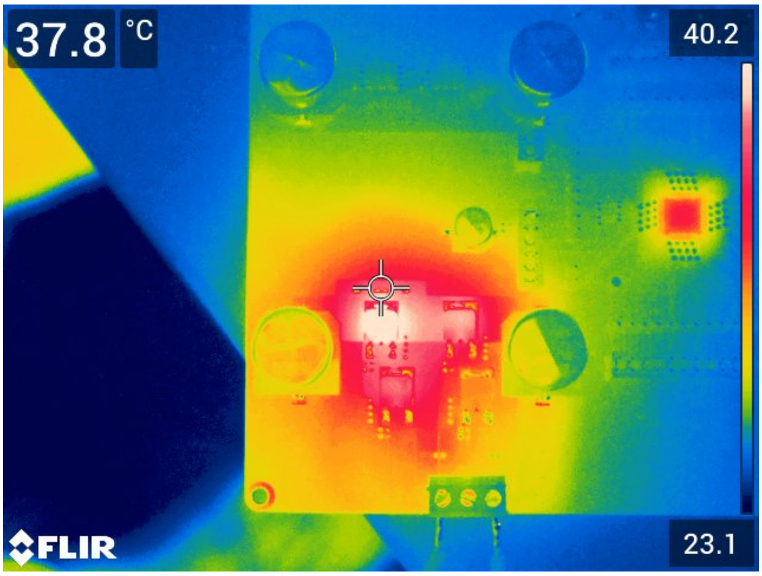
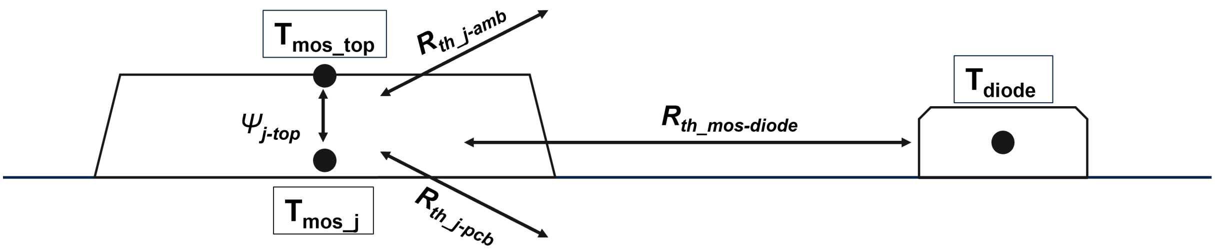


Figure 9. **Thermal map of diodes chain while a 2 A load maximum current flows in the closest MOSFET**



Through these maps and the relative temperature differences acquired it is possible to further improve the temperature measurement, adopting different approaches that involve some thermal parameters of the MOSFET (some are shown in the image).

Figure 10. **Some thermal parameters**



The main goal of these considerations is to get the junction temperature of the MOSFET (T_{mos_j} in Figure 10, or simply T_j), starting from the working temperature estimated by the forward voltage of the diode, so T_{diode} (behaving as a passive element, it is assumed that the diode does not experience a different temperature internally).

This analysis elaborates then a **Thermal Corrective Factor (ΔT)** to be integrated in the measuring algorithm, in order to obtain:

$$T_j = T_{diode} + \Delta T$$

In this document some suggested approaches are shown. The first approach uses the thermal resistance MOSFET-to-diode between the MOSFET and the diode to recover the T_j, junction temperature, of the examined MOSFETs calculated through the equation:

$$T_j = T_{diode} + R_{th_{mos-diode}} \cdot P_{mos}$$

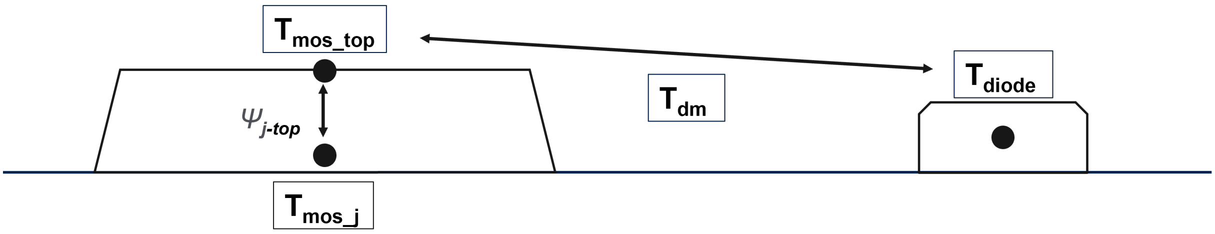
Where: T_j is the junction temperature of the MOSFET, T_{diode} is the temperature read by the diode, P_{mos} is the power of the MOSFET and $R_{th\_mos\text{-}diode}$ is the thermal resistance MOSFET-to-diode.

In this case, the parameter to be evaluated experimentally is $R_{th\_mos\text{-}diode}$.

Moreover, if the diode is very close to the MOSFET, it is possible to approximate the MOSFET-to-diode thermal resistance with the junction-to-PCB thermal resistance, reported in the datasheet of the MOSFET itself, without the need to calculate the MOSFET-to-diode thermal resistance.

The second approach adopts the $\boldsymbol{\psi_{j\text{-}top}}$ parameter and the thermal difference between the top of the diode and the top of the MOSFET (Figure 11).

Figure 11. Thermal difference using $\psi_{j\text{-}top}$



In this case the equation used to reconstruct the MOSFET junction temperature is:

$$T_j = T_{diode} + T_{dm} + \psi_{j-top} \cdot P_{mos}$$

Where: T_{dm} is the temperature difference between the diode and the MOSFET (top temperatures) and $\psi_{j\text{-}top}$ is the junction to top characterization parameter of the MOSFET (PSI junction-to-top).

In this case, the parameter to be evaluated is in fact $\psi_{j\text{-}top}$ and experimentally T_{dm}.

Example: thermal corrective factor

In the example carried out, the working temperature estimated during the 2 ms acquisition is 34.67°C. Then the thermal corrective factor has been evaluated knowing:

- T_{diode} = 34.67°C (temperature read by the chain of diode close to the selected HS)
- T_{mos_top} = 40°C (top temperature of the piloting HS read by the Thermocam during examination)
- T_{dm} = 40°C - 34.67°C = 5.33°C
- $\psi_{j\text{-}top}$ = 5.50°C/W (thermal characterization parameter of the selected MOSFET)
- P_{mos} = 0.61 W (averaged power consumption of the MOSFET during acquisition, evaluated experimentally)

Then:

T_j = 34.67°C + 5.33°C + 5.5°C/W * 0.61W = 43.33°C

1.6 Current calculation

After measuring the V_{ds} and the temperature at which the piloting MOSFET is working, the microcontroller is able to evaluate the working current. To achieve this, the measurement algorithm must follow two simple calculations:

1. The calculation of $R_{ds(on)}$ at working temperature. The microcontroller has in its memory the trend of $R_{ds(on)}$ curve of dependence on the temperature value: from the working temperature, it can estimate the value of the $R_{ds(on)}$ of the piloting MOSFET.
2. The final calculation of the current. At this point the microcontroller has calculated the value of the $R_{ds(on)}$ and measured the value of the V_{ds} (acquired with ADC): the current flowing in the H-bridge is easily calculated as **the ratio between $V_{ds}/R_{ds(on)}$**.

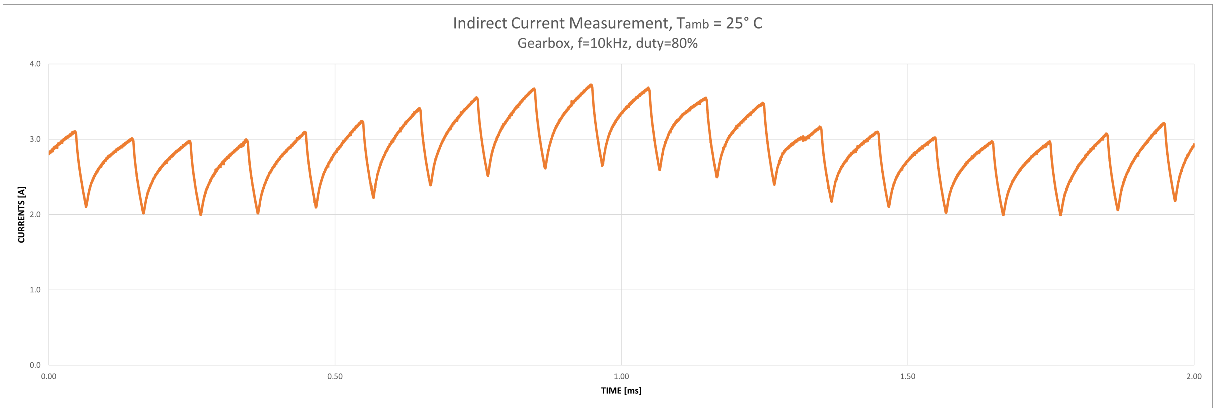
Example: $R_{ds(on)}$ estimation and current calculation

The final value of the $R_{ds(on)}$ is calculated as:

$R_{ds(on)}$ = 0.0008312 * (43.33)2 + 0.3532 * 43.33 + 52.987 = 69.85 mΩ

Now it is possible to evaluate the current flowing through the MOSFET, and in the DC motor, by dividing each V_{ds} sample (acquired by the digital oscilloscope, Figure 6) by this $R_{ds(on)}$ value, obtaining the following Figure:

Figure 12. **DC Motor (gearbox) indirect current sense measurement, T_{amb}=25°C**



It can be added that:

Table 5. **DC motor (Gearbox) indirect current sense measurements**

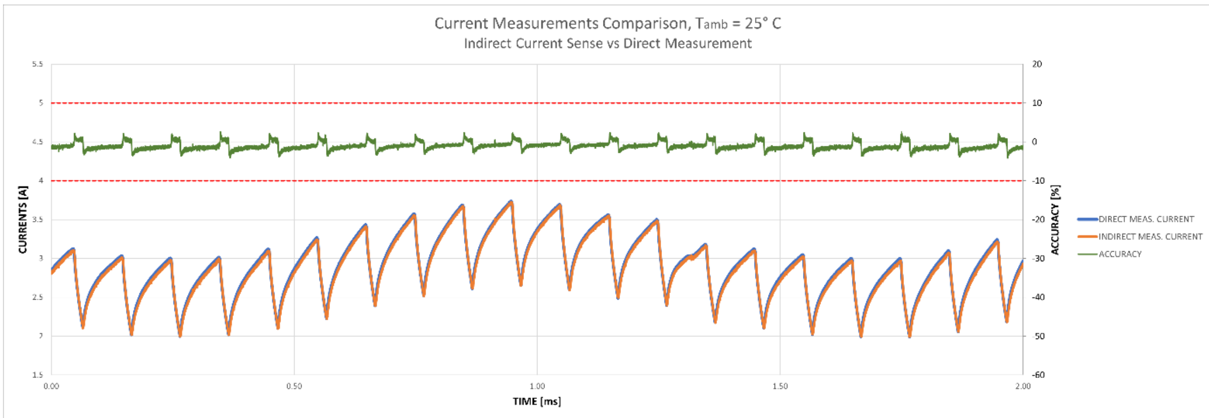
I_{AVG} (V)	I_{MAX} (V)	I_{MIN} (V)
2.868	3.715	1.993

1.7 Indirect current sense method: results

To evaluate the performance and validity of the indirect measurement system reported, a final estimation on the overall accuracy of the produced measurements can be evaluated. This result can be obtained by the comparison between two curves: the current calculated adopting the Indirect Current Sense of L99MH98 and the real value of the current flowing through the load.

Referring to the example already examined, since the curve of the measured current is a series of samples elaborated starting from the CSO_1 acquisition of the digital oscilloscope, a real measurement of the load current has been performed probing (with a current probe) the terminal of the motor connected to the piloting HS MOSFET.

Figure 13. DC motor (gearbox) indirect current sense measurement, T_{amb}=25°C



The above figure is the graphical comparison between the two different curves. The blue curve (in the background) is the current flowing through the HS STD12NF06LAG MOSFET, obtained by probing the terminal of the DC motor connected to the HS and then sampled by the oscilloscope. Some additional information:

Table 6. DC motor (gearbox) direct current measurements

I^*_{AVG} (V)	I^*_{MAX} (V)	I^*_{MIN} (V)
2.897	3.737	1.991

The orange curve (almost superimposed to the real one) is the current reconstructed by the Indirect Current Sense method. It is possible to appreciate that the current measured with the new method follows exactly the trend of the current of the motor.

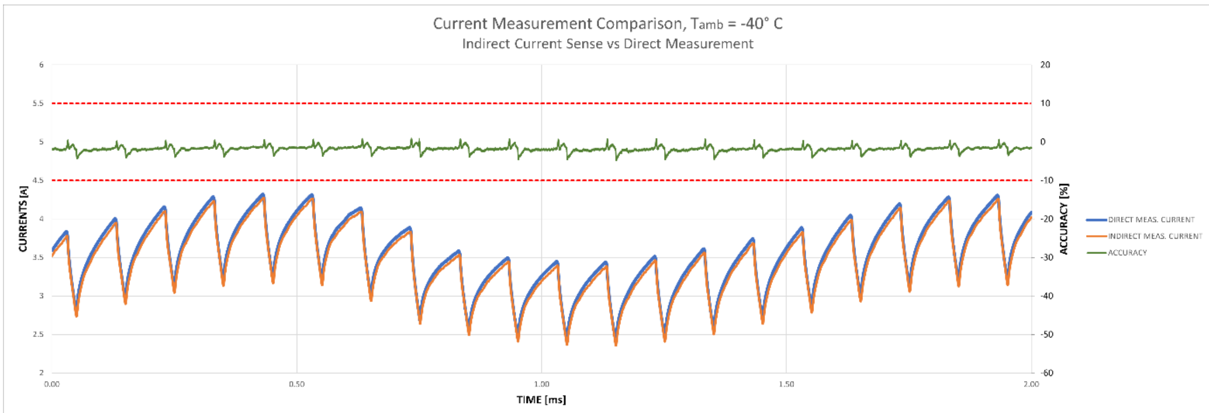
The accuracy parameter (green curve) has been evaluated as a relative error, point-to-point, by comparing both currents values, following the same time base (different vertical reference axis). It can be noticed that the accuracy of the acquisition is always within the range of ±10% (red dotted lines), which has been fixed as the maximum accuracy interval accepted by this indirect system. It can be evaluated in fact:

Table 7. Indirect measurement accuracy, T_{amb}=25°C

ΔI_{AVG}	$Range_{max}$
-1%	±4%

Moreover, additional measurements have been developed with the same calibration values but with different T_{amb}: -40°C, 85°C and 125°C.

Figure 14. DC motor (gearbox) indirect current sense measurement, T_{amb}=-40°C



The above figure has been obtained with an evaluated $R_{ds(on)}$ = 42.97 mΩ.

Table 8. Indirect measurement accuracy, T_{amb}=-40°C

ΔI_{AVG}	$Range_{max}$
-1.8%	±4%

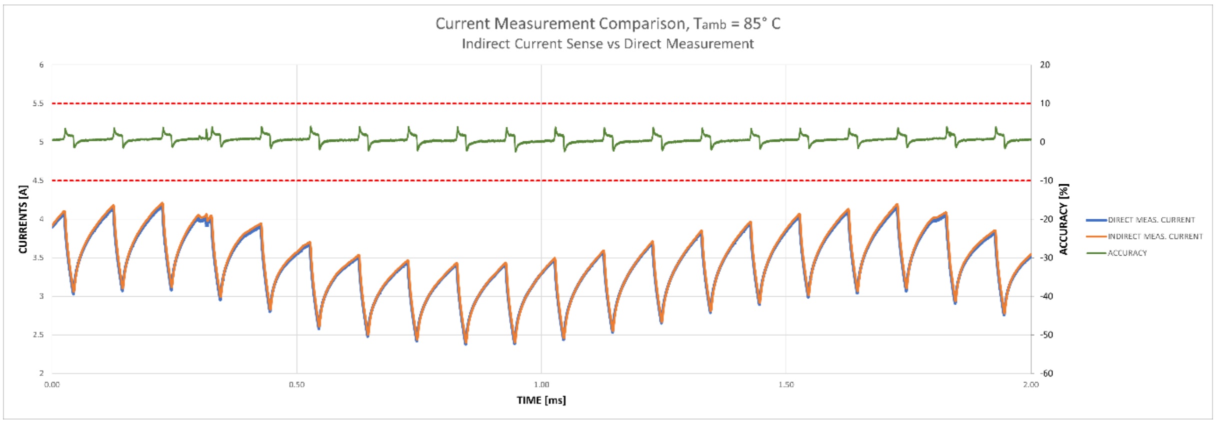


Figure 15. DC motor (gearbox) indirect current sense measurement, T_{amb}=85°C

The above figure has been obtained with an evaluated $R_{ds(on)}$ = 95.35 mΩ.

Table 9. Indirect measurement accuracy, T_{amb}=85°C

ΔI_{AVG}	$Range_{max}$
0.6%	±4%

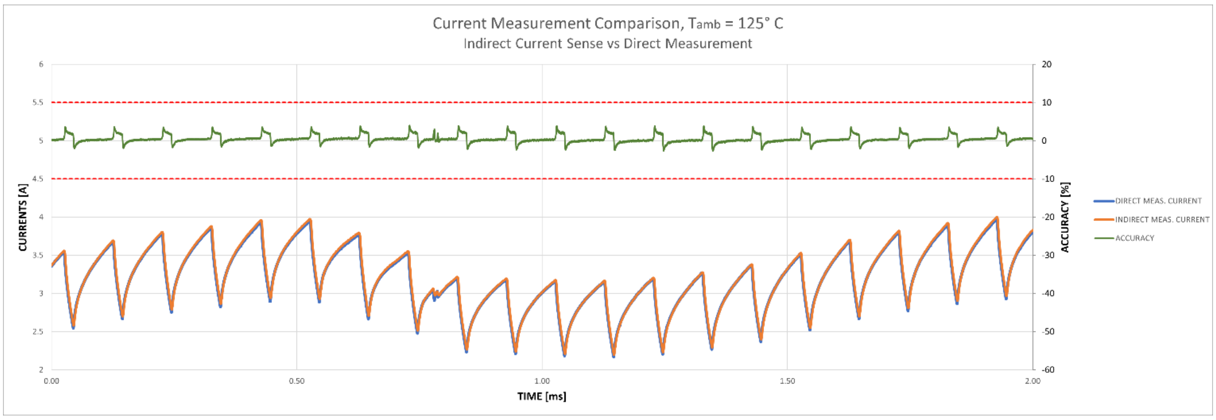


Figure 16. DC motor (gearbox) indirect current sense measurement, T_{amb}=125°C

The above figure has been obtained with an evaluated $R_{ds(on)}$ = 117.69 mΩ.

Table 10. Indirect measurement accuracy, T_{amb}=125°C

ΔI_{AVG}	$Range_{max}$
0.5%	±4%

A second example can be provided by showing another set of measurements executed by connecting a different, and closer to a potential application, DC motor: a Seat Control Sliding motor (forward direction). In this case a 20kHz PWM with 70% duty cycle has been configured as the piloting signal. Maintaining the same calibration values and following the same steps, the following figures have been developed:

Figure 17. **DC motor (seat adjustment) indirect current sense measurement, T_{amb}=25°C**

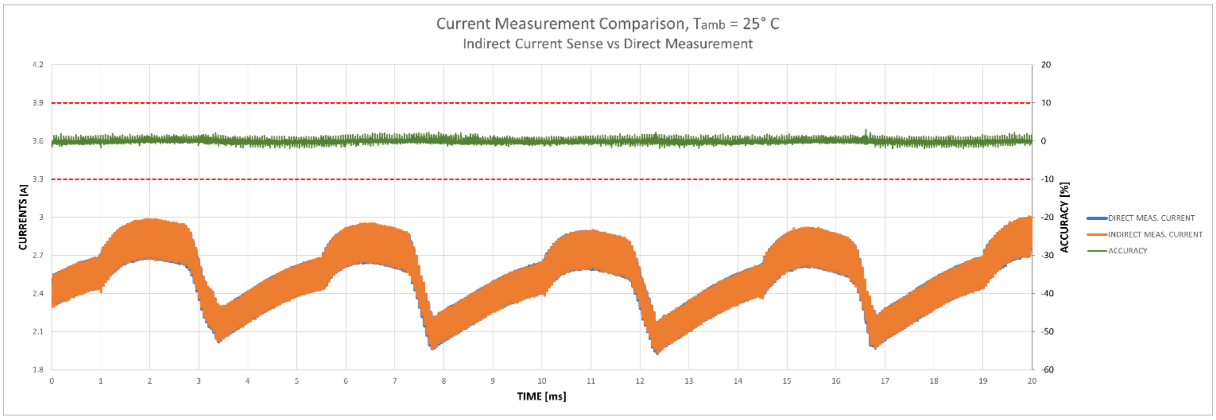


Figure 18. **DC motor (seat adjustment) indirect current sense measurement, T_{amb}=25°C - Section**

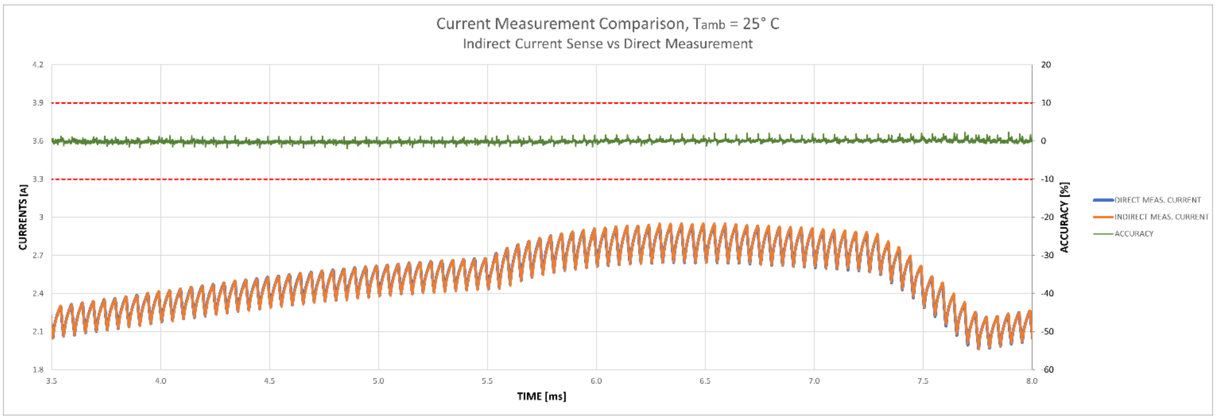


Figure 17 and Figure 18 have been obtained with an evaluated $R_{ds(on)}$ = 66.68mΩ.

Table 11. **DC motor (seat adjustment) indirect current sense measurements**

I_{AVG} (V)	I_{MAX} (V)	I_{MIN} (V)
2.539	3.006	1.929

Table 12. **Indirect measurement accuracy, T_{amb}=25°C**

ΔI_{AVG}	$Range_{max}$
-0.1%	±3%

2 PCB design guidelines

Two different PCBs were designed to test the functionality of the L99MH98, one called validation board and one called application board. The two PCBs have been used for the full validation of the device and some layout solutions have been adopted to optimize the indirect measurement system.

2.1 Reference schematic

The schematics of the two PCBs are reported here. The main difference in the two boards is in the external MOSFETs:

1. Validation board: 4 different external MOSFET topologies have been used to realize the 4 different H-Bridges.
2. Application board: in this case a single topology of MOSFET has been used to realize the 4 H-bridges.

Figure 19. Application board schematic, L99MH98 section

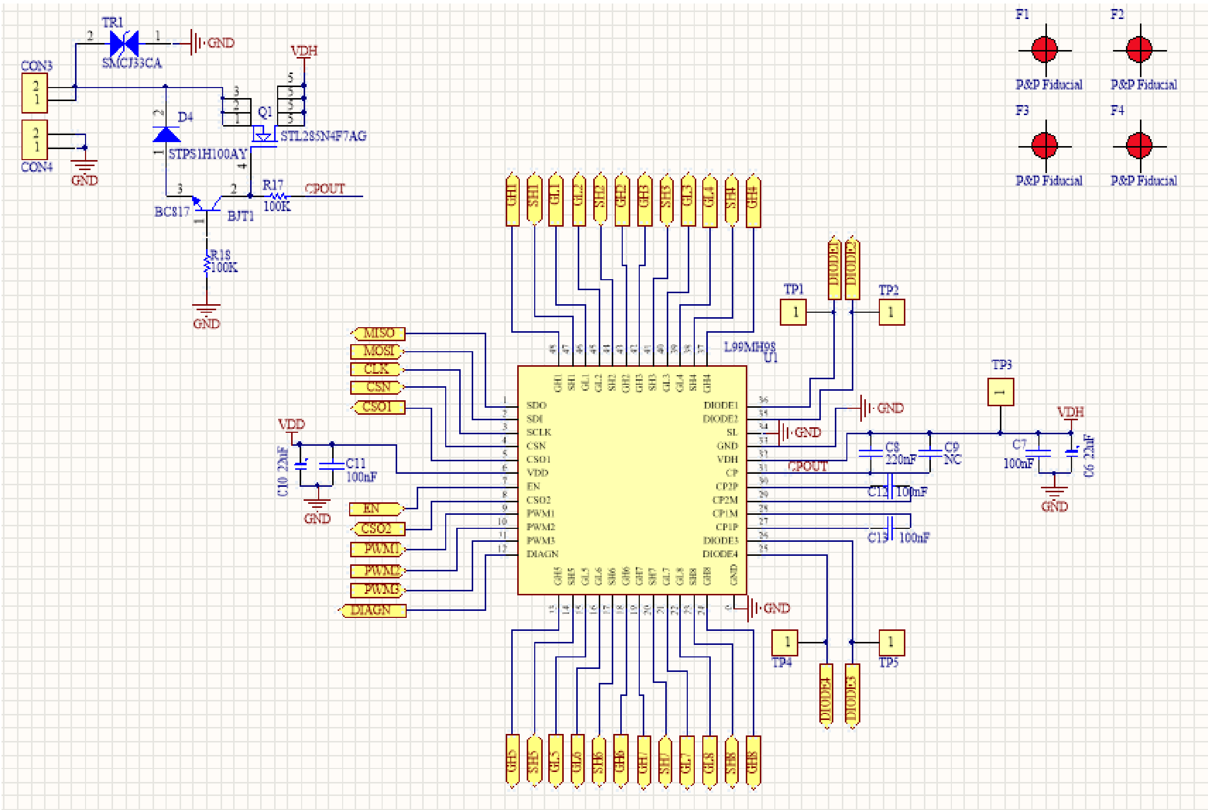


Figure 20. Application board schematic, power MOSFETs section

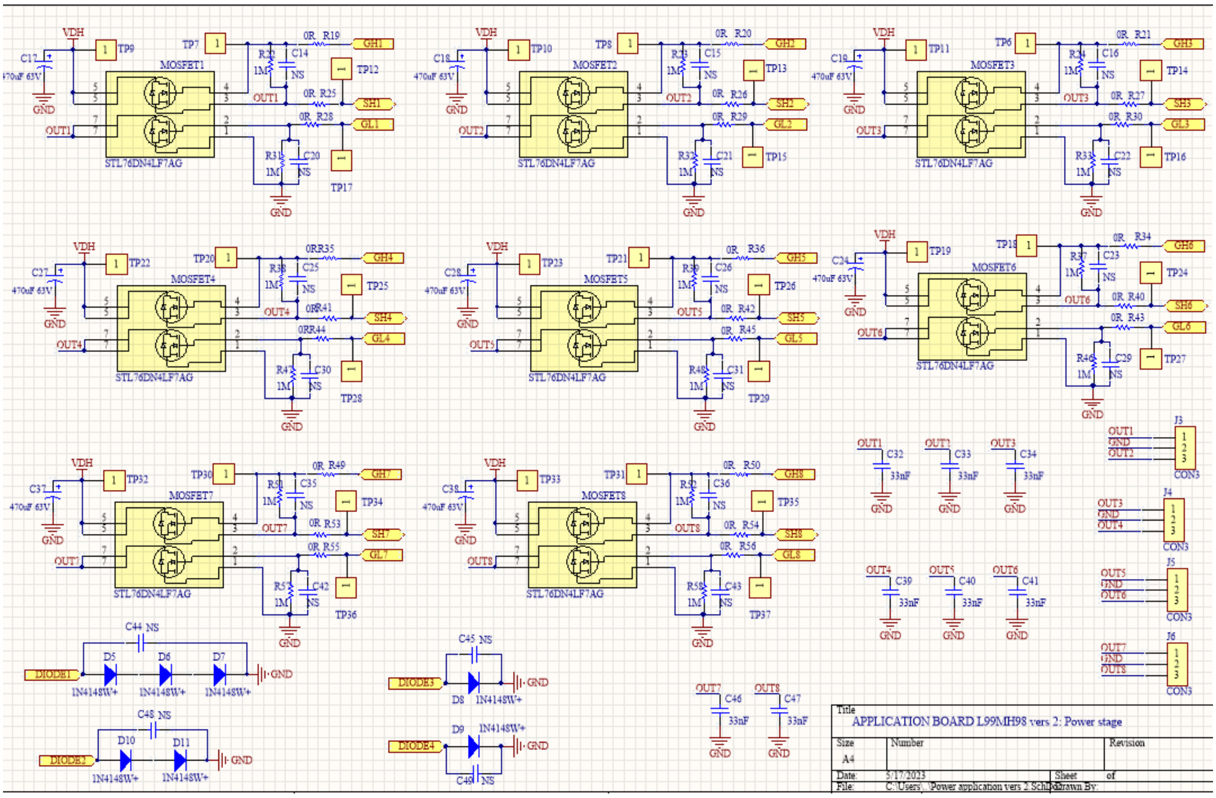


Figure 21. Validation board schematic, L99MH98 section

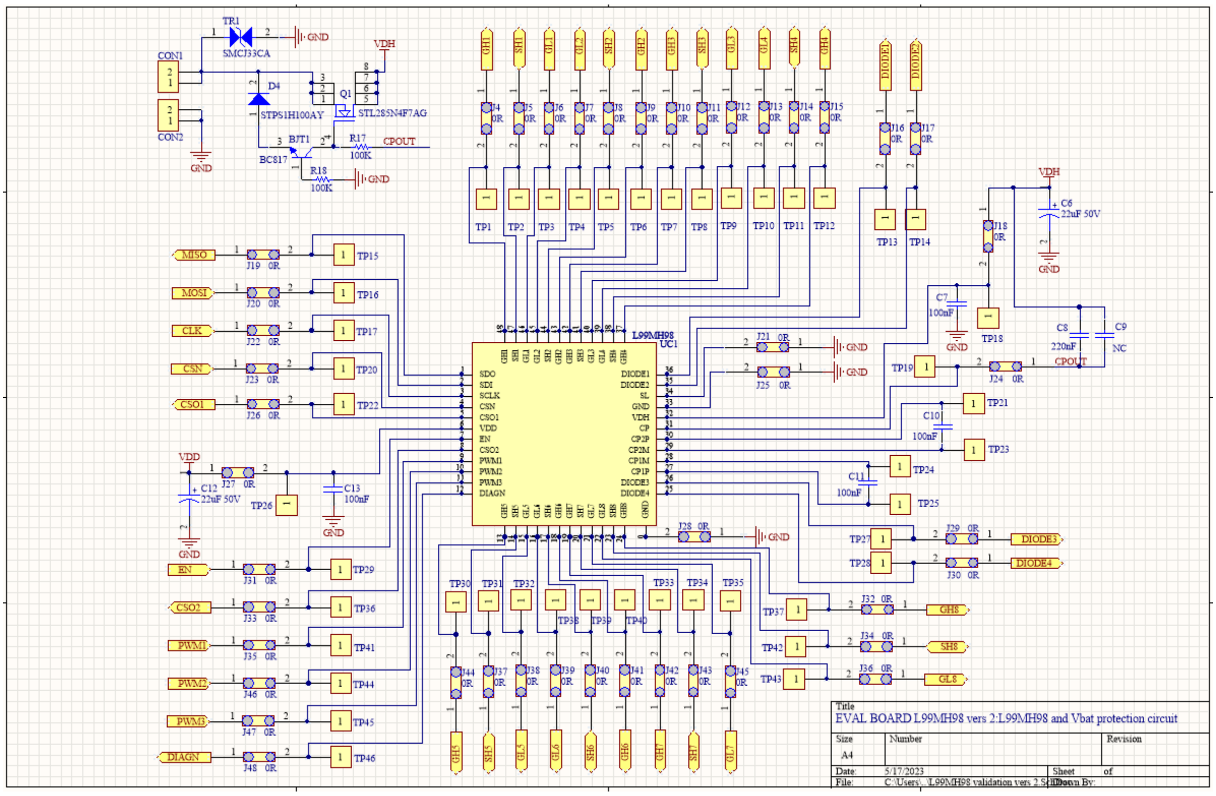
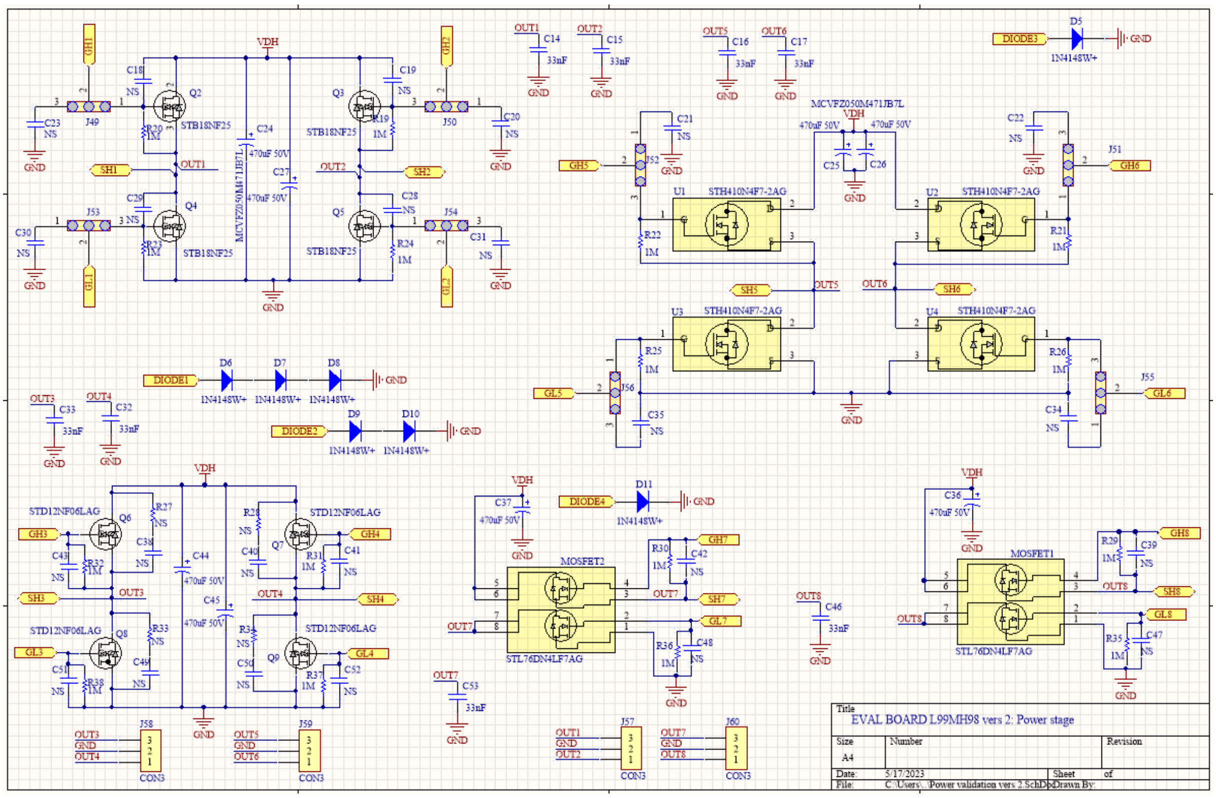


Figure 22. Validation board schematic, power MOSFETs section

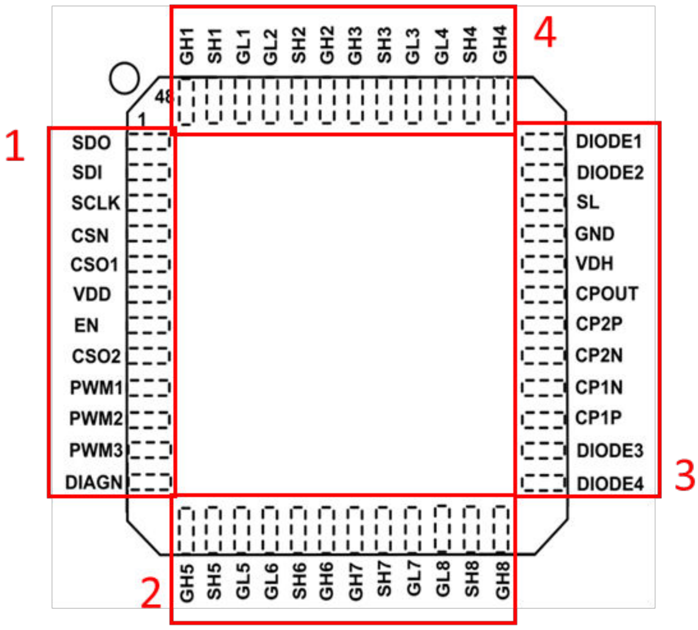


2.2 Layout recommendations

The L99MH98 pinout (Figure 23) was designed and thought to make the layout of the board as simple as possible. Four different zones can be identified in the pinout:

1. Zone 1, pins from 1 to 12: Digital pins + CSO pins + Vdd were inserted in this part. This is the part of the device that communicates with the microcontroller that controls the L99MH98.
2. Zone 2, pins from 13 to 24: all the pins (GH5-8 + SH5-8 + GL5-8) to control four half-bridges.
3. Zone 3, pins from 25 to 36: CP pins + DIODE1-4 pins + VDH + GND + SL were inserted in this part. In this part of the L99MH98 there are different functions: the pins to connect the charge pump capacitors, the battery and ground connections and the SL pins for the connection of the sources of the LS.
4. Zone 4, pins from 37 to 48: all the pins (GH1-4 + SH1-4 + GL1-4) to control four half-bridges.

Figure 23. **L99MH98 pinout**

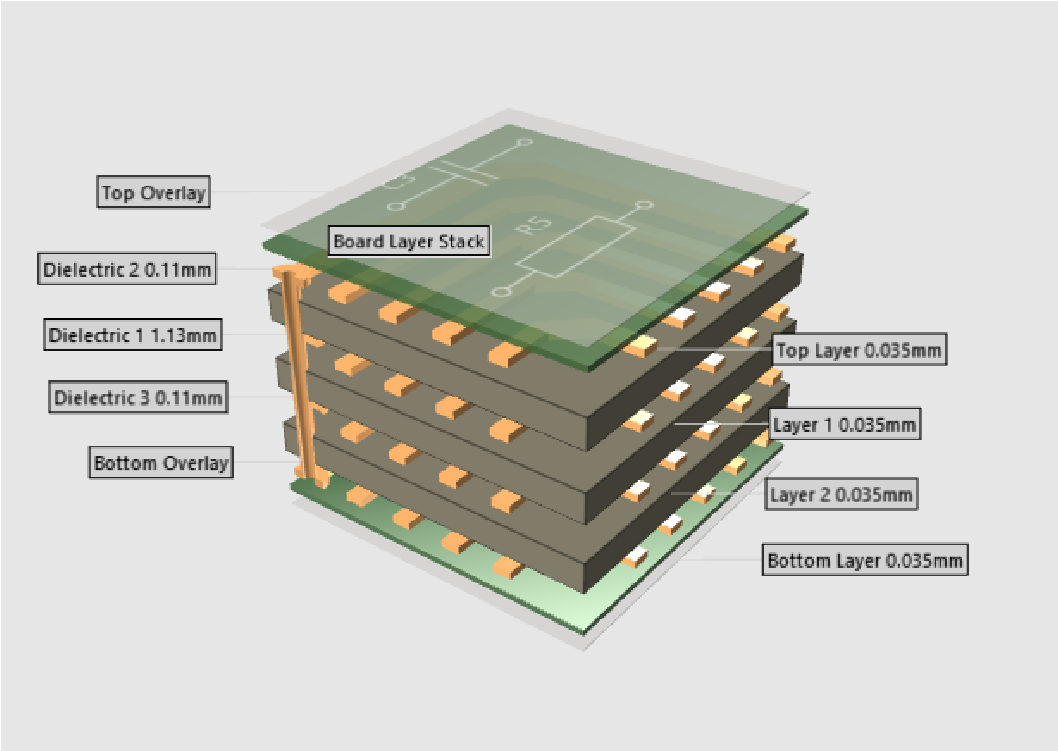


The board to evaluate L99MH98 was designed in 4 layers, FR4 standard. Figure 24 shows the stack-up layer of the application board. The layers are divided as follows:

1. Top layer: used to route the main signals as tracks for large currents
2. Layer 1: used as GND layer
3. Layer 2: used as VDD and VDH layer
4. Bottom layer: used to route signals layer

A coplanar waveguide (CPW) with ground structure has been used to design the board.

Figure 24. L99MH98 application board stack-up layer



When designing a motor control application board, one of the most important aspects is building an excellent ground. In a four layers board is not recommended to use tracks as ground, but it is necessary to use ground plane.

In the L99MH98 application board the ground plane solution is used: the Layer 1 (Figure 26) is used as ground layer. In the top (Figure 25) and bottom (Figure 28) layers, some ground islands are present also. Having one layer as a continuous ground plane lets each signal have the shortest return path and decrease coupling and interference. It is important to prevent breaks in the ground plane, so the best solution is avoiding tracks.

The ground planes present in the different layers must be electrically connected: vias are used to connect planes on multiple layers. Each via has a maximum current it can handle: the vias that have been used in the L99MH98 application board have a hole diameter of 12 mils and have a capability up to 0.84 A. To work with much larger currents, such as those flowing in a motor (tens of amperes), the multi-via solution is used: the planes present between the various layers relate to many vias. The multi-via solution will not only allow large currents to flow between the grounds in the various layers but it will also allow for uniform voltage across the entire ground of the board.

The indirect current measurement system uses the measurement of the Vds (drain source voltage) of each MOSFET. To do this, 16 operational amplifiers have been inserted inside the device: 8 operational amplifiers measure the voltage across the 8 HS, measuring the voltage between Vdh and SHx pins. The other 8 op amps measure the voltage between the SHx pins and the SL pin.

The pin of SL must be connected to the ground of the board. It is, therefore, very important to have a good ground plan on the board (as described above) and to connect the SL pin to the board ground in the best possible way.

Another important point is the connection of the Vdh pin on L99MH98. It is important that the Vdh pin of the device is at the same potential as the Vdh pin of the mosfet to be measured. To do this it is recommended to use Vdh planes in order to minimize the path resistance.

Figure 25. L99MH98 application board top layer

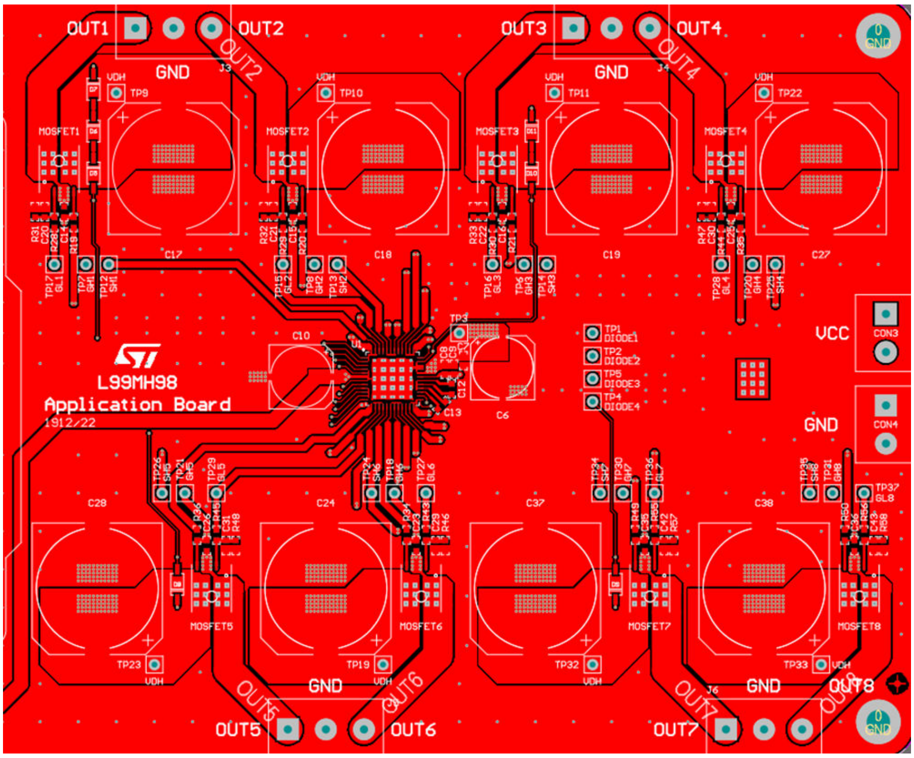
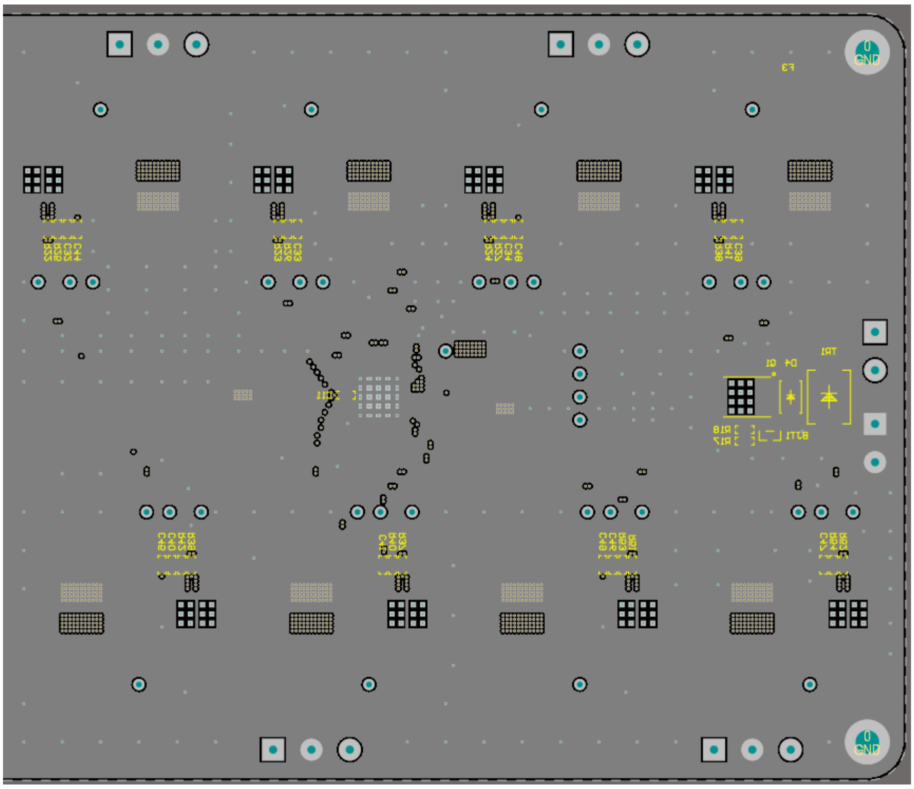


Figure 26. L99MH98 application board Layer 1 (GND layer)



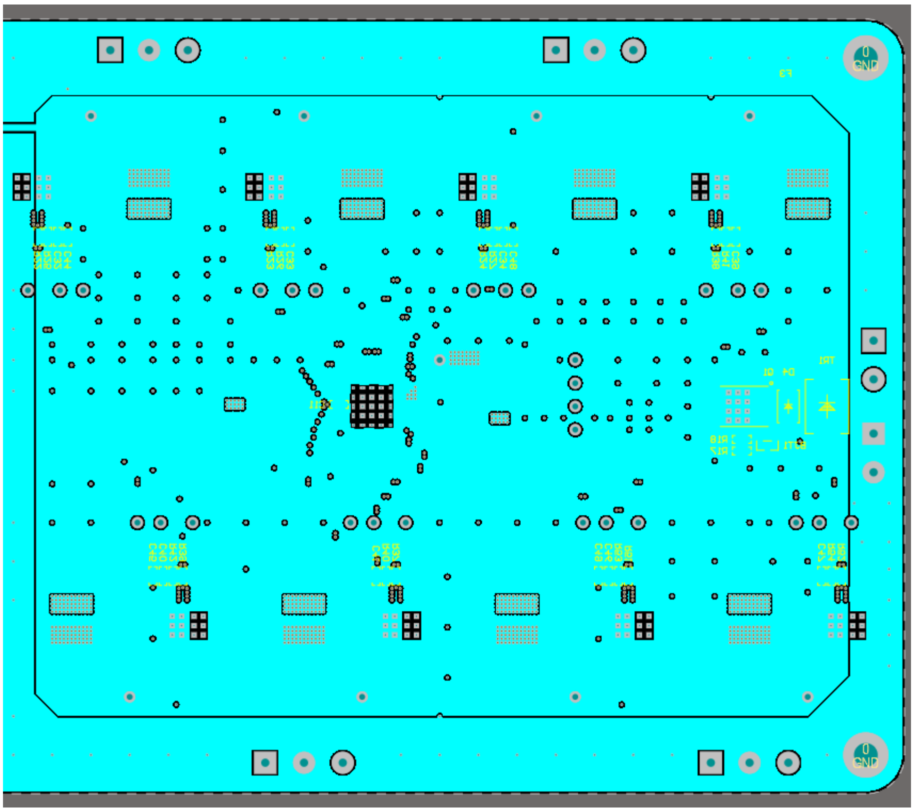


Figure 27. L99MH98 application board Layer 2 (Vdh layer)

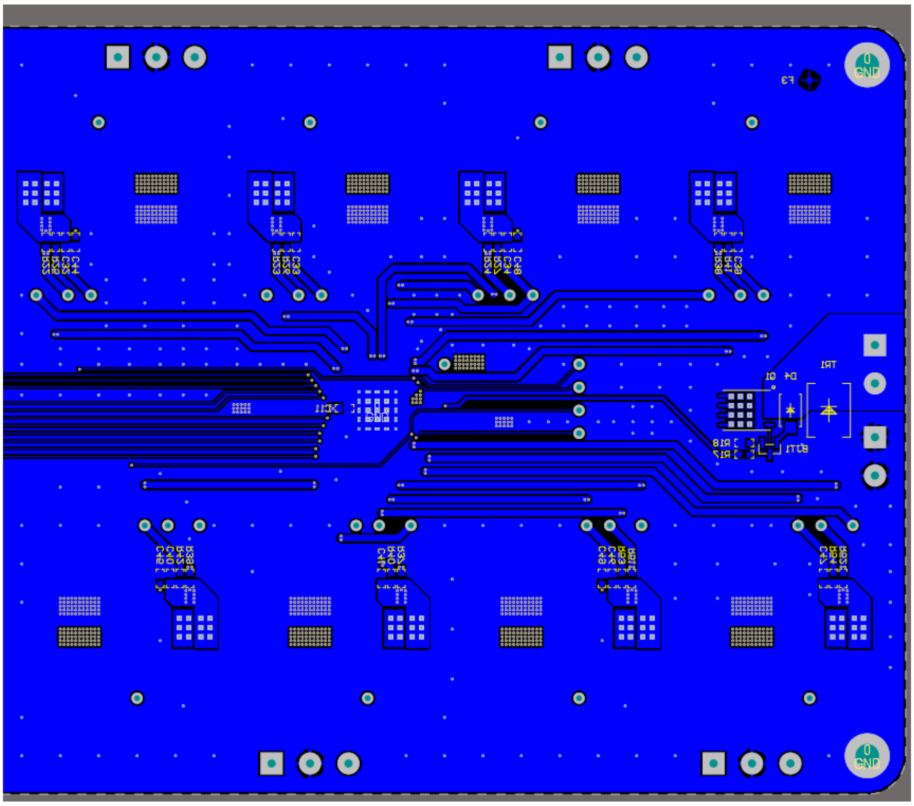


Figure 28. L99MH98 application board bottom layer

Revision history

Table 13. Document revision history

Date	Version	Changes
09-May-2025	1	Initial release.

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