# SOLUTIONS

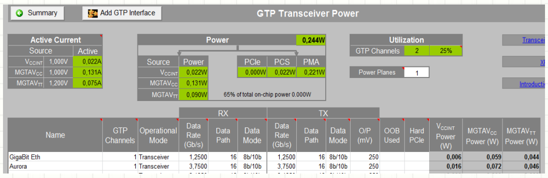


# ARTIX 7



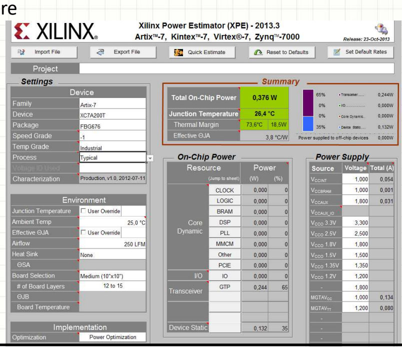
# Exercice de conception alimentation Artix-7

- Alimentation GTP Artix-7

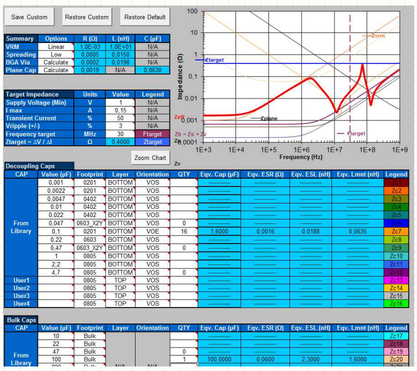


# Exercice de conception alimentation Artix-7

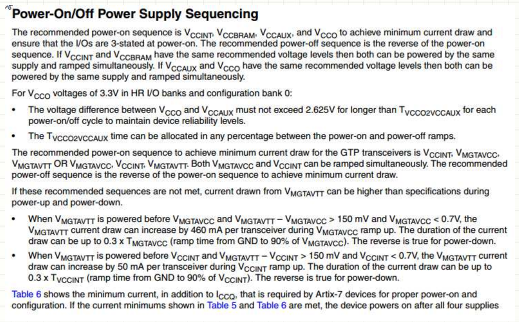
- Sommaire



# Exercice de conception alimentation Artix-7



# Exercice de conception alimentation Artix-7



**Power-On/Off Power Supply Sequencing**

The recommended power-on sequence is $V_{CCINT}$, $V_{CCBRAM}$, $V_{CCAUX}$, and $V_{CCO}$ to achieve minimum current draw and ensure that the I/Os are 3-stated at power-on. The recommended power-off sequence is the reverse of the power-on sequence. If $V_{CCINT}$ and $V_{CCBRAM}$ have the same recommended voltage levels then both can be powered by the same supply and ramped simultaneously. If $V_{CCAUX}$ and $V_{CCO}$ have the same recommended voltage levels then both can be powered by the same supply and ramped simultaneously.

For $V_{CCO}$ voltages of 3.3V in HR I/O banks and configuration bank 0:

- The voltage difference between $V_{CCO}$ and $V_{CCAUX}$ must not exceed 2.625V for longer than $T_{VCCO2VCCAUX}$ for each power-on/off cycle to maintain device reliability levels.
- The $T_{VCCO2VCCAUX}$ time can be allocated in any percentage between the power-on and power-off ramps.

The recommended power-on sequence to achieve minimum current draw for the GTP transceivers is $V_{CCINT}$, $V_{MGTAVCC}$, $V_{MGTAVTT}$ OR $V_{MGTAVCC}$, $V_{CCINT}$, $V_{MGTAVTT}$. Both $V_{MGTAVCC}$ and $V_{CCINT}$ can be ramped simultaneously. The recommended power-off sequence is the reverse of the power-on sequence to achieve minimum current draw.

If these recommended sequences are not met, current drawn from $V_{MGTAVTT}$ can be higher than specifications during power-up and power-down.

- When $V_{MGTAVTT}$ is powered before $V_{MGTAVCC}$ and $V_{MGTAVTT} - V_{MGTAVCC} > 150$ mV and $V_{MGTAVCC} < 0.7V$, the $V_{MGTAVTT}$ current draw can increase by 460 mA per transceiver during $V_{MGTAVCC}$ ramp up. The duration of the current draw can be up to 0.3 x $T_{MGTAVCC}$ (ramp time from GND to 90% of $V_{MGTAVCC}$). The reverse is true for power-down.
- When $V_{MGTAVTT}$ is powered before $V_{CCINT}$ and $V_{MGTAVTT} - V_{CCINT} > 150$ mV and $V_{CCINT} < 0.7V$, the $V_{MGTAVTT}$ current draw can increase by 50 mA per transceiver during $V_{CCINT}$ ramp up. The duration of the current draw can be up to 0.3 x $T_{VCCINT}$ (ramp time from GND to 90% of $V_{CCINT}$). The reverse is true for power-down.

Table 6 shows the minimum current, in addition to $I_{CCQ}$, that is required by Artix-7 devices for proper power-on and configuration. If the current minimums shown in Table 5 and Table 6 are met, the device powers on after all four supplies

# Exercice de conception alimentation Artix-7

*Table 2:* **Recommended Operating Conditions**[1][2]

| Symbol | Description | Min | Typ | Max | Units |
|---|---|---|---|---|---|
| **FPGA Logic** | | | | | |
| $V_{CCINT}$[3] | Internal supply voltage | 0.95 | 1.00 | 1.05 | V |
| | For -2L (0.9V) devices: internal supply voltage | 0.87 | 0.90 | 0.93 | V |
| $V_{CCAUX}$ | Auxiliary supply voltage | 1.71 | 1.80 | 1.89 | V |
| $V_{CCBRAM}$[3] | Block RAM supply voltage | 0.95 | 1.00 | 1.05 | V |
| $V_{CCO}$[4][5] | Supply voltage for 3.3V HR I/O banks | 1.14 | – | 3.465 | V |
| $V_{IN}$[6] | I/O input voltage | –0.20 | – | $V_{CCO}$ + 0.20 | V |
| | I/O input voltage (when $V_{CCO}$ = 3.3V) for $V_{REF}$ and differential I/O standards except TMDS_33[7] | –0.20 | – | 2.625 | V |
| $I_{IN}$[8] | Maximum current through any pin in a powered or unpowered bank when forward biasing the clamp diode. | – | – | 10 | mA |
| $V_{CCBATT}$[9] | Battery voltage | 1.0 | – | 1.89 | V |
| **GTP Transceiver** | | | | | |
| $V_{MGTAVCC}$[10] | Analog supply voltage for the GTP transmitter and receiver circuits | 0.97 | 1.0 | 1.03 | V |
| $V_{MGTAVTT}$[10] | Analog supply voltage for the GTP transmitter and receiver termination circuits | 1.17 | 1.2 | 1.23 | V |
| **XADC** | | | | | |
| $V_{CCADC}$ | XADC supply relative to GNDADC | 1.71 | 1.80 | 1.89 | V |
| $V_{REFP}$ | Externally supplied reference voltage | 1.20 | 1.25 | 1.30 | V |



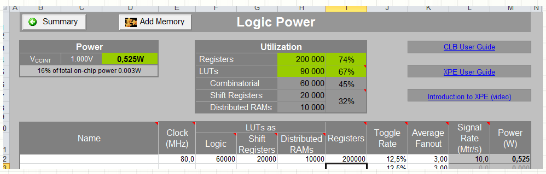
# Exercice de conception alimentation Artix-7

- Il y faut séquencer les alimentations dans leur ordre de « grandeur » : 1 v; 1,2 V; 1,8 V; 2,5 V; 3,3V.
- Donc VccInt et VmgtAVcc ensemble (attention Avcc sensible) puis VmgtAVtt puis VccAux et VccO 1,8 V, puis VccO 2,5 V enfin VccO 3,3 V.



# Exercice de conception alimentation Artix-7

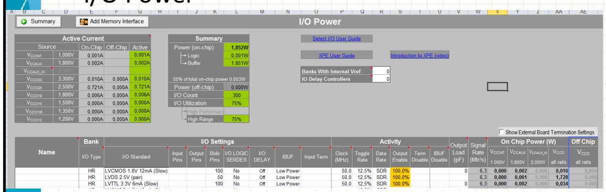
- Logic power






# Exercice de conception alimentation Artix-7

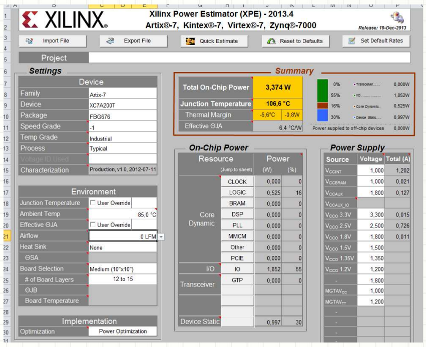
- I/O Power



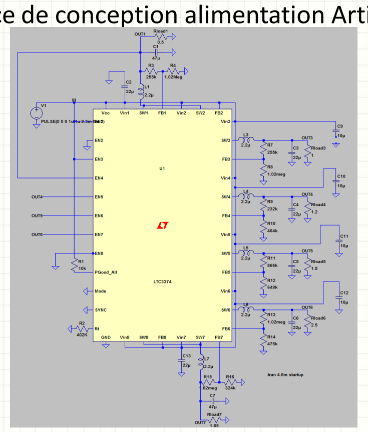



# Exercice de conception alimentation Artix-7

- Sommaire




# Exercice de conception alimentation Artix-7



# Exercice de conception alimentation Artix-7 (LTC3374)

**Buck Switching Regulator Output Voltage and Feedback Network**

The output voltage of the buck switching regulators is programmed by a resistor divider connected from the switching regulator's output to its feedback pin and is given by $V_{OUT} = V_{FB}(1 + R2/R1)$ as shown in Figure 2. Typical values for R1 range from 40k to 1M. The buck regulator transient response may improve with optional capacitor $C_{FF}$ that helps cancel the pole created by the feedback resistors and the input capacitance of the FB pin. Experimentation with capacitor values between 2pF and 22pF may improve transient response.

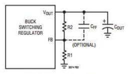



Figure 2. Feedback Components

**Table 1. Recommended Inductors for 1A Buck Regulators**

| PART NUMBER | L (µH) | MAX $I_{DC}$ (A) | MAX DCR (mΩ) | SIZE IN mm (L × W × H) | MANUFACTURER |
|---|---|---|---|---|---|
| IHLP1212ABER1R0M-11 | 1.0 | 3 | 38 | 3 × 3.6 × 1.2 | Vishay |
| 1239AS-H-1R0N | 1 | 2.5 | 65 | 2.5 × 2.0 × 1.2 | Toko |
| XFL4020-222ME | 2.2 | 3.5 | 23.5 | 4 × 4 × 2.1 | CoilCraft |
| 1277AS-H-2R2N | 2.2 | 2.6 | 84 | 3.2 × 2.5 × 1.2 | Toko |
| IHLP1212BZER2R2M-11 | 2.2 | 3 | 46 | 3 × 3.6 × 1.2 | Vishay |
| XFL4020-332ME | 3.3 | 2.8 | 38.3 | 4 × 4 × 2.1 | CoilCraft |
| IHLP1212BZER3R3M-11 | 3.3 | 2.7 | 61 | 3 × 3.6 × 1.2 | Vishay |

**Table 2. Recommended Inductors for 2A Buck Regulators**

| PART NUMBER | L (µH) | MAX $I_{DC}$ (A) | MAX DCR (mΩ) | SIZE IN mm (L × W × H) | MANUFACTURER |
|---|---|---|---|---|---|
| XFL4020-102ME | 1.0 | 5.1 | 11.9 | 4 × 4 × 2.1 | CoilCraft |
| 74437324010 | 1 | 9 | 27 | 4.45 × 4.06 × 1.8 | Wurth Elektronik |
| XAL4020-222ME | 2.2 | 5.6 | 38.7 | 4 × 4 × 2.1 | CoilCraft |
| FDV0530-2R2M | 2.2 | 5.3 | 15.5 | 6.2 × 5.8 × 3 | Toko |
| IHLP2020BZER2R2M-11 | 2.2 | 5 | 37.7 | 5.49 × 5.18 × 2 | Vishay |
| XAL4030-332ME | 3.3 | 5.5 | 28.6 | 4 × 4 × 3.1 | CoilCraft |
| FDV0530-3R3M | 3.3 | 4.1 | 34.1 | 6.2 × 5.8 × 3 | Toko |

**Buck Regulators**

All eight buck regulators are designed to be used with inductors ranging from 1µH to 3.3µH depending on the lowest switching frequency that the buck regulator must operate at. To operate at 1MHz a 3.3µH inductor should be used, while to operate at 3MHz a 1µH inductor may be used. Table 1 shows some recommended inductors for the buck regulators.

The input supply needs to be decoupled with a 10µF capacitor while the output needs to be decoupled with a 22µF capacitor. Refer to the Capacitor Selection section for details on selecting a proper capacitor.

**Combined Buck Regulators**

A single 2A buck regulator is available by combining two adjacent 1A buck regulators together. Likewise a 3A or 4A buck regulator is available by combining any three or four adjacent buck regulators respectively. Tables 2, 3, and 4 show recommended inductors for these configurations.

The input supply needs to be decoupled with a 22µF capacitor while the output needs to be decoupled with a 47µF capacitor for a 2A combined buck regulator. Likewise for 3A and 4A configurations the input and output capacitance must be scaled up to account for the increased load. Refer to the Capacitor Selection section for details on selecting a proper capacitor.

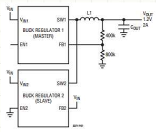
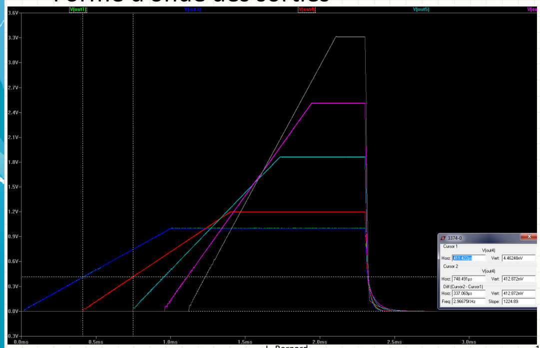



Figure 1. Buck Regulators Configured as Master-Slave

# Exercice de conception alimentation Artix-7

- Forme d'onde des sorties





# Exercice de conception alimentation Artix-7

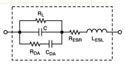
LTC3374

**APPLICATIONS INFORMATION**

**Table 1. Recommended Inductors for 1A Buck Regulators**

| PART NUMBER | L (µH) | MAX $I_{DC}$ (A) | MAX DCR (mΩ) | SIZE IN mm (L × W × H) | MANUFACTURER |
|---|---|---|---|---|---|
| IHLP1212ABER1R0M-11 | 1.0 | 3 | 38 | 3 × 3.6 × 1.2 | Vishay |
| 1239AS-H-1R0N | 1 | 2.5 | 65 | 2.5 × 2.0 × 1.2 | Toko |
| XFL4020-222ME | 2.2 | 3.5 | 23.5 | 4 × 4 × 2.1 | CoilCraft |
| 1277AS-H-2R2N | 2.2 | 2.6 | 84 | 3.2 × 2.5 × 1.2 | Toko |
| IHLP1212BZER2R2M-11 | 2.2 | 3 | 46 | 3 × 3.6 × 1.2 | Vishay |
| XFL4020-332ME | 3.3 | 2.8 | 38.3 | 4 × 4 × 2.1 | CoilCraft |
| IHLP1212BZER3R3M-11 | 3.3 | 2.7 | 61 | 3 × 3.6 × 1.2 | Vishay |

- Modèle Capacité




b) Continuous Conduction Mode:

$$\Delta Vout = ESR.(\Delta IL/2 + 4Iload)$$

Accepting the larger values of inductor current ripple allows the use of lower inductance, but results in higher output voltage ripple and greater core and power device conduction losses.

$$L = \frac{1}{fs \cdot \Delta IL_{max}} \cdot \frac{Vout_{max}}{4}$$

*Eqn. 12*

**MC13783 Buck and Boost Inductor Sizing Application Note, Rev. 0.1**

Freescale Semiconductor 5

# LT3020



# Composant LT3020 (R2) Alimentation 5 V vers 1 V 100 mA

- $R2 = \frac{V_{Out} - 200\ mV}{\frac{200\ mV}{R1} - I_{ADJ}}$ avec $R1 = 20k\ et\ I_{ADJ} = 20nA$
- R2 = 80,16 k exactement
- R2 = 82 k (E24)

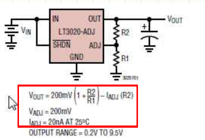



Figure 1. Adjustable Operation

3020fc

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# Composant LT3020 (C1) Alimentation 5 V vers 1 V 100 mA

- $C1 = 2{,}2\ \mu F$ (E3)

**Input Capacitance and Stability**

The LT3020 is designed to be stable with a minimum capacitance of 2.2µF placed at the IN pin. Ceramic capacitors with very low ESR may be used. However, in cases where a long wire is used to connect a power supply to the input of the LT3020 (and also from the ground of the LT3020 back to the power supply ground), use of low value input capacitors combined with an output load current of 20mA or greater may result in an unstable application. This is due to the inductance of the wire forming an LC tank circuit with the input capacitor and not a result of the LT3020 being unstable.

The self-inductance, or isolated inductance, of a wire is directly proportional to its length. However, the diameter

If the LT3020 is powered by a battery mounted in close proximity on the same circuit board, a 2.2µF input capacitor is sufficient for stability. However, if the LT3020 is powered by a distant supply, use a larger value input capacitor following the guideline of roughly 1µF (in addition to the 2.2µF minimum) per 8 inches of wire length. As power supply output impedance may vary, the minimum input capacitance needed to stabilize the application may also vary. Extra capacitance may also be placed directly on the output of the power supply; however, this will require an order of magnitude more capacitance as opposed to placing extra capacitance in close proximity to the LT3020. Furthermore, series resistance may be placed between the supply and the input of the LT3020 to stabilize the application; as little as 0.1Ω to 0.5Ω will suffice.

3020fc

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## Composant LT3020 (C2)
## Alimentation 5 V vers 1 V 100 mA

- $C2 = 2{,}2\ \mu F\ (E3)\ avec\ un\ ESR < 0{,}3\ ohm$

**Output Capacitance and Transient Response**

The LT3020's design is stable with a wide range of output capacitors, but is optimized for low ESR ceramic capacitors. The output capacitor's ESR affects stability, most notably with small value capacitors. Use a minimum output capacitor of 2.2µF with an ESR of 0.3Ω or less to prevent oscillations. The LT3020 is a low voltage device, and output load transient response is a function of output capacitance. Larger values of output capacitance decrease the peak deviations and provide improved transient response for larger load current changes. For output capacitor values greater than 20µF a small feedforward capacitor with a value of 300pF across the upper divider resistor (R2 in Figure 1) is required.

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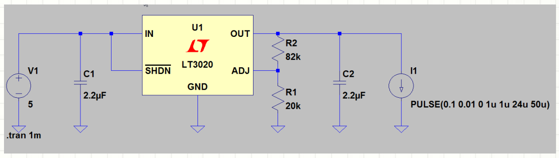
## Composant LT3020 (simulation)
## Alimentation 5 V vers 1 V 100 mA

- R2 = 82 k
- C1 = 2,2 µF
- C2 = 2,2 µF
- Simulation avec LTSpice IV

lt3020 avec valeurs.asc




L. Bernard 131

## Composant LT3020 (radiateur)
## Alimentation 5 V vers 1 V 100 mA

- $P = I_{OUTmax}(V_{INmax} - V_{OUT}) + I_{GND} \times V_{INmax}$
- $P = 100mA(5V - 1V) + 1500\ \mu A \times 5V = 0{,}41\ W$
- $T_{rise} = P \times Therm_{resistance}$
- $S = 100\ mm2 \Rightarrow Therm_{resistance} = 60°\ {}^{C}/_{W}$
- $T_{rise} = 0{,}41W \times 60°\ {}^{C}/_{W} = 24{,}5\ °C$

GND Pin Current

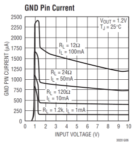



Table 1. Measured Thermal Resistance for DD Package

| COPPER AREA TOPSIDE* | COPPER AREA BACKSIDE | BOARD AREA | THERMAL RESISTANCE (JUNCTION-TO-AMBIENT) |
|---|---|---|---|
| 2500mm² | 2500mm² | 2500mm² | 35°C/W |
| 900mm² | 2500mm² | 2500mm² | 40°C/W |
| 225mm² | 2500mm² | 2500mm² | 55°C/W |
| 100mm² | 2500mm² | 2500mm² | 60°C/W |
| 50mm² | 2500mm² | 2500mm² | 70°C/W |

**Calculating Junction Temperature**

Example: Given an output voltage of 1.8V, an input voltage range of 2.25V to 2.75V, an output current range of 1mA to 100mA, and a maximum ambient temperature of 70°C, what will the maximum junction temperature be for an application using the DD package?

The power dissipated by the device is equal to:

$I_{OUT(MAX)}(V_{IN(MAX)} - V_{OUT}) + I_{GND}(V_{IN(MAX)})$

where

$I_{OUT(MAX)} = 100mA$

$V_{IN(MAX)} = 2.75V$

$I_{GND}$ at ($I_{OUT} = 100mA$, $V_{IN} = 2.75V$) = 3mA

so

P = 100mA(2.75V – 1.8V) + 3mA(2.75V) = 0.103W

The thermal resistance is in the range of 35°C/W to 70°C/W depending on the copper area. So the junction temperature rise above ambient is approximately equal to:

0.103W(52.5°C/W) = 5.4°C

The maximum junction temperature equals the maximum junction temperature rise above ambient plus the maximum ambient temperature or:

$T_{JMAX}$ = 70°C + 5.4°C = 75.4°C

**Thermal Considerations**

The LT3020's power handling capability is limited by its maximum rated junction temperature of 125°C. The power dissipated by the device is comprised of two components:

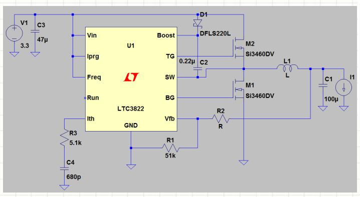
L. Bernard 2

# LT3822

L. Bernard 133

## Composant LT3822 (D et SF)
## Alimentation 3,3 V vers 1 V 3 A

- $D = \frac{V_{out}}{V_{in}}$
- $D = \frac{1}{3{,}3} = 30{,}3\ \%\ donc\ SF = 97\ \%\ (figure\ 1)$




**Design Example**

For a design example, $V_{IN}$ will be a 3.3V power supply. Output voltage is 1.2V with a load current requirement of 10A. The IPRG and FREQ pins will be left floating, so the maximum current sense threshold $\Delta V_{SENSE(MAX)}$ will be approximately 120mV and the switching frequency will be 550kHz.

$$\text{Duty Cycle} = \frac{V_{OUT}}{V_{IN}} = 36.4\%$$

From Figure 1, SF = 96%.

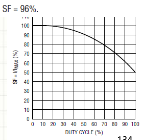
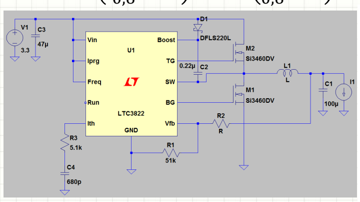



Figure 1. Maximum Peak Current vs Duty Cycle

L. Bernard 134

## Composant LT3822 (R2)
## Alimentation 3,3 V vers 1 V 3 A

- $V_{out} = 0{,}6(1 + \frac{R_2}{R_1})$
- $R_2 = \left(\frac{V_{out}}{0{,}6} - 1\right) R_1 = \left(\frac{1}{0{,}6} - 1\right) 51k = 34k\ donc\ 33k$




**Setting Output Voltage**

The LTC3822 output voltage is set by an external feedback resistor divider carefully placed across the output, as shown in Figure 3. The regulated output voltage is determined by:

$$V_{OUT} = 0.6V \cdot \left(1 + \frac{R_B}{R_A}\right)$$

For most applications, a 59k resistor is suggested for $R_A$. In applications where minimizing the quiescent current is critical, $R_A$ should be made bigger to limit the feedback divider current. If $R_B$ then results in very high impedance, it may be beneficial to bypass $R_B$ with a 50pF to 100pF capacitor $C_{FF}$.

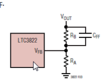



Figure 3. Setting the Output Voltage

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## Composant LT3822
## Alimentation 3,3 V vers 1 V 3 A

- $R_{DSmax} = \frac{5}{6} 0{,}9\ SF\ \left(\frac{\Delta V_{SENSEmax}}{I_{OUTmax}\ \rho_T}\right)\ avec$
- $\Delta V_{SENSEmax} = 200\ mV\ et\ \rho_T = 1{,}3$ à $70\ °\ C$
- $R_{DSmax} = \frac{5}{6}\ 0{,}9\ 0{,}97\ \left(\frac{200m}{3\ x\ 1{,}3}\right) = 0{,}037 ohm$

The topside MOSFET's on-resistance is chosen based on the required load current. The maximum average load current $I_{OUT(MAX)}$ is equal to the peak inductor current minus half the peak-to-peak ripple current $I_{RIPPLE}$. The LTC3822's current comparator monitors the drain-to-source voltage $V_{DS}$ of the top MOSFET, which is sensed between the $V_{IN}$ and SW pins. The peak inductor current is limited by the current threshold, set by the voltage on the $I_{TH}$ pin, of the current comparator. The voltage on the $I_{TH}$ pin is internally clamped, which limits the maximum current sense threshold $\Delta V_{SENSE(MAX)}$ to approximately 120mV when IPRG is floating (82mV when IPRG is tied low; 200mV when IPRG is tied high).

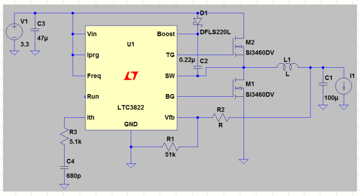
The output current that the LTC3822 can provide is given by:

$$I_{OUT(MAX)} = \frac{\Delta V_{SENSE(MAX)}}{R_{DS(ON)}} - \frac{I_{RIPPLE}}{2}$$

These must be further derated to take into account the significant variation in on-resistance with temperature. The following equation is a good guide for determining the required $R_{DS(ON)MAX}$ at 25°C (manufacturer's specification), allowing some margin for variations in the LTC3822 and external component values:

$$R_{DS(ON)MAX} = \frac{5}{6} \bullet 0.9 \bullet SF \bullet \frac{\Delta V_{SENSE(MAX)}}{I_{OUT(MAX)} \bullet \rho_T}$$

The $\rho_T$ is a normalizing term accounting for the temperature variation in on-resistance, which is typically about 0.4%/°C, as shown in Figure 2. Junction-to-case temperature $\Delta T_{JC}$ is about 10°C in most applications. For a maximum ambient temperature of 70°C, using $\rho_{80°C} \sim 1.3$ in the above equation is a reasonable choice.




L. Bernard 136

## Composant LT3822 (L)
## Alimentation 3,3 V vers 1 V 3 A

- $L_{min} >= \left(\frac{V_{in} - V_{out}}{fosc\ I_{ripple}}\right) \frac{V_{out}}{V_{in}}$
- $I_{ripple} = 0{,}4\ I_{OUTmax} = 0{,}4\ \times 3 = 1{,}2\ A$
- $L_{min} = \left(\frac{3{,}3\ -1}{750\ k\ 1{,}2}\right) \frac{1}{3{,}3} = 0{,}77$ µH donc 1µH

**Inductor Value Calculation**

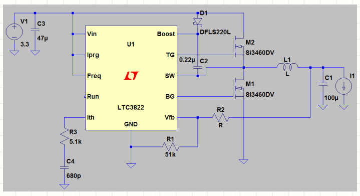
Given the desired input and output voltages, the inductor value and operating frequency, $f_{OSC}$, directly determine the inductor's peak-to-peak ripple current:

$$I_{RIPPLE} = \frac{V_{OUT}}{V_{IN}} \bullet \frac{V_{IN} - V_{OUT}}{f_{OSC} \bullet L}$$

Lower ripple current reduces core losses in the inductor, ESR losses in the output capacitors and output voltage ripple. Thus, highest efficiency operation is obtained at low frequency with a small ripple current. Achieving this, however, requires a large inductor.

A reasonable starting point is to choose a ripple current that is about 40% of $I_{OUT(MAX)}$. Note that the largest ripple current occurs at the highest input voltage. To guarantee that ripple current does not exceed a specified maximum, the inductor should be chosen according to:

$$L \geq \frac{V_{IN} - V_{OUT}}{f_{OSC} \bullet I_{RIPPLE}} \bullet \frac{V_{OUT}}{V_{IN}}$$




L. Bernard 137

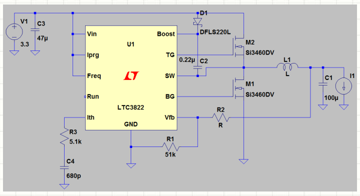
## Composant LT3822 ($I_{RMS}\ de\ C_{IN}$)
## Alimentation 3,3 V vers 1 V 3 A

- $I_{RMS} = I_{OUTmax}\left(\frac{V_{OUT}\sqrt{V_{IN} - V_{OUT}}}{V_{IN}}\right)$
- $I_{RMS} = 3\left(\frac{1\sqrt{3{,}3 - 1}}{3{,}3}\right) = 1{,}4\ A$

**$C_{IN}$ and $C_{OUT}$ Selection**

In continuous mode, the source current of the top MOSFET is a square wave of duty cycle ($V_{OUT}/V_{IN}$). To prevent large voltage transients, a low ESR input capacitor sized for the maximum RMS current must be used. The maximum RMS capacitor current is given by:

$$C_{IN}\ Required\ I_{RMS} \approx I_{MAX} \bullet \frac{V_{OUT} \bullet (V_{IN} - V_{OUT})^{1/2}}{V_{IN}}$$




L. Bernard 138

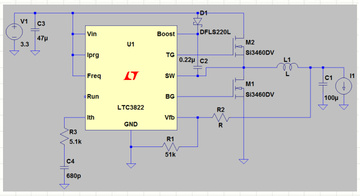
## Composant LT3822 (ESR $de\ C_{OUT}$)
## Alimentation 3,3 V vers 1 V 3 A

- $ESR = \frac{\Delta V_{OUT}}{I_{ripple}} - \frac{1}{8\ f\ C_{OUT}}$
- $ESR = \frac{50\ mV}{1{,}2} - \frac{1}{8 \times 750kHz\ 100µF} = 0{,}04\ ohm$

The selection of $C_{OUT}$ is driven by the effective series resistance (ESR). Typically, once the ESR requirement is satisfied, the capacitance is adequate for filtering. The output ripple ($\Delta V_{OUT}$) is approximated by:

$$\Delta V_{OUT} \approx I_{RIPPLE} \bullet \left(ESR + \frac{1}{8 \bullet f \bullet C_{OUT}}\right)$$

where f is the operating frequency, $C_{OUT}$ is the output capacitance and $I_{RIPPLE}$ is the ripple current in the inductor. The output ripple is highest at maximum input voltage since $I_{RIPPLE}$ increase with input voltage.

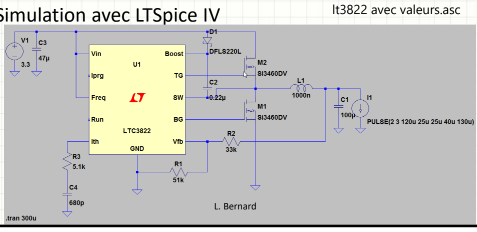



L. Bernard 139

## Composant LT3822
## Alimentation 3,3 V vers 1 V 3 A

- R2 = 33 k
- NMOS Si3460DV $R_{DSmax} = 0{,}03\ ohm$
- L1 = 1 µH
- C1 supporte 11 A RMS
- C2 a une ESR de 0,002 ohm.
- Simulation avec LTSpice IV

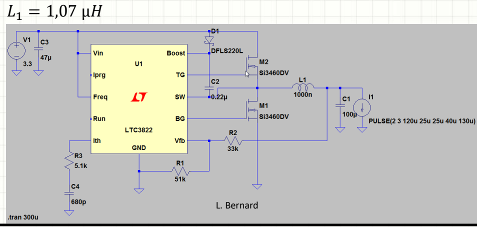
lt3822 avec valeurs.asc




L. Bernard 140

## Composant LT3822 (L formule)
## Alimentation 3,3 V vers 1 V 3 A

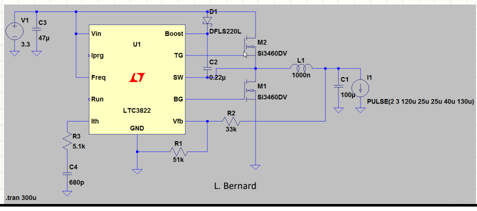
- $L = \frac{V_{InMax} - V_{Out}}{f_{osc} \times \Delta I_L} \times \frac{V_{Out}}{V_{InMax}}$
- $V_{InMax} = 3{,}6\ V$
- $V_{Out}$ = 1 V
- $f_{osc}$ = 750 kHz
- $\Delta I_L = 0{,}9\ A$
- $L_1 = 1{,}07$ µH




L. Bernard 141

# Composant LT3822 (C formule)
## Alimentation 3,3 V vers 1 V 3 A

- $C = \frac{L \times (I_{OutMax} + \Delta I_L)^2}{(\Delta V + V_{Out})^2 - V_{Out}^2}$
- $I_{OutMax} = 3\ A$
- $V_{Out} = 1\ \mathrm{V}$
- $\Delta I_L = 0{,}9\ A$
- $L_1 = 1\ \mu H$
- $\Delta V = 50\ mV$
- $C_1 = 148\ \mu F$




L. Bernard 142

# LT3872

L. Bernard 143

# Composant LT3872 (R2)
## Alimentation 3,3 V vers 12 V 1 A

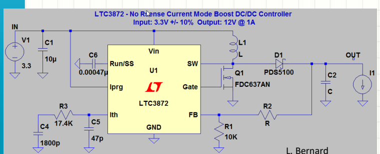
- $V_{out} = 1{,}2(1 + \frac{R_2}{R_1})$
- $R_2 = \left(\frac{V_{out}}{1{,}2} - 1\right) R_1 = \left(\frac{1}{1{,}2} - 1\right) 10k = 90k\ donc\ 91k$

**Output Voltage Programming**

The output voltage is set by a resistor divider according to the following formula:

$$\boxed{V_O = 1.2V \bullet \left(1 + \frac{R2}{R1}\right)}$$

The external resistor divider is connected to the output as shown in the Typical Application on the front page, allowing remote voltage sensing.




L. Bernard 144

# Composant LT3872 (D)
## Alimentation 3,3 V vers 12 V 1 A

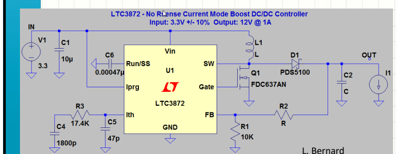
- $D = \frac{V_{out} + V_D - V_{in}}{V_{out} + V_D}$
- $D = \frac{12 + 0{,}4 - 3{,}3}{12 + 0{,}4} = 0{,}734$
- $D_{max} = \frac{12 + 0{,}4 - 3{,}3 \times 0{,}9}{12 + 0{,}4} = 0{,}760$

**Duty Cycle Considerations**

For a boost converter operating in a continuous conduction mode (CCM), the duty cycle of the main switch is:

$$\boxed{D = \left(\frac{V_O + V_D - V_{IN}}{V_O + V_D}\right)}$$

where $V_D$ is the forward voltage of the boost diode. For converters where the input voltage is close to the output voltage, the duty cycle is low and for converters that develop a high output voltage from a low voltage input supply, the duty cycle is high. The LTC3872 has a built-in circuit that allows the extension of the maximum duty cycle while keeping the minimum switch off time unchanged. This is accomplished by reducing the clock frequency when the duty cycle is close to 80%. This function allows the user to obtain high output voltages from low input supply voltages. The shift of frequency with duty cycle is shown in the Typical Performance Characteristics section.




L. Bernard 145

# Composant LT3872 ($I_{INpeak}\ et\ \Delta I_L$)
## Alimentation 3,3 V vers 12 V 1 A

- $I_{INpeak} = \left(1 + \frac{x}{2}\right) \frac{I_{OUTmax}}{1 - D_{max}}\ avec\ x = 40\%$
- $I_{INpeak} = \left(1 + \frac{0{,}4}{2}\right) \frac{1}{1 - 0{,}734} = 4{,}52\ A$
- $\Delta I_L = x \frac{I_{OUTmax}}{1 - D_{max}} = 0{,}4 \frac{1}{1 - 0{,}734} = 1{,}50\ A$


- $I_{INpeak} = \left(1 + \frac{x}{2}\right) \frac{I_{OUTmax}}{1 - D_{max}}\ avec\ x = 40\%$
- $I_{INpeak} = \left(1 + \frac{0{,}4}{2}\right) \frac{1}{1 - 0{,}734} = 4{,}52\ A$
- $\Delta I_L = x \frac{I_{OUTmax}}{1 - D_{max}} = 0{,}4 \frac{1}{1 - 0{,}734} = 1{,}50\ A$


**The Peak and Average Input Currents**

The control circuit in the LTC3872 is measuring the input current (either by using the $R_{DS(ON)}$ of the power MOSFET or by using a sense resistor in the MOSFET source), so the output current needs to be reflected back to the input in order to dimension the power MOSFET properly. Based on the fact that, ideally, the output power is equal to the input power, the maximum average input current is:

$$I_{IN(MAX)} = \frac{I_{O(MAX)}}{1 - D_{MAX}}$$

The peak input current is:

$$\boxed{I_{IN(PEAK)} = \left(1 + \frac{\chi}{2}\right) \bullet \frac{I_{O(MAX)}}{1 - D_{MAX}}}$$

**Inductor Selection**

Given an operating input voltage range, and having chosen the operating frequency and ripple current in the inductor, the inductor value can be determined using the following equation:

$$L = \frac{V_{IN(MIN)}}{\Delta I_L \bullet f} \bullet D_{MAX}$$

where:

$$\boxed{\Delta I_L = \chi \bullet \frac{I_{O(MAX)}}{1 - D_{MAX}}}$$

L. Bernard 146

# Composant LT3872 (L)
## Alimentation 3,3 V vers 12 V 1 A

- $L_{min} >= \frac{V_{INmin}}{\Delta I_L \times f} D_{max}$
- $L_{min} >= \frac{3{,}3 \times 0{,}9}{1{,}50 \times 550k} 0{,}760 = 3{,}01 \mu H\ donc\ 3{,}1 \mu H$

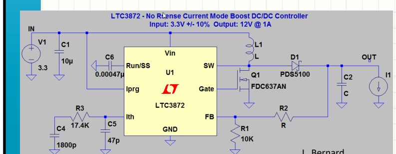
**Inductor Selection**

Given an operating input voltage range, and having chosen the operating frequency and ripple current in the inductor, the inductor value can be determined using the following equation:

$$\boxed{L = \frac{V_{IN(MIN)}}{\Delta I_L \bullet f} \bullet D_{MAX}}$$

where:

$$\Delta I_L = \chi \bullet \frac{I_{O(MAX)}}{1 - D_{MAX}}$$



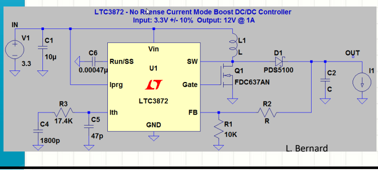

L. Bernard 147

## Composant LT3872 ($R_{DSmax}$) Alimentation 3,3 V vers 12 V 1 A

- $R_{DSmax} = V_{SENSEmax}\left(\frac{1-D_{max}}{(1+\frac{X}{2})I_{OUTmax} \times \rho_T}\right)$ *avec*
- $V_{SENSEmax} = 225\ mV\ (figure\ 3)\ et\ \rho_T$ = 0,4 %/ ° C
- $R_{DSmax} = 225m\left(\frac{1-0{,}760}{(1+\frac{0{,}4}{2})1 \times 0{,}4}\right)$ = 0,113ohm

The relationship between the maximum load current, duty cycle and the $R_{DS(ON)}$ of the power MOSFET is:

$$R_{DS(ON)} \le V_{SENSE(MAX)} \bullet \frac{1-D_{MAX}}{\left(1+\frac{\chi}{2}\right) \bullet I_{O(MAX)} \bullet \rho_T}$$



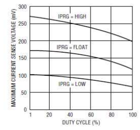




Figure 3. Maximum SENSE Threshold Voltage vs Duty Cycle



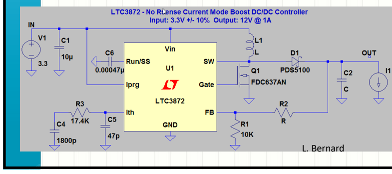
## Composant LT3872 ($I_{Dpeak}$) Alimentation 3,3 V vers 12 V 1 A

- $I_{Dpeak} = (1+\frac{X}{2}) \times \frac{I_{OUTmax}}{1-D_{max}}$
- $I_{Dpeak} = \left(1+\frac{0{,}4}{2}\right) \times \frac{1}{1-0{,}760} = 5\ A$

**Output Diode Selection**

To maximize efficiency, a fast switching diode with low forward drop and low reverse leakage is desired. The output diode in a boost converter conducts current during the switch off-time. The peak reverse voltage that the diode must withstand is equal to the regulator output voltage. The average forward current in normal operation is equal to the output current, and the peak current is equal to the peak inductor current.

$$I_{D(PEAK)} = I_{L(PEAK)} = \left(1+\frac{\chi}{2}\right) \bullet \frac{I_{O(MAX)}}{1-D_{MAX}}$$






## Composant LT3872 ($C_{OUT}$ et ESR) Alimentation 3,3 V vers 12 V 1 A

- $C_{out} \ge \frac{I_{OUTmax}}{0{,}01 \times V_{OUT} \times f} = \frac{1}{0{,}01 \times 12 \times 550k}$

  $= 15{,}2\mu F\ donc\ 22\ \mu F$
- $ESR \le \frac{0{,}01 \times V_{OUT}}{I_{INpeak}} = \frac{0{,}01 \times 12}{4{,}52} = 0{,}027\ ohm$

For a 1% contribution to the total ripple voltage, the ESR of the output capacitor can be determined using the following equation:

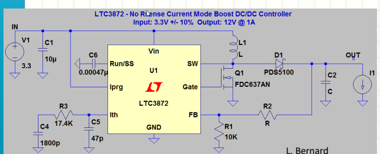
$$ESR_{COUT} \le \frac{0.01 \bullet V_O}{I_{IN(PEAK)}}$$

where:

$$I_{IN(PEAK)} = \left(1+\frac{\chi}{2}\right) \bullet \frac{I_{O(MAX)}}{1-D_{MAX}}$$

For the bulk C component, which also contributes 1% to the total ripple:

$$C_{OUT} \ge \frac{I_{O(MAX)}}{0.01 \bullet V_O \bullet f}$$





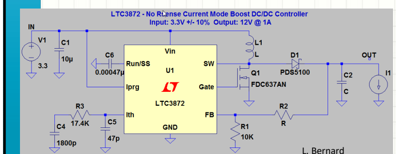

## Composant LT3872 ($I_{RMS}\ de\ C_{IN}$) Alimentation 3,3 V vers 12 V 1 A

- $I_{RMS} = 0{,}3 \times \frac{V_{INmin}}{L \times f} D_{max}$
- $I_{RMS} = 0{,}3 \times \frac{3{,}3 \times 0{,}9}{3{,}1\mu \times 550k} 0{,}760 = 0{,}40A$

The RMS input capacitor ripple current for a boost converter is:

$$I_{RMS(CIN)} = 0.3 \bullet \frac{V_{IN(MIN)}}{L \bullet f} \bullet D_{MAX}$$

Please note that the input capacitor can see a very high surge current when a battery is suddenly connected to the input of the converter and solid tantalum capacitors can fail catastrophically under these conditions. **Be sure to specify surge-tested capacitors!**






## Composant LT3872 Alimentation 3,3 V vers 12 V 1 A

- R2 = 91 k
- NMOS FDC637AN $R_{DSmax} = 0{,}024\ ohm$
- L1 = 3,1 µH
- D1 supporte 5 A.
- C1 supporte 10 A RMS.
- C2 = 22 µF et a une ESR de 0,005 ohm.
- Simulation avec LTSpice IV

ltc3872 12V avec valeurs.asc