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Evaluating the Power Capability of NCP101X Members

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APPLICATION NOTE

The NCP101X series is available in various combinations of peak current and switching frequencies. To help the designer quickly picking up the right part, it is important to present guidelines aimed to simplify the selection process. This application note details how to evaluate the power handled by each device and offers an overview of each part capability.

Each reference is affected by a key parameter that needs to be accounted for at the very beginning like the switching frequency, the maximum peak current and the Rds(ON). Among these parameters, we also find another set of constraints but linked to the adopted topology, i.e. current mode in our case:

- *DCM Operation:* As the device operates in current-mode, it is mandatory to keep operating in discontinuous mode with a duty-cycle bounded below 50% to avoid sub-harmonic oscillations when accidentally entering Continuous Conduction Mode (CCM). We will use 0.45 as the maximum value in our examples.
- *Nonnegative Reflection:* The built-in lateral MOSFET does not accept to see its body diode forward biased by an excessive Flyback voltage greater than the bulk voltage during the OFF time. Hence, for universal mains operation, the turn ratio Ns:Np will be selected to keep Vreflect = (Ns/Np) x (Vout + Vf) below the very minimum operating DC input voltage of your converter (including the input 120/100 Hz ripple in offline applications). For instance, if we have a 100 VAC input, it becomes 141 V once rectified minus the selected ripple. If choose a ±20% ripple, then the very minimum DC voltage is 141–20% = 112 V. We can select a 100 V maximum reflection voltage (noted Vr) for a safe operation on wide mains, whereas this number will grow-up above 200 V for European mains operation (230 VAC ±15%).
- *Maximum Peak Current:* In the NCP101X series, the peak current is internally fixed. This parameter plays an obvious role in the maximum transmitted power since it obeys the DCM Flyback formula, Pout = 0.5 x η x Lp x Ip^2 x Fsw. One input of our process selection will thus be the reference peak current, 100 mA, 250 mA, 350 mA or 450 mA.

Once the theoretical power capability has been evaluated, it will finally be necessary to check the power dissipation of the switcher itself (given conduction losses, switching losses, etc.) and see what remains available for a reliable operation.

Operating in DCM

When the switch closes, Vin is applied across the primary inductance Lp until the current reaches the level imposed by the feedback loop. The duration of this event is called the ON time and can be defined by:

$$Ton = \frac{Lp \cdot Ip}{Vin} \qquad \text{(eq. 1)}$$

Lp, primary inductance, Vin the input voltage, Ip the operating peak current.

At the switch opening, the primary energy is transferred to the secondary and the flyback voltage appears across Lp, resetting the transformer core with a slope of $\frac{N \cdot (Vout + Vf)}{Lp}$. Toff, the OFF time is thus:

$$Toff = \frac{Lp \cdot Ip}{N \cdot (Vout + Vf)} \qquad \text{(eq. 2)}$$

Lp, primary inductance, Vout the output voltage, Ip the operating peak current, Vf the secondary rectifier voltage drop, N the transformer turn ratio, Ns:Np.

If one wants to keep DCM only, but still need to pass the maximum power, we will not allow a dead-time after the core is reset, but rather immediately restart (fixed frequency boundary mode operation). The switching period can be expressed by:

$$Tsw = Toff + Ton = Lp \cdot Ip \cdot \left(\frac{1}{Vin} + \frac{1}{N \cdot (Vout + Vf)}\right) \quad \text{(eq. 3)}$$

with Vin the input voltage

The Flyback transfer formula dictates that: $\frac{Pout}{\eta} = \frac{1}{2} \cdot Lp \cdot Ip^2 \cdot Fsw$ (eq. 4) which, by extracting Ip (with η the converter efficiency) and plugging into equation 3, leads to:

$$Tsw = Lp \sqrt{\frac{2 \cdot Pout}{\eta \cdot Fsw \cdot Lp}} \cdot \left(\frac{1}{Vin} + \frac{1}{N \cdot (Vout + Vf)}\right) \quad \text{(eq. 5)}$$

Extracting Lp from equation 5 gives:

$$Lp_{critical} = \frac{(Vin \cdot Vr)^2 \cdot \eta}{2 \cdot Fsw \cdot [Pout \cdot (Vr^2 + 2 \cdot Vr \cdot Vin + Vin^2)]}$$

(eq. 6) , with Vr = N. (Vout + Vf) our reflected voltage…

Selecting a primary inductance value lower than the one given by equation 6 ensures discontinuous operation at the lowest line for a given reflected voltage.

Nonnegative Reflection

If we operate on universal mains from 100 VAC to 250 VAC, then the maximum reflected voltage is:

VinAC x 1.414 – ripple = 112 V with a selected ±20% ripple (eq. 7) . We can take 100 V to include a safety margin.

Running the part from a single Europeans mains offers greater flexibility. The voltage is 230 VAC ±15%, which leads to a minimum AC operating voltage of: 230-15% = 195 VAC. The maximum reflected voltage is thus:

VinAC x 1.414 – ripple = 220 V with a selected ±10% ripple (eq. 8) . We can take 210 V to include a safety margin.

Maximum Peak Current

The maximum peak current is given by the particular part reference. If we follow the data sheet, these values are:

	NCP1010			NCP1011			NCP1012			NCP1013			NCP1014	
Rdson (Ω)	23						11							
Ipeak (mA)	100			250			250			350			450	
Freq (kHz)	65	100	130	65	100	130	65	100	130	65	100	130	65	100

Combining all these parameters together, we can finally calculate what theoretical maximum power we can pass satisfying the three bullets expressed at the beginning of this document.

From the maximum peak current, duty-cycle constraint and minimum input voltage (universal or narrow mains), we deduce the maximum inductor Lpmax we can use:

$$Lpmax = \frac{DC_{max} \cdot Vinput_{min} \cdot Tsw}{Ip_{max}} \quad \text{(eq. 9)}$$

. If we apply the following parameters (45% max DC, 120 VDC minimum voltage, 65 kHz switching frequency and 100 mA peak current), it gives an upper inductance boundary of: 0.45 x 120 x 15.4 μ/0.1 = 8.3 mH.

If we now equate equation 9 with equation 6, we can obtain the maximum power constrained by a 45% duty-cycle, a maximum peak current and a given minimum input voltage:

$$Pmax := Tsw^2 \cdot Vinmin^2 \cdot Vr^2 \cdot \eta \cdot \frac{Fsw}{(2 \cdot Lpmax \cdot Vr^2 + 4 \cdot Lpmax \cdot Vr \cdot Vinmin + 2 \cdot Lpmax \cdot Vinmin^2)} \quad \text{(eq. 10)}$$

Keep the same parameters as above, we obtain, Pmax = 2.2 W. Please note that increasing the switching frequency will not expand the power capability of the given converter but will reduce the inductance and allow a smaller magnetic element (the L x I^2 goes down). Running the same chart with all the listed references, gives the first following correspondence between a given peak current and a *theoretical* maximum power obtained from a *converter* operated at high line and low line:

Table 1. Theoretical Transmitted Power Depending on Peak Current *Only*

Peak Current	Wide Mains Operation	230 ± 15% Operation
450 mA	8.9 W	18.6 W
350 mA	6.9 W	14.5 W
250 mA	5.0 W	10.3 W
100 mA	2.0 W	4.1 W

Accounting for the Part Parameters

We now need to refine these calculations knowing our thermal constraints and internal power dissipation (conduction and switching losses, Dynamic Self-Supply, etc.) in order to offer a final selection table. The following chart will be used to assess all possible power combinations given the NCP101X family:

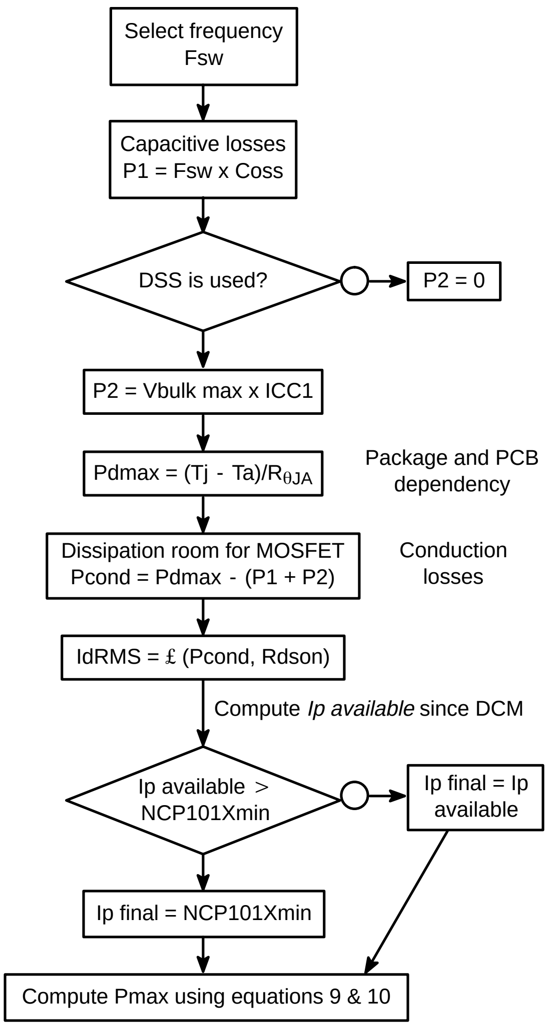

Figure 1. Power Flowchart - Used methodology for assessing the maximum power capability from a particular part number.

Power Dissipation of the PDIP7 Package

The power dissipated by the PDIP7 package is dependent upon its internal consumption and the total thermal resistance junction-to-ambient $R_{\theta JA}$. If we start from a 70°C ambient temperature and a $R_{\theta JA}$ of 75°C/W (which was measured on the demoboard with 35 μ added copper – please see data sheet for suggested layout), then the maximum dissipation from the PDIP7 is: (Max Tj–Ta)/ $R_{\theta JA}$ = (125–70)/75 = 730 mW which grows up to 1.0 W for a lower maximum ambient temperature of Ta = 50°C.

Calculating the total power consumption of a monolithic circuit implies splitting the budget with the various contributors:

1. ***Dynamic Self Supply (DSS):*** The *average* current flowing through the DSS is directly the current needed by the chip to operate (neglecting the switching losses on the DSS itself. Therefore, P_{DSS} = ICC1 x V_{HV}. Therefore, if we in average, the parts exhibit an average ICC1 consumption of 1.0 mA, then the maximum DSS dissipated power is: 1.0 m x 370 V = 0.37 W. (This number drops to 0 W with an auxiliary winding and thus offers better margin for the MOSFET.)
2. ***Switching Losses:*** Theoretically, the turn-on losses are null since we turn on the MOSFET at zero current (DCM). However, there still is a parasitic capacitance on the MOSFET (Coss and Crss) which play a role in the power dissipation budget. To assess the value of this capacitor, we can measure the time taken by the current to come back to zero:

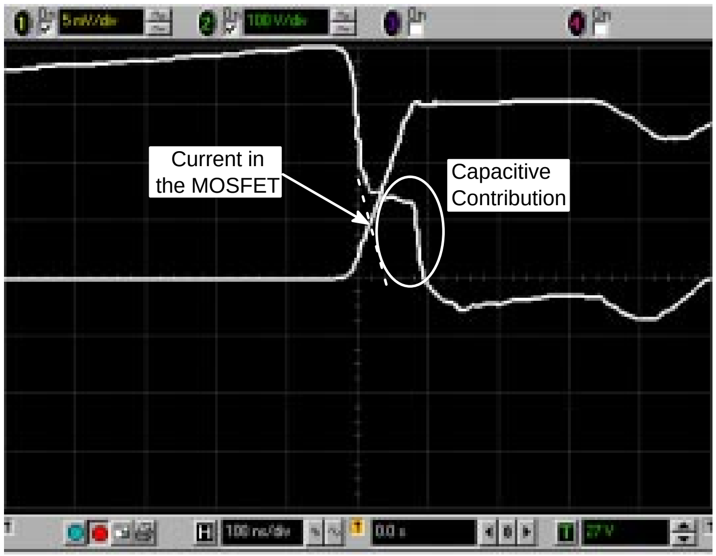

Figure 2. Typical Turn-Off Behavior of an NCP101X Series Member

The presence of the "square" corresponds to a capacitive current flowing inside the MOSFET…but also through the 10 pF scope probe. This current does not create turn-off losses (except losses due to ohmic paths) but it reveals the existence of a capacitor that will create additional losses during turn-on. Based on Figure 2, this capacitor roughly equals: 70 mA x 100 ns/300 V = 25 pF-10 pF = 15 pF. This is pretty low and is typical of lateral MOSFETs. In DCM, this capacitor can be charged to Vin in the worse case. Therefore, the energy stored in the capacitor is: 0.5 x 15 p x 370^2 = 1.03 μJ. Depending on the switching frequency, we will have the following average losses:

1.03 μ x 65000 = 67 mW

1.03 μ x 100000 = 103 mW

1.03 μ x 130000 = 134 mW

3. ***Conduction Losses:*** These losses are dependent on the device $Rds_{(ON)}$ and the RMS current flowing into it. In lack of internal ramp compensation, the converter has to operate in Discontinuous Conduction Mode (DCM) to avoid sub-harmonic oscillations. If we add powers described at bullets 1 and 2, we have the remaining power for the MOSFET alone at different ambient temperatures (Ploss in the previous chart):

Table 2. Available Losses Budget in Function of the Switching Frequency Selection

Fswitching	DSS 50°C	DSS 70°C	A.W. 50°C	A.W. 70°C
65 kHz	490 mW	220 mW	930 mW	670 mW
100 kHz	450 mW	190 mW	900 mW	630 mW
130 kHz	420 mW	150 mW	860 mW	600 mW

As we can see from Table 2, applications needing the maximum power at high ambient temperature will require the use of an auxiliary to improve the room for MOSFET power dissipation. By definition, we know that Pcond = $Idrms^2$ x $Rds_{(ON)}$, hence the maximum allowable ***RMS*** current can be deducted from the previous table. For instance, 490 mW = $Idrms^2$ x $Rds_{(ON)}$ $\rightarrow$ Idrms < $\sqrt{\frac{Pcond}{Rds_{(ON)}}} = \sqrt{\frac{0.49}{25}}$ = 140 mA RMS (eq. 11) .

By definition, the RMS value of a triangular current waveform (SMPS operated in DCM) is: Irms = Ipeak x sqrt (d/3). Since d = 45% maximum, then Ipeak = Irms / 387 m. By computing all RMS and peak values for the different versions, we obtain the following arrays (RMS current/ Peak current):

Rdson of 25 Ω @ Tj = 125°C (NCP1012/13/14)

Table 3. Available RMS/Peak Current for the 11 Ω Version, Respecting Table 2 Figures

Fswitching	DSS 50°C	DSS 70°C	A.W. 50°C	A.W. 70°C
65 kHz	140 mA/361 mA	94 mA/242 mA	193 mA/498 mA	163 mA/423 mA
100 kHz	134 mA/346 mA	87 mA/225 mA	190 mA/490 mA	158 mA/410 mA
130 kHz	129 mA/335 mA	77 mA/200 mA	185 mA/479 mA	155 mA/400 mA

Rdson of 52 Ω @ Tj = 125°C (NCP1010/11)

Table 4. Available RMS/Peak Current for the 23 Ω Version, Respecting Table 2 Figures

Fswitching	DSS 50°C	DSS 70°C	A.W. 50°C	A.W. 70°C
65 kHz	97 mA/250 mA	65 mA/168 mA	130 mA/345 mA	113 mA/293 mA
100 kHz	93 mA/240 mA	60 mA/156 mA	131 mA/340 mA	110 mA/284 mA
130 kHz	90 mA/232 mA	54 mA/139 mA	129 mA/332 mA	107 mA/277 mA

Final Selection Table

From the two above tables 3 and 4, we can now compute the maximum power handled by the component applying equations 9 and 10, where the duty-cycle is constrained to 45%, Vreflec = 100 V for universal mains but rises up to 210 V in single mains, offering more flexibility. If the device offers a peak current capability greater than the value recommended by table 3 or 4, then unfortunately table 3 and 4 set the priority as the flow chart indicated. To the opposite, if the available peak current room exceeds the maximum part peak setpoint, then the part peak current takes the lead. Two short examples can detail this methodology with a 100 kHz, 350 mA device featuring a 12 Ω $Rds_{(ON)}$:

Table 3 states that the maximum peak current at Ta = 70°C equals 225 mA when the DSS is used. The peak current value entered in equation 9 is thus 225 mA which gives a maximum power of 3.41 W (eq. 10). Here, we cannot use the full dynamic of the 350 mA because of the dissipation constraint imposed by table 3.

If we now wire an auxiliary winding, the peak current room given by table 3 rises up to 410 mA. But this time, the peak limit is bounded by the part setpoint of 350 mA – 10% = 315 mA min (350 m < 465 m). This value will therefore be entered into equation 9 and gives a maximum power of 6.8 W.

These calculations were performed for a universal mains application (Vreflect = 100 V) and can also for a narrow mains input with a Vreflect of 210 V. All results are gathered in table 5 and 6, offering a power handling capability device by device.

Universal Mains Applications (100-260 VAC)

Table 5. Power Capability Per Selected Device in Universal Mains Applications

Part Reference	Key Parameters	DSS 50°C	DSS 70°C	A.W. 50°C	A.W. 70°C
NCP1012P06	65 kHz – 23 Ω – 250 mA	4.9	4.9	4.9	4.9
NCP1013P06	65 kHz – 23 Ω – 350 mA	6.8	5.3	6.8	6.8
NCP1014P06	65 kHz – 23 Ω – 450 mA	7.8	5.3	8.8	8.8
NCP1012P10	100 kHz – 23 Ω – 250 mA	4.9	4.8	4.9	4.9
NCP1013P10	100 kHz – 23 Ω – 350 mA	6.8	4.8	6.8	6.8
NCP1014P10	100 kHz – 23 Ω – 450 mA	7.6	4.8	8.8	8.8
NCP1012P13	130 kHz – 23 Ω – 250 mA	4.9	4.4	4.9	4.9
NCP1013P13	130 kHz – 23 Ω – 350 mA	6.8	4.4	6.8	6.8
NCP1010P06	65 kHz – 52 Ω – 100 mA	2.0	2.0	2.0	2.0
NCP1011P06	65 kHz – 52 Ω – 250 mA	4.9	3.7	4.9	4.9
NCP1010P10	100 kHz – 52 Ω – 100 mA	2.0	2.0	2.0	2.0
NCP1011P10	100 kHz – 52 Ω – 250 mA	4.9	3.3	4.9	4.9
NCP1010P13	130 kHz – 52 Ω – 100 mA	1.9	1.9	1.9	1.9
NCP1011P13	130 kHz – 52 Ω – 250 mA	4.9	3.0	4.9	4.9

Narrow Mains Applications (230 VAC ± 15%)

Table 6. Power Capability Per Selected Device in Narrow Mains Applications

Part Reference	Key Parameters	DSS 50°C	DSS 70°C	A.W. 50°C	A.W. 70°C
NCP1012P06	65 kHz – 23 Ω – 250 mA	10.6	10.6	10.6	10.6
NCP1013P06	65 kHz – 23 Ω – 350 mA	14.8	11.7	14.8	14.8
NCP1014P06	65 kHz – 23 Ω – 450 mA	17.3	11.7	19	19
NCP1012P10	100 kHz – 23 Ω – 250 mA	10.6	10.6	10.6	10.6
NCP1013P10	100 kHz – 23 Ω – 350 mA	14.8	10.6	14.8	14.8
NCP1014P10	100 kHz – 23 Ω – 450 mA	16.6	10.7	19	19
NCP1012P13	130 kHz – 23 Ω – 250 mA	10.6	9.6	10.6	10.6
NCP1013P13	130 kHz – 23 Ω – 350 mA	14.8	9.6	14.8	14.8
NCP1010P06	65 kHz – 52 Ω – 100 mA	4.2	4.2	4.2	4.2
NCP1011P06	65 kHz – 52 Ω – 250 mA	10.6	8	10.6	10.6
NCP1010P10	100 kHz – 52 Ω – 100 mA	4.2	4.2	4.2	4.2
NCP1011P10	100 kHz – 52 Ω – 250 mA	10.6	7.4	10.6	10.6
NCP1010P13	130 kHz – 52 Ω – 100 mA	4.2	4.2	4.2	4.2
NCP1011P13	130 kHz – 52 Ω – 250 mA	10.6	6.7	10.6	10.6

Taking the Right Part

As one can see, there is a lot of overlap between the part themselves. We can use this characteristic to fine tune the final design and reach the optimal price/performance ratio. For instance, there are a few questions which, once answered, will naturally push toward the exact reference.

1. *"Do I need an accurate Over Current Protection point?":* If the answer is yes, then go to the DSS option only. If the answer is no, an auxiliary winding will help you passing more power with a cooler part.
2. *"Is the EMI filtering a big constraint on my design?":* A yes means you need the DSS to provide the frequency jittering. If not, auxiliary winding is an option as in point 1. Also, filtering a 65 kHz pattern is easier than a 100 kHz or 130 kHz.
3. *"I need to operate my converter at the highest ambient temp!":* In that case, go for the auxiliary winding.
4. *"My application requires a protection against optocoupler failures…":* By sensing the auxiliary current flowing into the Vcc pin, the part self-protects against open-loop runaways. Go for the auxiliary winding option.
5. *"I have a converter that already runs at 130 kHz with an auxiliary winding!":* In that case, no option, provided that power budget is compatible…
6. *"I need the smallest possible size…":* If the frequency increase does not help to pass more power, it certainly provides a size reduction in the magnetic element. Go for a 130 kHz version.

Notes

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