International IOR Rectifier

Application-Specific Current Rating of Advanced Power Modules for Motion Control

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Abstract

With the advent of new energy efficiency requirements in motion control applications, 3-phase electric motors have reached broader adoption thanks to the availability of integrated power modules. These modules provide an easy migration path from mechanically commutated motors to electronically controlled motors. However, design engineers need to pay closer attention to the control strategy that affects the current rating requirements of the power elements selected or, more specifically, the integrated power module of choice. This paper illustrates application specific methods of rating advanced multi-chip inverter power modules as a function of motor speed and semiconductor frequency operation. Examples on silicon and (IGBT and FRED) and shunt resistor are illustrated.

Introduction

55% of the total worldwide produced energy is used to run motors, of which, only a small percentage is inverter-driven, while the vast majority use some sort of mechanical control. For this reason the potential for energy saving by the use of electronic (inverter based) regulation is huge: it has been calculated that up to $72B of equivalent electric power can be saved with the adoption of variable speed motion control [1].

In this paper we will concentrate on the appliance and light industrial market, where the adoption of motion control is steadily growing. To help speed up the adoption of electronically controlled motors, the power electronics industry has started to develop a new family of integrated products that goes beyond the concept of a 6-pack power module. By integrating in the same package the inverter bridge and most of the required external control circuitry, systems designers can develop a complete motor drive system with low component count and without - the need of designing the power stage. However, selection of the appropriate power module and definition of operational limits remain in the hands of the system designer. In particular, thermal design still represents the most important factor in the selection of the correct power stage.

In general, designers need to be able to evaluate specific conditions in their application:

?? Power losses

?? Current ratings

?? T_J safety margins

A process flow illustrating the main steps necessary to design and select a power stage are shown in Figure 1. The method shown in this paper streamlines this process in a simple procedure that can be easily automated.

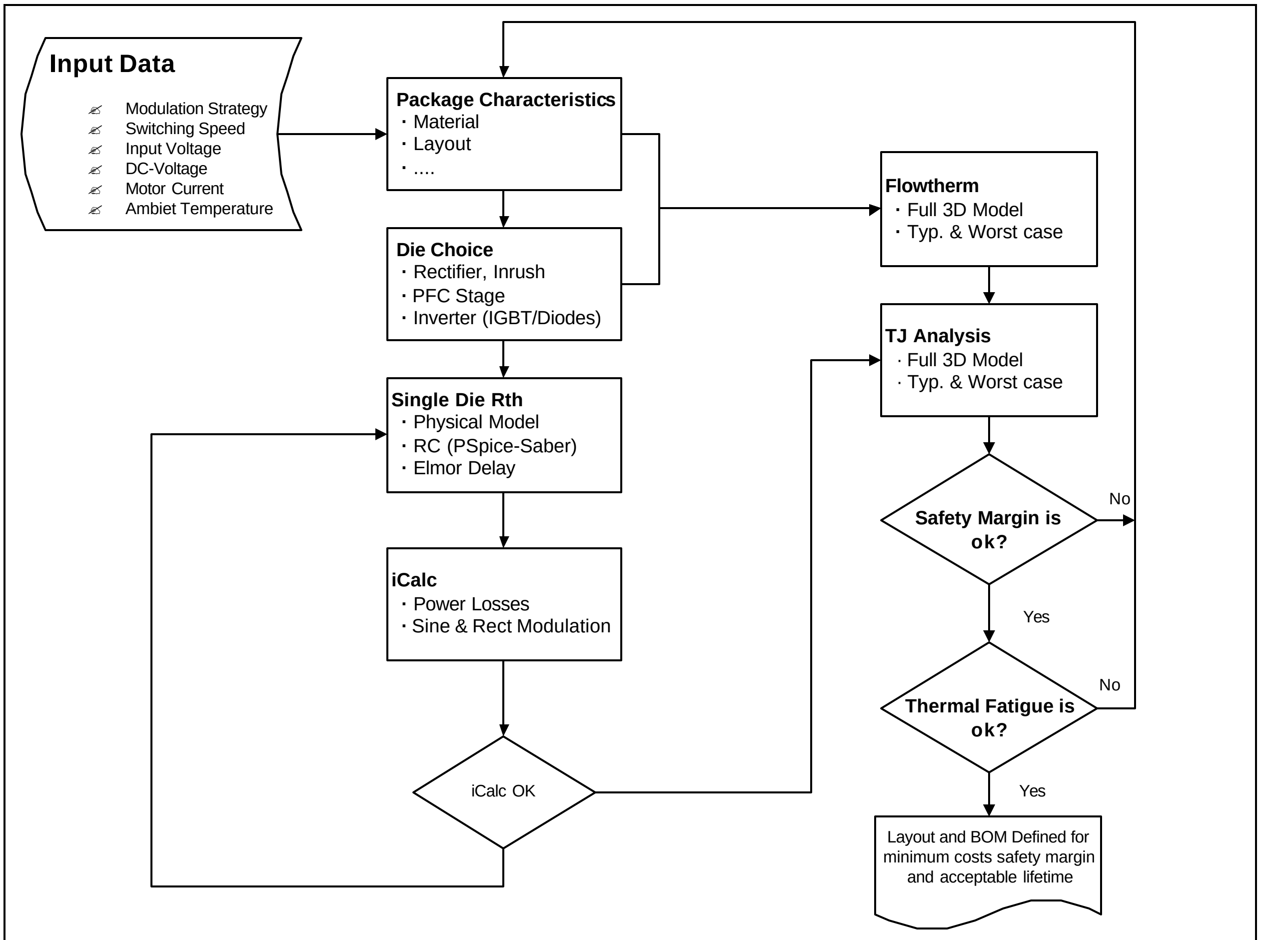

Figure 1 - Power stage design process flow

Electrical model

The complexity of the modulation techniques used in even basic motor drives (PWM, SVM) requires use of simulation tools. The system that needs to be analyzed presents some challenges from the simulation point of view since it includes events that have much different time constants: switching transients in the sub-microsecond range and thermal models in the second range. Spice models, widely available, are based on physical models. These models are not well suited for this type of problem, since they require extremely long simulation times. Another problem with Spice models is the accuracy of power losses calculations, which usually require an even further increase in complexity, modelling time and simulation time to provide accurate results.

An alternative approach to physical models is the use of behavioural models. These models do not try to represent the internal workings of the various silicon devices, but are "black boxes" that have only the input/outputs that are required by system being analyzed

The models in the present papers are used to calculate the conduction losses and power losses with the following equations:

$$V_{CEON} ? V_T ? a.I^b$$

$$V_F ? V_{TD} ? ad.I^{bd}$$

$$E_{ON} ? (h1 ? h2I^x) I^k$$

$$E_{OFF} ? (m1 ? m2I^y) I^n$$

$$E_{DIODE} ? d1.I^{d2}$$

Variation of switching losses with bus voltage is assumed to be linear. In the above equations, parameters V_T, a, b, V_{TD}, ad, bd, h1, h2, x, y, m1, m2, y, n, d1, d2 are extracted with curve fitting methods from measurements done at different currents in the following conditions:

?? V_{BUS}= 400V

?? T_J = 150°C

?? Driver/R_G: internal to the power module

Behavioural models tend to be quite accurate, especially when the conditions are close to the measured conditions. International Rectifier has used these models now for several years in the definition of new products and for the preparation of IGBT product technical documents.

Thermal model

Also for thermal models, a similar trade-off as before can be found. In this case the physical models are based on finite element analysis (FEA) tools, which are accurate but require very long simulation times. The approach used here is again to use a behavioural model that does not contain specific information of the inner workings of the thermal stack.

First, an FEA tool is used to calculate the step response of the power module. The mutual heating of adjacent dies is included by distributing the power losses between the 6 IGBT and 6 diodes of the inverter. Typically this distribution is 85% on the IGBT and 15% on the diode: this is common for the range of power modules considered in this paper.

Second, the normalized curve of the step response, which is the thermal impedance, for the worst IGBT is utilized: the curve is saved by points, and linear interpolation is used when necessary. This effectively represents the thermal behavioural model. An example of thermal impedance curve is shown in Figure 2. This curve is obtained through a FEA tool, but it is always verified experimentally for its steady-state value, in this case 4°C/W.

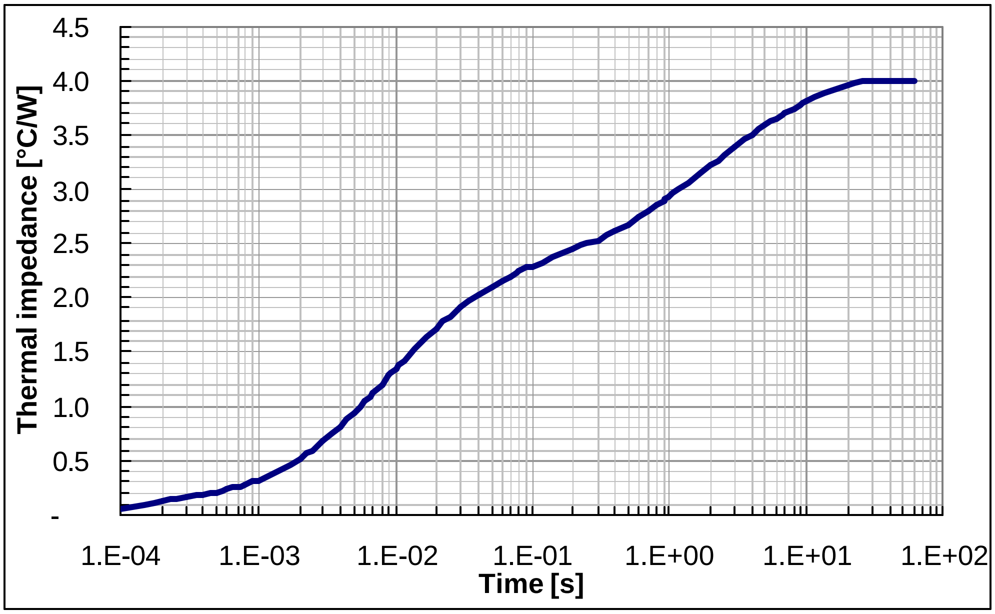

Figure 2 - Thermal impedance curve - IRAMX16UP60A

Sinusoidal approximation method

The purpose of this model is the calculation of power losses and junction temperature. This information is then used to generate the maximum current rating for the part in a specific application.

In an inverter configuration, IGBT and diodes share the current depending on the modulation technique, power factor and modulation index. In the case of pure sinusoidal modulation, a closed form solution can be calculated for the average power dissipated. This approach is valid when the operational frequency of the motor is sufficiently high (>50Hz) that the ripple in junction temperature due to the non-constant power dissipation is negligible. This method has two main limitations:

?? For low operation frequencies (which are common in inverter controlled motors) the ripple in junction temperature is significant and cannot be ignored

?? Different modulation techniques (like space vector modulation) require a complete new set of equations (assuming a closed form solution exists)

In this paper we want to be able to overcome these limitations by calculating the ripple in junction temperature. For these reasons power losses are not calculated with a closed form equation, but the following method is used:

?? The dissipated energy is calculated over half modulation cycle

?? In each switching cycle, the conduction and switching losses are calculated assuming a constant current (inside the switching cycle period)

?? The maximum value of power dissipated in a switching frequency period is saved

As indicated before, the thermal model used in this paper is based on the thermal impedance curve. For this reason, the exact waveform for the dissipated power cannot be used to calculate the junction temperature, because the parameters for the differential equation of the thermal model are unknown. The thermal impedance curve enables us to calculate the maximum junction temperature when the power dissipated in the device is a series of repetitive square pulses.

For this reason the waveform of total power dissipated calculated previously is approximated to a series of square pulses by using the following assumptions:

?? The peak power dissipated in the device is equal to the peak of the square waveform

?? The average power dissipated in the device is equal to the average of the square waveform over this half modulation period

An example of this approximation is shown in Figure 3 for the module IRAMX16UP60A. To verify the effectiveness of these assumptions, an RC ladder model was used for the thermal stack. This model was then solved using Pspice. In Figure 4 we can see the comparison between two runs of this model where we compare the ? T_j with the continuous power input and the square wave input (from Figure 3).

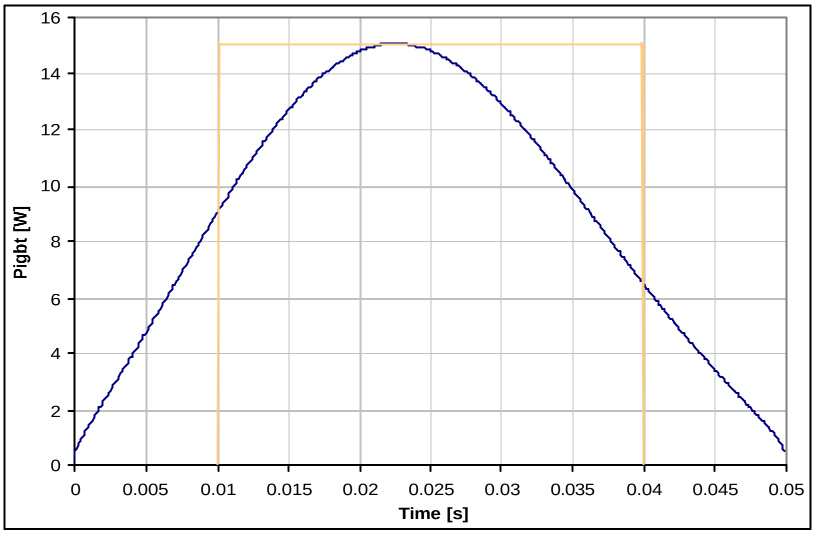

Figure 3 - Power dissipation waveform approximation IRAMX16UP60A; 5Arms; 10kHz; pf=0.6; mi=80%; motor speed= 10Hz

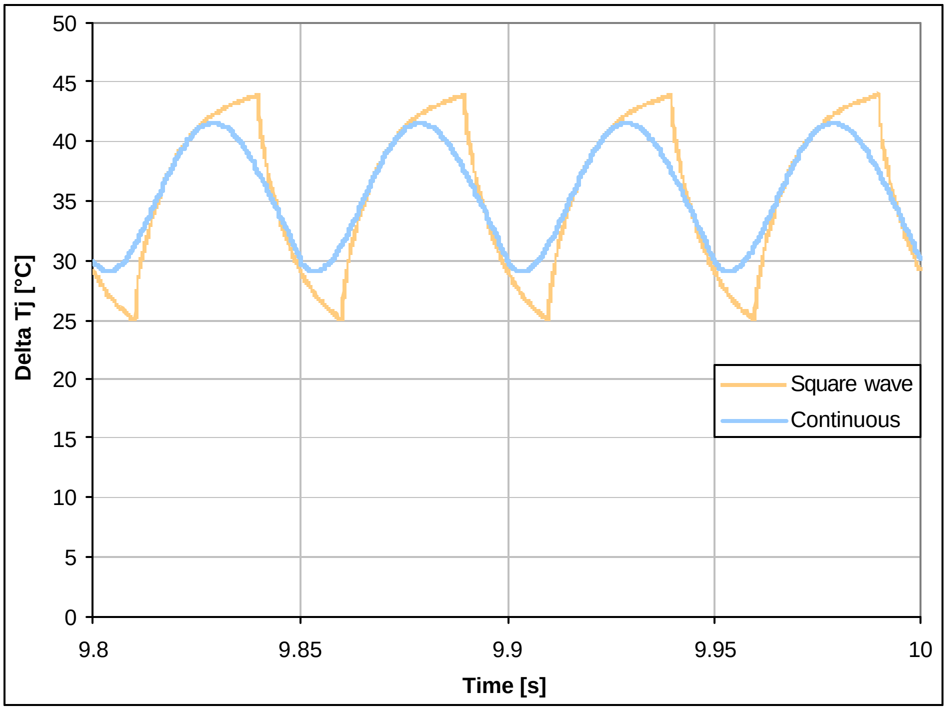

Figure 4 – Effects of power dissipation approximation on junction temperature IRAMX16UP60A @ motor speed= 10Hz

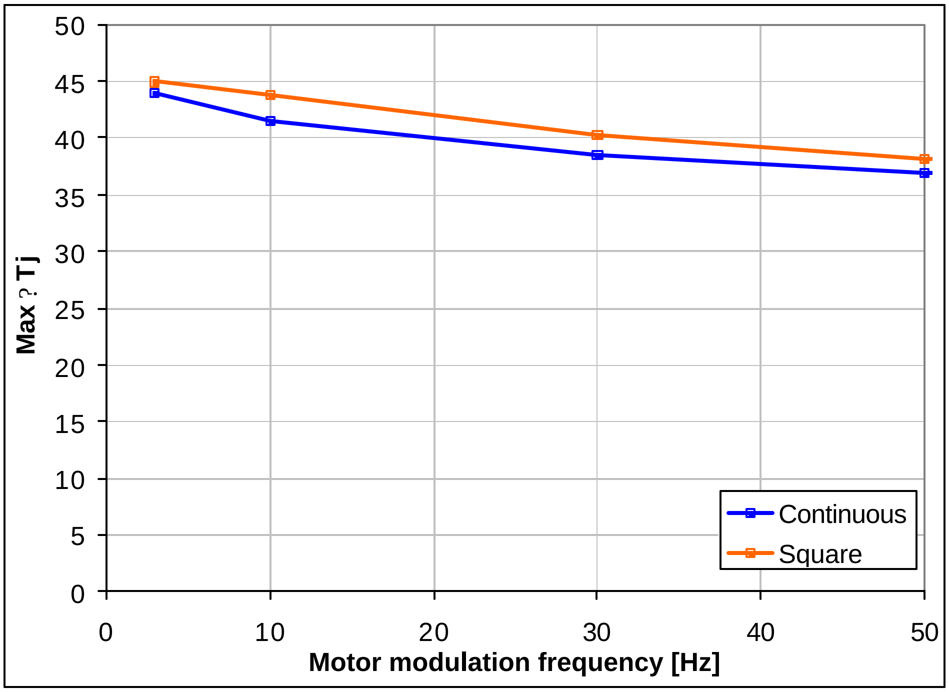

Figure 5 - ? T_J estimation with 2 methods as a function of motor modulation frequency IRAMX16UP60A; Vbus= 400V; pf= 0.6; mi= 0.8; fsw=10kHz; I= 5Arms

As can be seen, the maximum junction temperature calculated with the proposed approximations is 41.5°C with the continuous waveform, and 43.8°C with the square wave input: this is less than 6% higher. This error is acceptable; moreover the approximated temperature is higher so it increases the safety margin of the application. In Figure 5 is shown the difference between these 2 methods as a function of motor modulation frequency. As can be seen the approximation proposed in this paper can be used throughout the modulation frequency range used in motor drive applications.

Current rating curves

Based on the previous calculation methods, it is possible to calculate the maximum current that a specific power module can deliver to the motor in specific application conditions. The curves are generated with the following assumptions:

?? The case temperature is assumed constant (e.g. 100°C)

?? The maximum current is the current that will cause the junction temperature to reach the maximum allowed (e.g. 150°C)

?? Specific modulation technique (sinusoidal or space vector modulation), and specific operational parameters (power factor, modulation index)

First the current rating is given as a function of switching frequency (see Figure 6). This curve shows the influence of the switching losses.

The current capability can also be given as a function of motor modulation frequency (see Figure 7). This curve is generated with a constant switching frequency and shows the effect of the ripple in junction temperature: at lower modulation frequencies, the ripple in junction temperature increases, causing the current rating of the power module to be lower to allow the peak junction temperature not to exceed the preset limit (150°C).

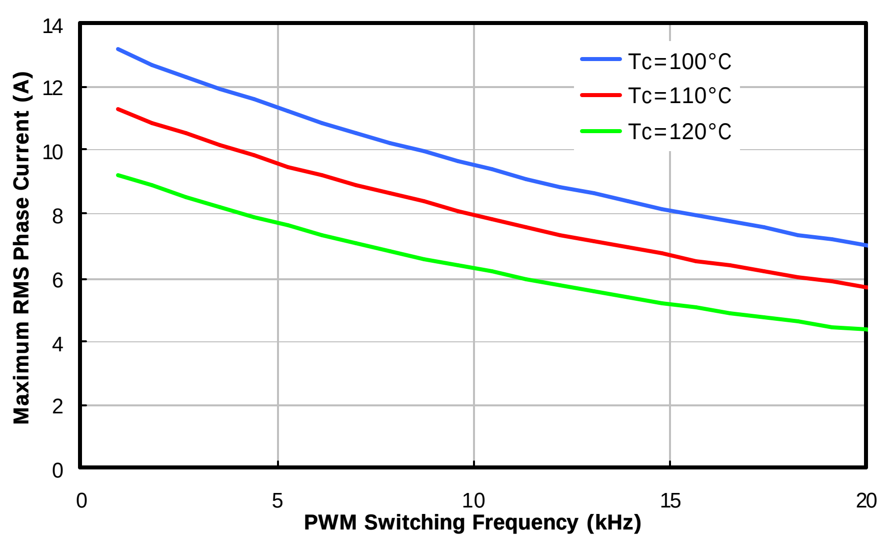

Figure 6 - Power module current rating vs switching frequency IRAMX16UP60A; Tc= 100°C; Vbus= 400V; pf= 0.6; mi= 0.8

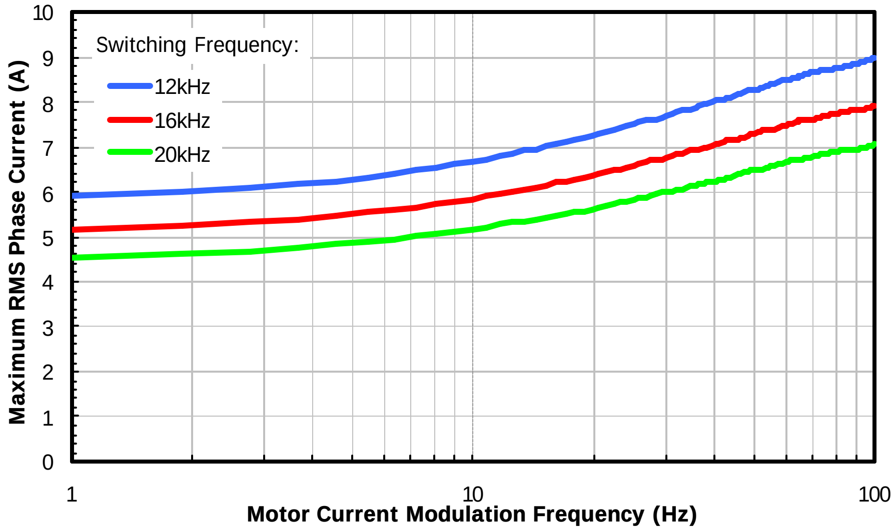

Figure 7 - Power module current vs motor modulation frequency IRAMX16UP60A; Tc= 100°C; Vbus= 400V; pf= 0.6; mi= 0.8

Shunt resistor power dissipation

In a typical motor drive application, shunt resistors are used to provide current feedback and over-current protection. Often a shunt resistor is placed in the low side bus (as shown in Figure 8). This resistor sees a current waveform that depends on the modulation technique. An important part of the motor drive design is the definition of the power dissipation in this shunt resistor. Again, it is quite hard to link with a formula the RMS phase current and the RMS current in the shunt. But it is possible to do this by using a simple model and running a few modulations under the most common operational conditions. The models considered are meant to represent only the different switching strategies and the topology shown in Figure 8. In our case we used Pspice where the IGBT and diodes of the inverter were replaced by almost ideal switches.

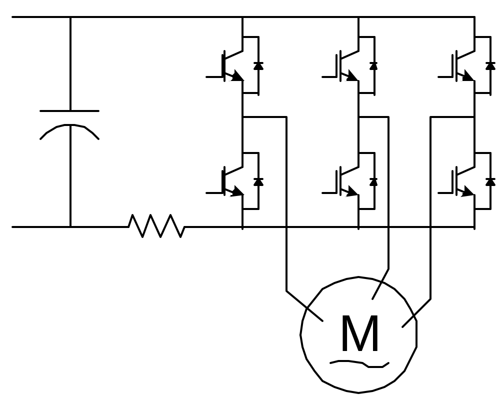

Figure 8 - Location of DC bus shunt

An example of the results of these simulations is shown in Figure 9. As can be seen the waveform is highly non linear and a numerical method is well suited to analyse this problem.

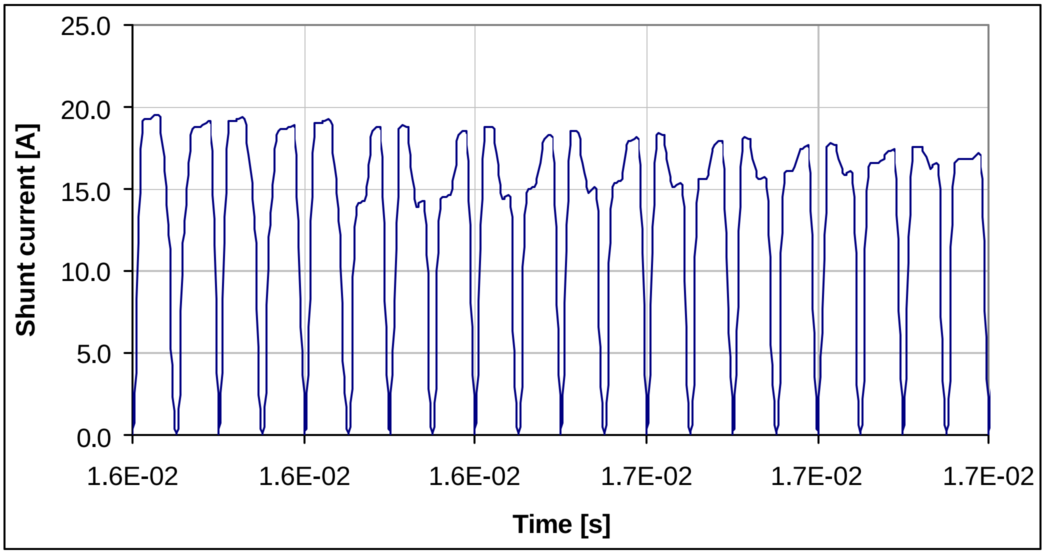

Figure 9 - Example of current in a low side shunt resistor SVM modulation; cos? =0.9; mi=0.98; fmod=50Hz; fsw=10kHz

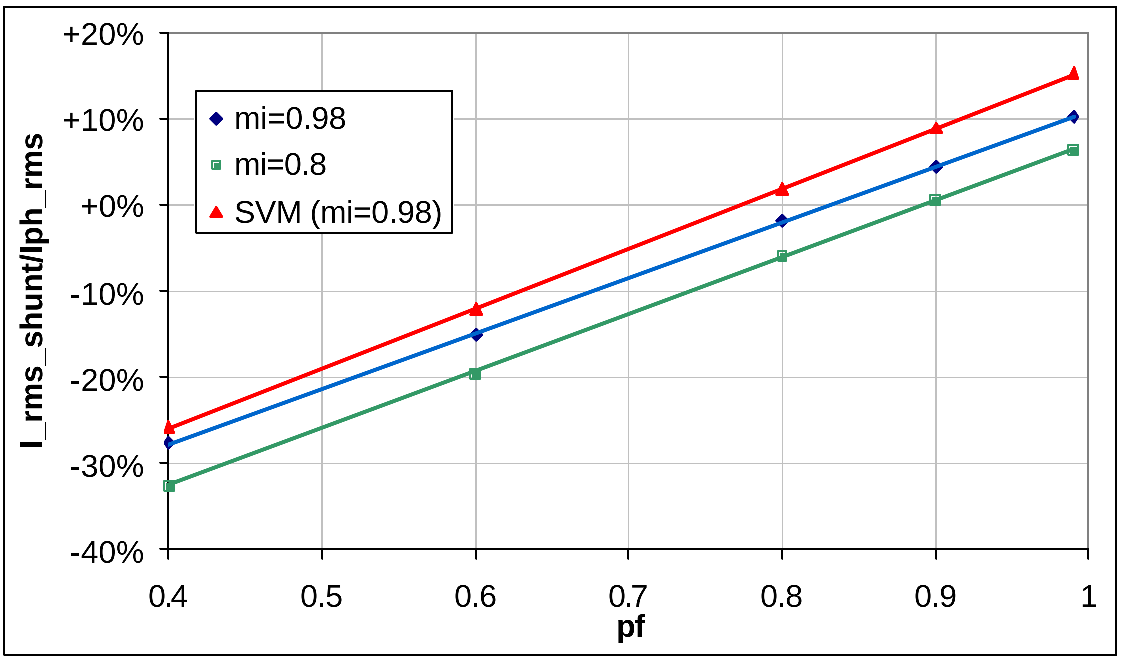

Figure 10 - Shunt RMS current vs Motor RMS current

The results of several simulations with different control strategies and modulation parameters are shown in Figure 10. The worst case is when using space vector modulation, with high power factor and modulation index. In this condition the RMS current in the shunt resistor can be up to 15% higher than the RMS current in the motor. This result can be used to estimate the worst-case power dissipation in the shunt resistor, although in a real application both modulation index and power factor cannot be 1.

Conclusion

In this paper we have shown a method that can greatly simplify the calculations necessary to correctly choose a power stage in motor drive applications. First of all we chose to use only behavioural models to allow simple implementation and fast simulation. Second, a method to estimate the ripple in junction temperature (crest factor) is shown which uses these models.

This method was applied to the case of IPM (integrated power modules), and current ratings are shown as an example. These current ratings are much more useful than typical DC ratings because they show the capability of the power module in real application conditions.

But because of this users will not usually find their exact conditions in the standard datasheet. For this reason IR is developing a tool that uses this method and includes the models for the IRAM PlugnDrive^{TM} family of power modules. A screen shot of the main panel where all the parameters are input is shown in Figure 11. This tool will allow the user to rapidly recalculate the current rating curves shown in this paper based on their specific application conditions.

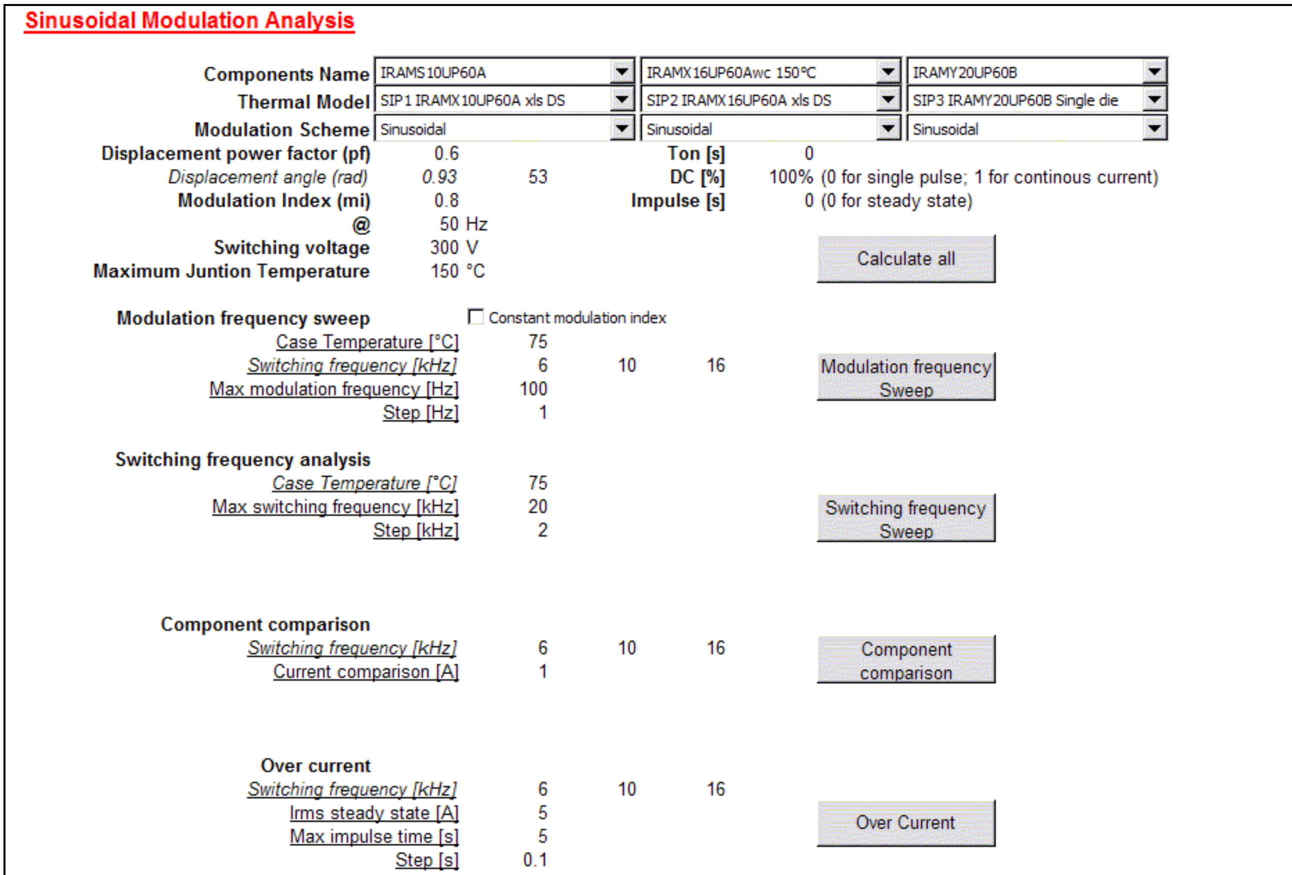

Figure 11 - Main panel of simulation tool