Abstract

The high working junction temperature of power components in vehicle’s Electronic Control Unit (ECU) is the most common reason for their failure. The peak junction temperature under a pulsed condition could be substantially different from that at the steady-state condition. An accurate junction temperature prediction plays an important role in the design and optimization of power devices. In this paper, a transient temperature prediction model (TTPM) for power components in ECU is proposed. The measured temperature-rise results by experiments are provided to verify model prediction accuracy. The prediction results match the experimental results very well. Although temperature prediction is based on the solenoid valve drive circuit, the analysis results can be easily extended to other power device drive circuitry topology.

Keywords: Thermal analysis, Temperature prediction, Model, Power loss, Transient thermal resistance

1. Introduction

The number of electrical devices in modern vehicles has been rapidly increasing over the last two decades. As a crucial part of the modern automobile, power-flow-regulating devices in the form of power electronics are indispensable in a wide range [1-2]. The Power MOSFET, as one of the most important power electronics, is widely used in the area of motor drives control, large switching mode power supplies, dc-dc converters, and engine solenoid valve drive, due to their lower drive requirement and fast switching speeds [3-4]. More than 60% of electric power in the vehicles today is processed through some forms of Power MOSFETs [5].

Ideal electronic component without power consumption does not exist in the real world. Every power component will consume some power and the consumed power will be transformed into heat which is the main cause for its temperature-rise. Excessive heat may degrade the performance of the device by lowering its operating speed, and in the worst case, damage the device, rendering it inoperable. The reliability of operation and the life expectancy of electronic components reduce exponentially as the junction operating temperature rises [6]. The high working junction temperature of power components are the most common reasons for their failure. As such, small differences in operating temperature (order of 10-15°C) can result in approximately two times the difference in the lifespan of the components [7]. Reliability and performance analyses in the design process necessitate the need for accurate prediction of electronic device junction temperatures.

The peak junction temperature under a pulsed condition could be substantially different from that at the steady-state condition [8-9]. In order to predict the accurate peak junction temperature, transient thermal analysis is very necessary. The problem of transient thermal analysis appears very early in literature [10]. Transient thermal analysis is performed analytically using asymptotic functions and the convolution integral in Refs [11-12]. In these works, the analogy between heat conduction and electrical conduction in an RC network is stated in a way that allows modeling the thermal system with lumped thermal resistances and capacitances. A classical example is the work presented in Ref [13-14] on the synthesis of a dynamic model consisting of a three-stage RC ladder network. However, the aforementioned analysis works are all focus on the component’s temperature-rise response excited by a given power loss which has been known in advance.

In order to investigate the temperature-rise performance of the power circuit, a practical temperature prediction model is very necessary before the hardware has been built without expensive prototype
building and testing. Based on some typical parameter values which were easily acquired from datasheet of MOSFET, the model can predetermine the degree of temperature-rise to keep the junction temperature below the maximum admissible security value.

In this paper, a transient temperature prediction model (TTPM), which is composed by the power loss model and the corresponding thermal resistance model, is proposed in section 2. A common rail fuel injector solenoid valve driving circuit topology is chosen as an application example for the proposed model in section 3. When the mathematical model is used as a tool for thermal analysis, experimental validation is always required to validate the accuracy of the mathematical model in section 4. Based on the drive circuitry structure parameters and operation conditions, the model can predict the junction temperature of Power MOSFET. Since device selection and circuit design are normally carried out exclusively on datasheet information provided by manufacturers, this is a very valuable model for practical use.

2. The theory foundation of model

2.1. Transient Thermal Resistance Model

The basic principles of thermal analysis are similar to those in the electrical domain. In electrical domain, a material could offer resistance to the flow of current when there is voltage difference at two ends of it; it also can offer resistance to the flow of power in thermal domain if temperature difference exists. Resistance to heat (power) flow is called thermal resistance. Just as the familiarity of electrical resistance definition \( R = \frac{V}{I} \), thermal resistance can be expressed with Equation 1:

\[
R_{th} = \frac{\Delta T}{P}
\]  

where \( \Delta T \) is temperature difference and \( P \) is consumed power. Heat is generated in the central junction position of the power device and released to surrounding environment through thermal resistance path.

Fig. 1. Steady-state thermal resistance model of a Power MOSFET mounted on a printed circuit board.

Both the concept of thermal resistance and capacitance and the idea that the transient behavior of any packaged component can be expressed by a set of exponential terms comes from the well established thermal-electrical analogy [15-16]. Equivalent circuit models of thermal systems are still a convenient method for thermal analysis [17-21]. Fig. 1 shows an engine electrical control unit with several Power MOSFETs mounted on the surface of PCB and the corresponding steady-state thermal resistance model. The popularity of this method is probably due to the familiarity of circuit analysis to electrical designers. In thermal analysis, temperature and heat flow are analogous to voltage and current, respectively.

In the case of a vertical Power MOSFET, where the thickness is small compared to other dimensions, heat is generated at the top surface of junction silicon and flows perpendicularly to the silicon surface through the die to the die attach pad, through the lead frame to the surrounding case material, to the mounting-base printed circuit board, and eventually to the surrounding ambient environment. The temperature of mounting-base is assumed to be equal to the PCB. Convection and radiation are assumed to be negligible. So, a one-dimensional (1-D) heat transfer may be considered.

As has been aforementioned, the rise of MOSFET junction temperature is due to the consumed power and associated thermal resistance. Thus the junction temperature rise above the mounting-base is directly proportional to the dissipated power and the junction to mounting-base thermal resistance.
According to the definition of thermal resistance, junction temperature-rise \( T_j \) of MOSFET can be written as:

\[
\Delta T_j = T_j - T_{amb} = P \times R_{thj-mb} \tag{2}
\]

where \( P \) is consumed power, \( R_{thj-mb} \) is thermal resistance from junction to mounting-base, while \( T_{amb} \) demonstrates the mounting-base temperature.

Steady-state operating junction temperature for a circuit under DC or very low frequency conditions can be computed through use of above described thermal resistance. But it is not satisfactory for finding transient peak junction temperature for pulsed applications. Because of the heat storage effects of semiconductor material, the critical junction temperature will not be reached instantaneously, even when excessive power is being dissipated in the device. The actual junction temperature will be overestimated if plugging in the peak power while using the average power value will underestimate the peak junction temperature at the end of the power pulse. The reason for the discrepancy lies in the thermal capacity of the semiconductor and its ability to store heat and to cool down before the next pulse.

In order to estimate peak junction temperature for pulsed application, thermal capacitance must be taken into consideration. The corresponding simple dynamic thermal model [18-21] for the MOSFET, which exactly corresponds to the thermal characterization provided by most device manufacturers, is shown in Fig. 2. The normally distributed thermal capacitances have been lumped into single capacitors labeled \( C_j \) and \( C_{mb} \). This model represents the junction temperature in terms of the static junction to mounting-base thermal resistance and the thermal time constant \( \tau_\theta = R_{thj-mb} \times C_{\theta} \).

\[
\text{Fig. 2. Transient RC thermal resistance model corresponds to steady-state model}
\]

The existence of thermal capacitance results in thermal responses like electrical RC responses which we are familiar with in the electrical domain. When a step pulse of heating power \( P \) is introduced at the junction, junction temperature \( T_j \) will rise at an exponential rate to a certain steady-state value; naturally, when the input power is terminated, junction temperature will decrease along the similar exponential slope rate curve back to its initial value [21-23].

\[
T_j = R_{thj-mb} \times P \times \left( 1 - e^{-\frac{t}{\tau_\theta}} \right) + T_{amb} \tag{3}
\]

Transient thermal resistance could be obtained from inspecting equation 4:

\[
Z_{thj-mb}(t) = R_{thj-mb} \times \left( 1 - e^{-\frac{t}{\tau_\theta}} \right) \tag{4}
\]

Initially, assuming the time approaches to zero and is far less than time constant, the thermal resistance could be negligible. Taking the limit of the first-order differential of the transient thermal impedance equation gives:

\[
\lim_{t \to +\infty} \frac{dZ_{thj-mb}(t)}{dt} = \frac{\Delta Z_{thj-mb}}{\Delta t} = \frac{R_{thj-mb}}{\tau_\theta} \tag{5}
\]

On the contrary, as the time passes and is far greater than thermal time constant, transient thermal effects will die out; transient thermal resistance will increase until it eventually reaches the steady−state thermal resistance \( R_{thj-mb} \).

\[
\lim_{t \to +\infty} Z_{thj-mb}(t) = R_{thj-mb} \tag{6}
\]

The practical method of handling the transient thermal problem is to measure the thermal response of the semiconductor to various pulses of input power. Fortunately, almost every power
semiconductor’s datasheet will provide the transient thermal response chart, on which the thermal resistance against pulse duration, often accompanying a curves family of “duty cycle” will be plotted. Fig. 3 shows a typical curves family of NXP Semiconductor BUK9675-100A for thermal impedance against pulse duration, with duty factor as a parameter. From the transient thermal response chart, the steady-state thermal resistance can be directly read out; afterwards the thermal time constant $\tau_n$ can be acquired in single-pulse heating curve ($\delta = 0$) according to Equation 5 through differentiating transient thermal resistance $\Delta Z_{th}/\Delta t$.

Thus, the transient thermal resistance under different pulse conditions can be easily approximated by the following expression [23], the required parameter value can be easily obtained from published datasheet of power device.

$$Z_{th}(t, \delta) = \delta \times R_n + (1 - \delta) \times Z_n \left( t + \frac{t}{\delta} \right) + Z_n(t) - Z_n \left( \frac{t}{\delta} \right)$$

(7)

Fig. 3. Transient thermal impedance curves family as a function of pulse duration

2.2. Power Loss Model

Without loss of generality, two major power losses of Power MOSFET are conduction loss and dynamic switching loss. The on-resistance of Power MOSFET dissipates most conduction power while the dynamic switching loss is mainly produced during every switching cycle. The switching losses are made up of a series of pulse pairs (the first one refers to the device turn-on while the second one to the device turn-off) having a large amplitude but a relatively short duration. A classical piecewise-linear analytical loss model [24-25] has been widely used due to its simplicity and good performance. The model treats the switch turn-on and turn-off waveforms as piecewise linear. Fig. 4 shows an approximated piecewise linear loss model of Power MOSFET.

During the turn-on process, the MOSFET gate is connected to gate charge voltage $V_{GS}$ starts to increase until it reaches gate-source threshold voltage $V_{th}(a)$ at $t_1$ and there are no switching losses in this period. At time $t_1$, the drain current $I_D$ begins to rise linearly and the gate-source capacitance $C_{GS}$ starts to charge. During the period $t_1$ to $t_2$, $C_{GS}$ continues to charge, the gate voltage continues to rise and the MOSFET is sustaining the entire input voltage across it. At time $t_2$, $C_{GS}$ is completely charged and the drain current reaches the predetermined current $I_p$ and stays constant while the drain voltage starts to fall linearly. At this time, the drive current starts to charge the Miller capacitance $C_{DG}$, this continues until time $t_3$. During this period the current is constant (at $I_p$) and the supply voltage is falling linearly from $V_{ds}$ to drain-source on-state voltage $V_{ds(on)}$ which is equal to $R_{ds(on)} \times I$ due to on-resistance. Actually, charge time for the Miller capacitance is very larger than that for the gate to source capacitance $C_{GS}$ due to the rapidly changing drain-source voltage between $t_1$ and $t_2$. Once both
of the capacitances $C_{gs}$ and $C_{gd}$ are fully charged, gate voltage $V_{gs}$ starts to increase again until it reaches the gate supply voltage at time $t_4$. The turn-off process is just opposite.

Fig. 4. Piecewise linear approximation of power loss model

Integrating the product of current and voltage over the switching interval yields the instantaneous switching power losses. The instantaneous product of current and voltage results in large values of instantaneous energy dissipation.

$$E_{\text{on-off}} = \frac{1}{2} (V_{ds} \times I_d \times t_{\text{turn-on}})$$

$$E_{\text{off-on}} = \frac{1}{2} (V_{ds} \times I_d \times t_{\text{turn-off}})$$

where $V_{ds}$ is the supply voltage, $I_d$ is the load drive current. Approximating drain-source voltage falling linearly from $V_{ds}$ to 0 simplifies the calculation and introduces no significant error because the switching time is very small. The turn-on time $t_{\text{turn-on}}$ and turn-off time $t_{\text{turn-off}}$ is determined by how long it takes the gate driver to deliver the required charge to switch the device.

$$t_{\text{turn-on}} = \frac{Q_g}{I_{\text{drive}}}$$

$$t_{\text{turn-off}} = \frac{Q_g}{I_{\text{drive}}}$$

The driver current $I_{\text{drive}}$ relies on the selected driver chip and drive circuit construction. The gate charge $Q_g$ can be directly read from MOSFET datasheet, if it is not specifically specified in datasheet, it can also be read from the gate charge graph. The advantage of using gate charge is that the designer can easily calculate the time required to switch the device.

The product of the on-state voltage of the MOSFET and the load current during the on time is the conduction energy loss.

$$E_{\text{cond}} = I_d \times R_{\text{DS(on)}} \times t_{\text{cond}}$$

where $R_{\text{DS(on)}}$ is the drain-source on-state resistance and $t_{\text{cond}}$ is the turn-on time of MOSFET.

Thus the total power loss during every switching process of Power MOSFET can be equivalent to the following:

$$P_{\text{total}} = \frac{E_{\text{on-off}} + E_{\text{cond}} + E_{\text{off-on}}}{t_{\text{turn-on}} + t_{\text{cond}} + t_{\text{turn-off}}}$$

2.3 Temperature Predicting Model

From the view of thermal design reliability, it is most important to know what the transient peak junction temperature will be when a series of power pulses waveform is applied. Peak junction
temperature will usually occur at the end of the applied pulses and its calculation will undoubtedly involve instantaneous power dissipation and transient thermal impedance.

While predicting peak junction temperature in response to a series of power pulses becomes very complex, superposition \([23, 26-27]\) of power pulses offers a rigorous numerical method of using the equivalent rectangle power pulse and the transient thermal resistance to secure a solution. The superposition principle allows the summation of the individual power pulse effects to calculate the resulting junction temperature. Power inputs may be thought of as combinations of step inputs, and the response of the system is the sum of the step inputs.

\[
\Delta T(t) = \sum_{i=1}^{n} P_i \times Z_a(t - t_i, \delta)
\]  

(14)

where \(P_i\) is the input power, \(t_i\) is the time step input, \(Z_a(t - t_i, \delta)\) is transient thermal resistance of different duration \(t\) and duty factor \(\delta\), \(n\) is numbers of power contributions. In practice, a power device frequently has to borne periodic pulse waveforms. This type of signal can be simulated by superimposing several rectangular pulses which have a common period, but both positive and negative amplitudes, in addition to suitable values of pulse duration and duty factor. The positive pulse occurs at the start of the pulse, with a magnitude equal to that of the pulse. The negative pulse occurs at the end of the pulse, with a magnitude opposite that of the pulse.

In order to apply the superposition thermal calculations to non rectangular power waveforms, a waveform that is not square can be approximated by a rectangle wave. The equivalent rectangle must be equal in area to the section of the waveform it represents (i.e. the same energy) and also be of the approximate similar peak power, through adjusting the duration. As long as the total power remains constant, the calculated results will be similar.

For a series of pulses, the junction temperature at the end of the \(n\)th pulse is the algebraic sum of all preceding pulse contributions. To be able to add the various effects of the pulses at this time, all the pulses, both positive and negative, must end at this time. Positive pulses increase the junction temperature while negative pulses decrease it.

\[
\Delta T(t_{\text{end}}) = \sum_{i=1}^{n} P_i \left[ Z_a(t_{\text{end}} - t_{i,\text{start}}) - Z_a(t_{\text{end}} - t_{i,\text{end}}) \right]
\]  

(15)

(a) Power dissipation                        (b) Equivalent power input by a set of step function

(c) Junction temperature response to individual pulse        (d) Use of superposition to determine peak \(T_j\)

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Ren Guofeng, Tian Feng, Yang Lin
The corresponding transient thermal resistance of different duration and duty factor

Fig. 5 illustrates a typical example for how the superposition principle is used to calculate the peak junction temperature when four power pulses are applied, the peak junction temperature-rise at the end of the fourth pulse can be summed as follows:

\[
\Delta T(t_{4,\text{end}}) = \Delta T(t_1) = \sum_{i=1}^{4} P_i \cdot Z_{th}(t_{i,\text{end}} - t_{i,\text{start}}) - Z_{th}(t_{i,\text{end}} - t_{i,\text{start}})
\]

\[
= P_1 \cdot Z_{th}\left(t_1 - t_0, \frac{t_2 - t_0}{t_2 - t_0}ight) - P_2 \cdot Z_{th}\left(t_1 - t_2, \frac{t_2 - t_1}{t_2 - t_1}\right)
+ P_2 \cdot Z_{th}\left(t_1 - t_2, \frac{t_3 - t_1}{t_3 - t_1}\right) - P_3 \cdot Z_{th}\left(t_1 - t_3, \frac{t_3 - t_2}{t_3 - t_2}\right)
+ P_3 \cdot Z_{th}\left(t_1 - t_3, \frac{t_4 - t_1}{t_4 - t_1}\right) - P_4 \cdot Z_{th}\left(t_1 - t_4, \frac{t_4 - t_3}{t_4 - t_3}\right)
\]

3. Application example

The transient temperature prediction model (TTPM) proposed in this paper is a general model for predicting the peak junction temperature of power devices as long as the transient thermal impedance curve families, which is similar with Fig. 3, is provided in manufacturer’s data books.

In this section, a common rail fuel injector solenoid valve driving circuit topology is chosen for the present analysis example; the framework of driving circuitry is depicted in Fig. 6(a). The six solenoid valves are placed in parallel connection with each other, each corresponding to one cylinder injector. Every solenoid valve is connected in serial with a power control switch which will be turned on in sequence of cylinder firing order via low side selecting switch pulse signal (LS_Switch_x). The control switch BUK9675-100A from NXP semiconductor is employed in driving circuitry. The driving circuitry features Peak & Hold two-stage constant current control function and dual-voltage driven capability. An initial larger current driven by a high supply voltage is applied to the solenoid coil to provide the necessary strong electromagnetic force to pull the plunger in more quickly; the high supply voltage is turned off immediately as soon as the peak current is reached, and the peak current will last for a constant 500\text{us} to ensure the plunger is reliably seated, when the plunger is seated, a relative smaller current (called hold current) driven by low supply voltage is sufficient to maintain the plunger in the seated position. The dual-voltage is composed of low supply voltage which is connected to battery directly and high supply voltage which is derived from low supply voltage through a high efficient DC/DC convertor. The related timing signals and waveforms during single injection pulse are also illustrated in Fig. 6(b).

In order to get a constant drive current, the PWM method is adopted to frequently switch the control MOSFET; this will lead to two major power loss of Power MOSFET, conduction loss and dynamic switching loss. The power dissipation will be transformed to heat and becomes the main cause for its temperature-rise.
The power dissipation of each Power MOSFET during single injection pulse is proportional to the injection pulse width. Each cylinder of engine completes one fuel inject during every work-cycle. So the fuel injection frequency is also proportional to the engine speed. The injection pulse width is decided by fuel injection amount which is decided by accelerator pedal position (ACCP) and engine speed at certain fuel injection pressure, it is a function of ACCP and engine speed, the relation map graph about ACCP and engine speed is depicted in Fig. 7.

As can be seen from the timing signal of Fig. 6(b), the low side power switch MOSFET will be switched more times during single injection pulse, unlike high side switch will be turned on only once, so it will always dissipate more power compared to the high side power switch, thus its operating temperature will be higher than high side MOSFET. As the six low side switches operate under the same electrical and thermal stress, only one low side switch is chosen for the temperature-rise analysis (in the following LS_Switch_1 is chosen).

In order to apply the above TTPM to predict the peak junction temperature, the switching number $n$ must be calculated out in advance, and then the transient thermal resistance under different duration and duty factor could be obtained successively. This is also considered in the Peak & Hold two-stage constant current waveform shown in Fig. 8. Every ripple of modulated current waveform is accompanied a switching process of the Power MOSFET.
When the high side voltage supply control switch and the low side cylinder selecting control switch are turned on simultaneously, the voltage equation of driving circuitry can be expressed by equation 17.

\[ U = i \times R + L \times \frac{di}{dt} \] (17)

The magnetizing current of solenoid valve during power-up and power-down stages are solved through equation 18 and equation 19 respectively.

\[ I_{\text{power-up}} = \frac{U - e^{- \frac{R_{up}}{L} \cdot \ln(U - I_0 \times R_{up})}}{R_{up}} \] (18)
\[ I_{\text{power-down}} = I_0 \times e^{- \frac{R_{down}}{L}} \] (19)

where \( U \) is magnetizing voltage, \( I_0 \) is initial current value when current begins to rise or fall, \( L \) is inductance value of solenoid valve, \( R_{up} \) and \( R_{down} \) are the electrical resistance values of current loop path, including solenoid resistance, on-resistance of MOSFET, shunt resistance \( R_{\text{shunt}} \).

Based on equation 18 and equation 19, if the current ripple \( \Delta I \) is predetermined, the turn-on time \( t_{on} \) for magnetizing current rising to peak-threshold from 0 and the turn-off time \( t_{off} \) from peak-threshold to valley hold-threshold can be solved:

\[ t_{on} = \frac{L}{R_{up}} \times \ln \left( \frac{U}{U - R_{up} \times I_{peak}} \right) \] (20)
\[ t_{off} = \frac{L}{R_{down}} \times \ln \left( \frac{I_{peak}}{I_{hold} - \Delta I} \right) \] (21)

The rising time \( t_r \) for current rising from valley-threshold to peak-threshold and the falling time \( t_f \) for current falling from peak-threshold to valley-threshold can also be solved:

\[ t_r = \frac{L}{R_{up}} \times \ln \left( \frac{U - R_{up} \times I_{th}}{U - R_{up} \times (I_{th} - \Delta I)} \right) \] (22)
\[ t_f = \frac{L}{R_{down}} \times \ln \left( \frac{I_{th}}{I_{th} - \Delta I} \right) \] (23)

where \( I_{th} \) means current threshold in peak or hold stage. For an injection pulse width \( T \), the number \( n \) of switching MOSFET can be given by:

\[ n = \frac{500 \mu s - t_{on}}{t_r + t_f} + \frac{T - 500 \mu s - t_{off}}{t_r + t_f} \] (24)

The power dissipation of LS_Switch_1 depends on injection pulse width and engine speed, it will consume maximum power according to 100% engine load line (100% accelerator pedal position, shown in Fig. 7) as it always has maximum injection pulse width under different engine speed. The
required parameter of proposed model includes power loss part and thermal resistance part; the input parameter values for the model are shown in table 1.

<table>
<thead>
<tr>
<th>Table 1. The input parameter for the proposed model</th>
</tr>
</thead>
<tbody>
<tr>
<td>Power loss model</td>
</tr>
<tr>
<td>Drain-source on-state resistance</td>
</tr>
<tr>
<td>Gate charge</td>
</tr>
<tr>
<td>Gate charge current</td>
</tr>
<tr>
<td>Gate charge current</td>
</tr>
<tr>
<td>High supply voltage</td>
</tr>
<tr>
<td>Low supply voltage</td>
</tr>
<tr>
<td>Peak current threshold</td>
</tr>
<tr>
<td>Hold current threshold</td>
</tr>
<tr>
<td>Electrical resistance when power-up</td>
</tr>
<tr>
<td>Electrical resistance when power-down</td>
</tr>
<tr>
<td>Solenoid valve inductance</td>
</tr>
<tr>
<td>Current ripple</td>
</tr>
<tr>
<td>Thermal resistance model</td>
</tr>
<tr>
<td>Steady-state thermal resistance</td>
</tr>
<tr>
<td>Differential term</td>
</tr>
</tbody>
</table>

4. Result and verification

4.1. Experimental Results

When measuring junction temperature of a Power MOSFET, it is basically impossible to put a physical thermometer onto its junction while under power. Instead, we can monitor a temperature sensitive parameter ($TSP$) internal of the component. The inherent body-diode’s forward voltage drop $V_{SD}$ is one of the most commonly used $TSP$s. The forward biased voltage drop of this pn junction has a very linear relationship with temperature [14, 23]; it is approximately 2.39mV/℃ as shown in Fig. 9. So, when this voltage drop is measured out, we can use it to tell us what junction temperature results from any power condition.

![Fig. 9. The body-diode voltage drop versus temperature](image)

![Fig. 10. The measure circuit for temperature verification](image)
A contactless infrared thermometer is an easy-to-use tool for taking temperature measurement. The print circuit board temperature can be easily obtained through it. It is assumed that the mounting-base temperature $T_{mb}$ is equal to the board temperature as they are in close contact with each other. Using a constant voltage supply and a large resistor in series with the power device, a constant current supply is approximately acquired, the current used here should be very low (1.0mA) to prevent significant self-heating. The thermal measure circuit is shown in Fig. 10.

### Table 2. The mounting-base temperature under different engine speed

<table>
<thead>
<tr>
<th>Speed (r/min)</th>
<th>$T_{mb}$ (°C)</th>
</tr>
</thead>
<tbody>
<tr>
<td>300</td>
<td>33.7</td>
</tr>
<tr>
<td>500</td>
<td>34.1</td>
</tr>
<tr>
<td>750</td>
<td>34.3</td>
</tr>
<tr>
<td>1000</td>
<td>35.2</td>
</tr>
<tr>
<td>1250</td>
<td>35.9</td>
</tr>
<tr>
<td>1500</td>
<td>36.3</td>
</tr>
<tr>
<td>1750</td>
<td>36.7</td>
</tr>
<tr>
<td>2000</td>
<td>37.1</td>
</tr>
<tr>
<td>2250</td>
<td>37.6</td>
</tr>
<tr>
<td>2500</td>
<td>38.1</td>
</tr>
<tr>
<td>2750</td>
<td>37.6</td>
</tr>
</tbody>
</table>

The junction temperature-rise $\Delta T_{j-mb}$ at several specific engine speed points (300, 500, 750 ... 2500, 2750r/min) is verified through monitoring TSP of LS_Switch_1, the experimental verification platform is shown in Fig. 11. When the fuel solenoid valve is activated, a series of pulses are applied to the low side power switch LS_Switch_1, the source-drain diode voltage $V_{SD}$ is measured immediately 30us following the end of the power pulses. At the same time, the mounting-base temperature $T_{mb}$ is also measured under different engine speeds and the $T_{mb}$ increases slightly from 33.7°C to 38.1°C with engine speed as is shown in table 2. From the change of the source-drain voltage $V_{SD}$, the print circuit board temperature, and the body-diode voltage relationship with temperature, the junction temperature rise relative to mounting-base temperature can be calculated out:

$$\Delta T_{j-mb} = \frac{0.55 - V_{SD}}{0.00239} + 25 - T_{mb} \text{[°C]}$$

(25)

![Fig. 11. Experimental verification platform](image)

### 4.2. Model Results and Comparison

Based on the proposed model and some conveniently acquired parameters from device datasheets as shown in table 1, the temperature rise of LS_Switch_1 is calculated out at the same engine speed points. Spreadsheet tools are very convenient for this calculation. In order to investigate the predicting accuracy of the proposed model, a lot of comparisons have been carried out. One similar example is provided in Ref [14] which constructs the equivalent thermal model as second order accuracy, but the transient power loss model is not given out, it is assumed that the power loss value has been known in advance. The temperature rise of LS_Switch_1 is also calculated out based on the method provided in Ref [14], the power consumption value using the same results with the proposed model. The experimental measurement results are used as a benchmark for this comparison because they were considered the most accurate device physical characteristics. The comparison result is shown in table 3.

The temperature-rise of LS_Switch_1 at very low engine speed is higher than that in subsequent normal running state because longer injection pulse width will make the transient thermal effects vanished, transient thermal resistance will approach the steady-state thermal resistance, and the longer pulse width will also result in more power dissipation. The temperature-rise does not always increase with the engine speed because the working frequency increases with engine speed, whereas, the injection pulse width decreases with engine speed. The maximum temperature rise appears at
2500r/min engine speed point, so this engine state (2500r/min engine speed, 100% accelerator pedal position) should be chosen for the reference point of the worst thermal analysis.

Compared to the experimental results, the temperature change tendency of calculation shows a good agreement with TSP verification results with the maximum error of 2.3%. The proposed model predicting results are almost identical with the literature model results, but the power loss of literature model must be obtained through other means.

### Table 3. Comparisons of temperature-rise based on different models

<table>
<thead>
<tr>
<th>Speed (r/min)</th>
<th>300</th>
<th>500</th>
<th>750</th>
<th>1000</th>
<th>1250</th>
<th>1500</th>
<th>1750</th>
<th>2000</th>
<th>2250</th>
<th>2500</th>
<th>2750</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Experimental Results (°C)</strong></td>
<td>25.3</td>
<td>25.9</td>
<td>27.6</td>
<td>30.1</td>
<td>33.4</td>
<td>36.3</td>
<td>38.9</td>
<td>41.3</td>
<td>43.7</td>
<td>44.6</td>
<td>44.1</td>
</tr>
<tr>
<td><strong>Literature Model Result (°C)</strong></td>
<td>24.5</td>
<td>26.3</td>
<td>27.1</td>
<td>31.4</td>
<td>34.2</td>
<td>37.2</td>
<td>40</td>
<td>42.5</td>
<td>44.1</td>
<td>45.2</td>
<td>44.5</td>
</tr>
<tr>
<td><strong>Error (%)</strong></td>
<td>-3.2</td>
<td>1.5</td>
<td>-1.8</td>
<td>4.3</td>
<td>2.4</td>
<td>2.5</td>
<td>2.8</td>
<td>0.9</td>
<td>1.3</td>
<td>0.9</td>
<td>1.3</td>
</tr>
<tr>
<td><strong>Proposed Model Result (°C)</strong></td>
<td>25.7</td>
<td>26.4</td>
<td>28.1</td>
<td>30.8</td>
<td>33.2</td>
<td>36.6</td>
<td>38.8</td>
<td>41.8</td>
<td>44</td>
<td>44.7</td>
<td>43.6</td>
</tr>
<tr>
<td><strong>Error (%)</strong></td>
<td>1.6</td>
<td>1.9</td>
<td>1.8</td>
<td>2.3</td>
<td>-0.6</td>
<td>0.8</td>
<td>-0.3</td>
<td>1.2</td>
<td>0.7</td>
<td>0.2</td>
<td>-1.2</td>
</tr>
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### 5. Conclusion

Thermal analysis and thermal design in automotive application has long been a challenging issue as its impact on the component’s reliability and endurance. For Power MOSFETs working under pulse mode, the transient thermal resistance is suitable for its peak junction temperature assessment. Combined with the power loss model and the transient thermal resistance model, the proposed TTPM can predict the peak junction temperature of power component to investigate its reliability and endurance before the hardware circuit has been built, and the required input parameters can be easily acquired from datasheet.

In a specific application example, the temperature-rise of power component at different engine state is calculated out according to the proposed model. The measured results by experiments are provided to verify the model results. A reasonable agreement between the model results and the experimental results is achieved. Although temperature prediction is based on solenoid valve drive circuit, the analysis results can easily be extended to other power device drive circuits such as motor control.

### 6. Acknowledgement

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### 7. REFERENCES


A Transient Temperature Prediction Model for Power MOSFET in ECU
Ren Guofeng, Tian Feng, Yang Lin