Quasi-Resonant type AC/DC converter IC

BM1Q0XX series Quasi-Resonant converter Technical Design

This application note describes the design of Quasi-Resonant converters using ROHM's AC/DC converter IC BM1Q0xx series devices. It explains the selection of external components and provides PCB layout guidelines. Please note that all performance characteristics have to be verified. They are not guaranteed by the PCB layout shown here.

● Description
The BM1Q0xx series of ICs are AC/DC converters for Quasi-Resonant switching, incorporating a built-in starter circuit having withstanding voltage of 650V. Use of external switching MOSFET and current detection resistors, provides a higher degree of design freedom. Power efficiency is improved by the built-in starter circuit and the reduction of switching frequency under light load conditions.

● Key features
Quasi-resonant method
Built-in 650V tolerate start circuit
Low power when load is light (Burst operation)
Maximum frequency control (120kHz)
Frequency reduction function
AC voltage correction function
VCC pin : under voltage protection
VCC pin : overvoltage protection
Over-current protection (cycle-by-cycle)
OUT pin : H voltage 12V clamp
Soft start
ZT trigger mask function
ZT Over voltage protection
FB Over Load protection [Auto-restart]
CS pin open protection [Auto-restart]

● Basic specifications
Operating power supply voltage range(VCC) : 8.9V to 26.0V
VH voltage range(VH pin) : up to 600V
Operating current    : Normal mode 0.60 mA (Typ.)
                   : Burst mode 0.35 mA (Typ.)
Maximum frequency   : 120 KHz (Typ.)
Operating temperature range : -40°C to +85°C

● BM1Q0xx Series line-up

<table>
<thead>
<tr>
<th>Product</th>
<th>Package</th>
<th>VCC OVP</th>
<th>ZT OVP</th>
</tr>
</thead>
<tbody>
<tr>
<td>BM1Q001FJ</td>
<td>SOP-J8</td>
<td>Auto restart</td>
<td>None</td>
</tr>
<tr>
<td>BM1Q002FJ</td>
<td></td>
<td>Latch stop</td>
<td>Latch stop</td>
</tr>
</tbody>
</table>

● Applications
AC adapters, TVs, household appliances (vacuum cleaners, humidifiers, air filters, air conditioners, refrigerators, induction heating cookers, rice cookers, etc.)
1. Design Example of Isolated Type Quasi-Resonant Converter

Quasi-Resonant Converters become DCM (Discontinuous Conduction Mode) under light load, and switching frequency increases with the load increasing. When the load increased further, Quasi-Resonant Converters become BCM (Boundary Conduction Mode), and switching frequency decreases with the load increasing.

Figure 1-1. Isolated Type Quasi-Resonant Converter Circuit Example

Figure 1-2. Switching Frequency – Output Load Characteristics

Figure 1-3. Switching waveform (MOSFET Vds,Ids)
1-1. Transformer T1 design

1-1-1. Determination of flyback voltage VOR
Flyback voltage VOR is determined along with turns-ratio Np:Ns and duty-ratio.

\[
VOR = \frac{VO}{Ns} \times \frac{ton}{toff} \times VIN
\]

\[
\Rightarrow Np = \frac{VOR}{VO} \times Ns
\]

\[
\Rightarrow \text{Duty} = \frac{VOR}{VIN + VOR}
\]

When \( VIN = 95V (AC85V \times 1.4 \times 0.8) \), \( VOR = 78V \), \( Vf = 1V \):

\[
\frac{Np}{Ns} = \frac{VOR}{VO} = \frac{78V}{20V + 1V} = 3.714
\]

\[
\text{Duty(max)} = \frac{VOR}{VIN(\text{min}) + VOR} = \frac{78V}{95V + 78V} = 0.45
\]

(*) When duty is 0.5 or above, VOR is adjusted to set it below 0.5 in consideration of MOSFET’s loss, etc.

1-1-2. Determination of Minimum frequency \( fsw \) and calculation of primary -side winding inductance \( Lp \) and primary-side maximum current \( Ippk \)
When \( VIN = 95V \), set minimum frequency to 38kHz.

Other’s parameter is following:
Since of \( Po = 20V \times 3A = 60W \), \( Po(\text{max}) = 70W \) in consideration of over current protection.
Transformer efficiency: \( \eta = 90\% \)
Resonance capacitor: \( Cv = 100pF \)

\[
Lp = \left[ \frac{\left( \frac{VIN(\text{min}) \times \text{Duty(max)}}{2 \times Po(\text{max}) \times fsw} + \frac{VIN(\text{min}) \times \text{Duty(max)} \times fsw \times \pi \times \sqrt{Cv}}{\eta} \right)}{1} \right] ^{-\frac{1}{2}} = 297uH
\]

\[
Ippk = \frac{2 \times Po(\text{max})}{\eta \times Lp \times fsw} = 3.713A
\]

1-1-3. Determination of transformer size
Based on \( Po(\text{max}) = 70W \), the transformer’s core size is EER35.

<table>
<thead>
<tr>
<th>Output power ( Po(W) )</th>
<th>Core size</th>
<th>Core sectional area Ae (mm²)</th>
</tr>
</thead>
<tbody>
<tr>
<td>~30</td>
<td>EI25/EE25</td>
<td>41</td>
</tr>
<tr>
<td>~60</td>
<td>EI28/EE28/EER28</td>
<td>84</td>
</tr>
<tr>
<td>~80</td>
<td>EI33/EER35</td>
<td>107</td>
</tr>
</tbody>
</table>

(*) The above are guideline values. For details, check with the transformer manufacturer, etc.
1-1-4. Calculation of primary-side turn count Np

Generally, the maximum magnetic flux density B(T) for an ordinary ferrite core is 0.4T @100°C, so Bsaturated = 0.35T.

\[
Np > \frac{Lp \times Ipke}{Ae \times Bsat} = \frac{297\mu H \times 3.713A}{107\text{mm}^2 \times 0.35T} = 29.4 \text{ turns} \rightarrow Np \text{ is 30 turns or above}
\]

Since magnetic saturation does not result from this, Np is set based on the AL-value—NI characteristics.

When Np=30 turns,

AL – Value = \[
\frac{Lp}{Np^2} = \frac{297\mu H}{30 \text{turns}^2} = 330 \text{nH/turns}^2
\]

NI = Np \times Ipke = 30 \text{turns \times 3.713A} = 111.4A \times \text{turns}

In this case, transformer is saturated based on the AL-value—NI characteristics.

When Np=40 turns,

AL – Value = \[
\frac{Lp}{Np^2} = \frac{297\mu H}{40 \text{turns}^2} = 186 \text{nH/turns}^2
\]

NI = Np \times Ipke = 40 \text{turns \times 3.713A} = 148.5A \times \text{turns}

In this case, this point is within the tolerance range.

1-1-5. Calculation of secondary-side turn count Ns

\[
\frac{Np}{Ns} = 3.714 \rightarrow Ns = \frac{40}{3.714} = 10.8 \text{ turns} \rightarrow 11 \text{ turns}
\]

1-1-6. Calculation of VCC turn count Nd

When VCC=15V, Vf_{vcc}=1V,

\[
Nd = Ns \times \frac{VCC + Vf_{vcc}}{Vout + Vf} = 11 \text{turns} \times \frac{15V + 1V}{20V + 1V} = 8.8 \text{turns} \rightarrow 9 \text{ turns}
\]

As a result, the transformer specifications are as follows.

<table>
<thead>
<tr>
<th>Table 1-2. Transformer Specifications</th>
</tr>
</thead>
<tbody>
<tr>
<td>Core</td>
</tr>
<tr>
<td>Lp</td>
</tr>
<tr>
<td>Np</td>
</tr>
<tr>
<td>Ns</td>
</tr>
<tr>
<td>Nd</td>
</tr>
</tbody>
</table>
1-2. Selection of main components

1-2-1. MOSFET: Q1

Factors to select a MOSFET include the maximum drain-source voltage, peak current, loss due to Ron, and the package's maximum allowable loss.

In particular, in the case of world-wide input (AC 85V to AC264 V, etc.), the MOSFET's ON-period becomes long when the input voltage is low, and Ron-loss makes more heat generated. Confirm while the MOSFET is assembled in the product, and, when necessary, use a heat-sink or similar to dissipate the heat.

In this design example, ROHM's MOSFET R8008ANX (800V, 8A, 0.79Ω) is selected based on worldwide input and \( I_{ppk} = 3.713A \).

1-2-2. Input capacitor: C3

Use Table 1-3 to select the capacitance of the input capacitor.

Since \( P_{out} = 20V \times 3A = 60W \), \( C_{1} = 2 \times 60 = 120 \rightarrow 150\mu F \).

Table 1-3. Input Capacitor Selection Table

<table>
<thead>
<tr>
<th>Input voltage (Vac)</th>
<th>Cin(μF)</th>
</tr>
</thead>
<tbody>
<tr>
<td>85-264</td>
<td>2 x Pout(W)</td>
</tr>
<tr>
<td>180-264</td>
<td>1 x Pout(W)</td>
</tr>
</tbody>
</table>

(*) The above values are guidelines for full-wave rectification.

When selecting, also consider other specifications such as the retention-time.

The withstanding voltage of the capacitor becomes, \( V_{ac} \text{(max)} \times 1.41 \). Say for AC 264V, it is 264V \times 1.41 = 372V, so this should be 400V or more.

1-2-3. Setting resistor for changing of over current protection point: R13

When input voltage is high, ON time is short, and switching frequency increases. As a result, maximum output power increases for constant over current limiter. For that, monitoring input voltage, IC switches CS over current voltage level when ZT input current: \( I_{zt} = 1mA \).

Set input voltage to AC150V → DC212V when over current protection point changes.

\[
R_{13} = \frac{V_{IN(change)}}{Nd} \times \frac{1}{I_{zt}} = 212V \times \frac{9 \text{ turns}}{40 \text{ turns}} \times \frac{1}{1mA} = 47.7k\Omega \rightarrow 47k\Omega
\]

1-2-4. Setting resistor for ZT terminal voltage: R14

ZT bottom detected voltage is \( V_{zt1} = 100mV \text{(typ.)} \) (ZT fall), \( V_{zt2} = 200mV \text{(typ.)} \) (ZT rise), and ZT OVP(min) is 4.65V (BM1Q002FJ), so as a guide, set Vzt to 1V to 3V.

\[
V_{zt} = (V_{out} + V_{f}) \times \frac{Nd}{Ns} \times \frac{R_{14}}{R_{13} + R_{14}} = 1.5V \quad R_{14} = 4.495k\Omega \rightarrow 4.3k\Omega
\]
1-2-5. Current-sensing resistor: R10

The current-sensing resistor limits the current that flows on the primary side to provide protection against output overload.

\[ R_{10} = \frac{V_{cs}}{I_{ppk}} = \frac{0.5V}{3.713A} = 0.135\Omega \rightarrow 0.12\Omega \]

Check over current protection point after it was changed.

When IC switches CS over current voltage level, it is changed from 0.5V to 0.35V.

\[ V_{IN}(\text{change}) = R_{13} \times \frac{N_p}{N_d} \times I_{zt} = 47k\Omega \times \frac{40\text{turns}}{9\text{turns}} \times 1mA = 209V \]

\[ I_{ppk}' = \frac{V_{cs}}{R_{10}} = 0.35V \times 0.12\Omega = 2.917A \]

\[ t_{on}' = \frac{L_p \times I_{ppk}'}{V_{IN}(\text{change})} = \frac{297uH \times 2.917A}{209V} = 4.145us \]

\[ I_{spk}' = \frac{N_p}{N_s} \times I_{ppk}' = 40\text{turns} \times 11\text{turns} = 2.971A \]

\[ L_s = \frac{L_p \times N_s^2}{N_p} = 297uH \times \frac{11\text{turns}^2}{40\text{turns}} = 22.46uH \]

\[ t_{off}' = \frac{L_s \times I_{spk}'}{V_{out} + V_f} = \frac{22.46uH \times 10.61A}{20V + 1V} = 11.35us \]

\[ t_{delay} = \pi \times \sqrt{L_p \times C_v} = 3.14 \times \sqrt{297uH \times 100pF} = 0.541us \]

\[ f_{sw}' = \frac{1}{t_{on}' + t_{off}' + t_{delay}} = \frac{1}{4.145us + 11.35us + 0.541us} = \frac{1}{16.03us} = 62.36kHz \]

Transformer efficiency: \( \eta = 90\% \)

\[ P_o' = \frac{1}{2} \times L_p \times I_{ppk}'^2 \times f_{sw}' \times \eta = \frac{1}{2} \times 297uH \times 2.917A^2 \times 62.36kHz \times 0.9 = 70.92W \]

When \( P_o' \) is under the rated output power, \( V_{IN}(\text{change}) \), R10, etc. are adjusted to set \( P_o' \) above rated output power.

Confirm the overload protection point while the resistor is assembled in the product.

Sensing resistor loss \( P_{R10} \):

\[ P_{R10}(\text{peak}) = I_{ppk}'^2 \times R_{10} = 3.713A^2 \times 0.12\Omega = 1.654W \]

\[ P_{R10}(\text{rms}) = I_{rms}^2 \times R_{10} = \left( \frac{I_{ppk} \times \sqrt{\text{Duty}(\text{max})}}{3} \right)^2 \times R_{10} = \left( \frac{3.713A \times \sqrt{0.45}}{3} \right)^2 \times 0.12 = 0.248W \]

Set to 1W or above in consideration of pulse resistance.

With regard to pulse resistance, the structure of the resistance may vary even with the same power rating.

Check with the resistor manufacturers for details.

1-2-6. VCC-diode: D5

A high-speed diode is recommended as the VCC-diode.

When \( D_5_{\text{Vf}} = 1V \), reverse voltage applied to the VCC-diode:

\[ V_{dr} = VCC(\text{max}) + V_{f} + V_{IN}\text{max} \times \frac{N_d}{N_p} \]

When VCC (max) = 29V,

\[ V_{dr} = 29V + 1V + \frac{9\text{turns}}{40\text{turns}} \times 372V = 113.7V \]

With a design-margin taken into account, \( 122.5V / 0.7 = 175V \rightarrow 200V \) component is selected.

(Example: ROHM's RF05VA2S 200V, 0.5A)
1-2-7. VCC capacitor: C6
A VCC capacitor is needed to stabilize the IC’s VCC voltage. Capacitance of 2.2μF or above is recommended (example: 50V, 10μF).
Next, determine the startup time of the IC at power-on. Figure 1-7 illustrates VCC capacitor capacitance and startup time characteristics.

1-2-8. VCC winding surge-voltage limiting resistor: R12
Based on the transformer’s leakage inductance (Lleak), a large surge-voltage (spike noise) may occur during the instant when the MOSFET is switched from ON to OFF. This surge-voltage is induced in the VCC winding, and as the VCC voltage increases the IC’s VCC overvoltage protection may be triggered.
A limiting resistor R2 (approximately 5Ω to 22Ω) is inserted to reduce the surge-voltage that is induced in the VCC winding. Confirm the rise in VCC voltage while the resistor is assembled in the product.

Based on the transformer’s leakage inductance (Lleak), a large surge-voltage (spike noise) may occur during the instant when the MOSFET is switched from ON to OFF. This surge-voltage is applied between the MOSFET’s Drain and Source, so in the worst case damage to MOSFET might occur. RCD snubber circuits are recommended to suppress this surge-voltage.

(1) Determination of clamp voltage (Vclamp) and clamp ripple-voltage (Vripple)
Take a design-margin based on the MOSFET’s withstand voltage for the clamp voltage.
Vclamp = 800V × 0.8 = 640V
The clamp ripple-voltage (Vripple) is about 50V.

(2) Determination of R6

\[ R6 < 2 \times V_{\text{clamp}} \times \frac{V_{\text{clamp}} - V_{\text{OR}}}{L_{\text{leak}} \times f_{\text{sw}} \times \left( f_{\text{sw}}(\text{max}) \right)} \]

Calculation of Ip, fsw when Lleak = Lp × 10% = 297μH × 10% = 29.7μH, Po=60W and Vin(max)=372V.

\[ P_0 = \frac{1}{2} \times Lp \times I_p^2 \times f_{\text{sw}} \times \eta \]
\[ I_p = \frac{V_{\text{cs}}}{R_{\text{cs}}} \times f_{\text{sw}} = \frac{1}{t_{\text{on}} + t_{\text{off}} + t_{\text{delay}}} = \left( \frac{Lp}{V_{\text{IN}}} \times I_p \right)^2 + \left( \frac{Ls}{V_{\text{DS}}} \times \frac{Np}{N_s} \times I_p \right)^2 + \frac{\pi}{\sqrt{Lp \times C_v}} \]

⇒ Vcs=0.2657V, Ip=2.214A, fsw=91.6kHz

R6 is derived as:

\[ R6 < 2 \times 640V \times \frac{640V - 76.4V}{29.7μH \times 2.214^2 \times 91.6kHz} = 54k \Omega \rightarrow 47k \Omega \]

R6 loss \( P_{\text{R6}} \) is expressed as

\[ P_{\text{R6}} = \frac{(V_{\text{clamp}} - V_{\text{IN}})^2}{R6} = \frac{(640 - 372)^2}{47k \Omega} = 1.53W \]

A 3W component is determined with consideration for design margin.
(3) Determination of C4

\[
C4 \geq \frac{V_{\text{clamp}}}{V_{\text{ripple}} \times f_{\text{sw(min)}}} \times \frac{640V}{50V} \times \frac{91.6kHz}{47k} \times 47k = 2973pF \rightarrow 3300pF
\]

The voltage applied to C4 is 640V – 264 × 1.41 = 268V.
Set 400V or above with design margin.

(4) Determination of D3

Choose a fast recovery diode as the diode, with a withstanding voltage that is at or above the MOSFET’s Vds (max) value.

The surge-voltage affects not only the transformer’s leakage inductance but also the PCB substrate’s pattern.
Confirm the Vds voltage while assembled in the product, and adjust the snubber circuit as necessary.

1-2-10. Output rectification diode: D6

Choose a high-speed diode (Schottky barrier diode or fast recovery diode) as the output rectification diode.

When D6_Vf=1V, reverse voltage applied to output diode is

\[
V_{\text{dr}} = V_{\text{out(max)}} + V_{\text{f}} + V_{\text{in}} \times N_s = \frac{N_s}{N_p}
\]

When Vout(max)=20V+5%=21V:

\[
V_{\text{dr}} = 21V + 1V + 372V \times \frac{11}{40} = 124.3V
\]

A 124.3V/0.7=178V → 200V component is determined with consideration for design margin.
Also, diode loss (approximate value) becomes Pd = Vf × Iout = 1V × 3A = 3W.
(Example: Rohm RFN10T2D: 200V, 10A, TO-220F package)

Use of a voltage margin of 70% or less and current of 50% or less is recommended.
Check temperature rise while assembled in the product. When necessary, reconsider the component and use a heat sink or similar to dissipate the heat.

1-2-11. Output capacitors: C11, C12

Determine the output capacitors based on the output load’s allowable peak-to-peak ripple voltage (ΔVpp) and ripple-current.

When the MOSFET is ON, the output diode is OFF. At that time, current is supplied to the load from the output capacitors.
When the MOSFET is OFF, the output diode is ON. At that time, the output capacitors are charged and a load current is also supplied.

When ΔVpp = 200mV,

\[
Z_{\text{C}} < \frac{\Delta V_{\text{pp}}}{I_{\text{spk}}} \times \frac{N_s}{N_p} \times \frac{1}{I_{\text{ppk}}} = \frac{0.2V}{40 \times 3.713A} = 0.0148 \Omega \text{ at 60kHz (fsw min)}
\]

With an ordinary switching power supply electrolytic-capacitor (low-impedance component), impedance is rated at 100kHz, so it is converted to 100kHz.

\[
Z_{\text{C}} < 0.0148 \Omega \times \frac{60}{100} = 0.009 \Omega \text{ at 100kHz}
\]

Ripple-current Is (rms):

\[
I_{\text{rms}} = I_{\text{spk}} \times \sqrt{\frac{1 - \text{Duty}}{3}} = \frac{40}{11} \times 3.713A \times \sqrt{\frac{1 - 0.45}{3}} = 5.781A
\]

The capacitor’s withstanding voltage should be set to about twice the output voltage.
Vout x 2 = 20V x 2 = 40V → 50V or above
Select an electrolytic capacitor that is suitable for these conditions.
(Example: low impedance type 35V, 1000 μF x 2 parallel for switching power supply )
(*) Use the actual equipment to confirm the actual ripple-voltage and ripple-current.
1-2-12. MOSFET gate circuits: R7, R8, D4

The MOSFET’s gate circuits affect the MOSFET’s loss and generate noise. The switching speed for turn-on is adjusted using R7+R8, and for turn-off is adjusted using R7, via the drawing diode D4.

(Example: R7: 10Ω/ 0.25W, R8: 150Ω, D4: SBD 60V, 1A)

In the case of Quasi-Resonant converters, switching-loss basically does not occur during turn-on, but occurs predominantly during turn-off. To reduce switching-loss when turned off, turn-off speed can be increased by reducing R7 value, but sharp changes in current will occur, which increases the switching-noise. Since there is a trade-off between loss (heat generation) and noise, measure the MOSFET’s temperature rise and noise while it is assembled in the product, and adjust it as necessary.

Also, since a pulse current flows to R7, check the pulse resistance of the resistors being used.

1-2-13. FB terminal capacitor: C7

C7 is a capacitor for stability of FB voltage (approximately 1000pF to 0.01µF).

1-2-14. ZT terminal capacitor: C8

C8 is a capacitor for stability of ZT voltage and for timing adjustment of bottom detection.

Check the waveform of ZT terminal and the timing of bottom detection, and adjust it as necessary.

1-2-15. Output voltage setting resistors: R17, R18, R19

When Shunt regulator IC2: Vref=2.495V,

\[ V_o = \left(1 + \frac{R_{17} + R_{18}}{R_{19}}\right) \times V_{ref} = \left(1 + \frac{82k\Omega + 2.2k\Omega}{12k\Omega}\right) \times 2.495V = 20.00V \]

1-2-16. Parts for adjustment of control circuit: R15, R16, R20, C10

R20 and C10 are parts for phase compensation. Approximately R20:1k to 30kΩ , C10=0.1µF, and adjust them while they are assembled in the product.

R15 limits a control circuit current. Approximately R15:300 to 2kΩ , and adjust it while it assembled in the product.

R16 is a resistor for adjustment of minimum operating current of shunt regulator IC2.

In case of IC2: TL431, minimum operating current is 1mA. And when Optocoupler:PC1_Vf is 1V,

R16 = 1V / 1mA = 1kΩ

1-3. EMI countermeasures

Confirm the following with regard to EMI countermeasures.

(*) Constants are reference values. Need to be adjusted based on noise effects.

- Addition of filter to input block
- Addition of capacitor between primary-side and secondary-side (approximately C20: Y-Cap 2200pF)
- Addition of RC snubber to diode (approximately C30: 500V 1000pF, R30: 10Ω, 1W)
1-4. Output noise countermeasures
As an output noise countermeasure, add an LC filter (approximately L:10μH, C: 10μF to 100μF) to the output.
(*) Constants are reference values.
Need to be adjusted based on noise effects.

1-5. Proposed PCB layout
A proposed layout (example) for these circuits is shown in Figure 1-9.

- Single-sided board, lead component view
- Components in red are surface-mounted components
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