1 Scope

The scope of this application note is to review:

- the design process to realize a CISPR25 class 5 compliant reference board
- and the actual measurement results

for the MLX10803 in a Buck Boost topology.

2 Introduction

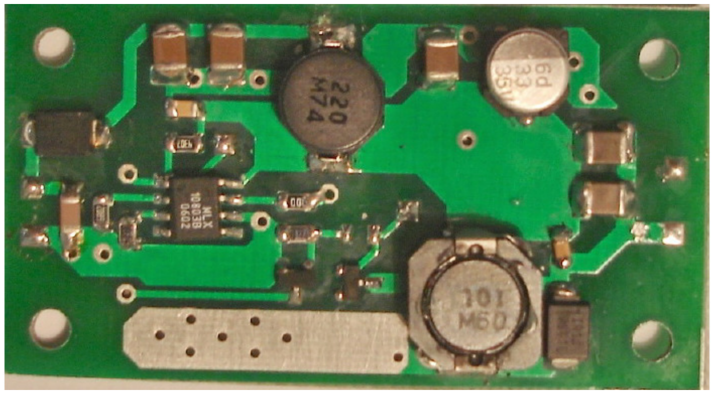
The reference design board has been developed for a 4W blinker application with 12 amber/red LEDs at 150mA, but can be applied for any Buck-Boost signaling applications.
Typical examples are Daylight Running Lights (DRL), Center High Mounted Stop Light (CHMSL), Rear and Front Fog lamp, Reversing or Back up lamp, … .

The applied bill of materials has not been optimized for cost.
The reference design is intended to give the best possible result for a wide range of buck-boost applications from 1W to 12W or more.

It should be used as a starting point to subsequently depopulate the board until the required EMC specification is achieved.

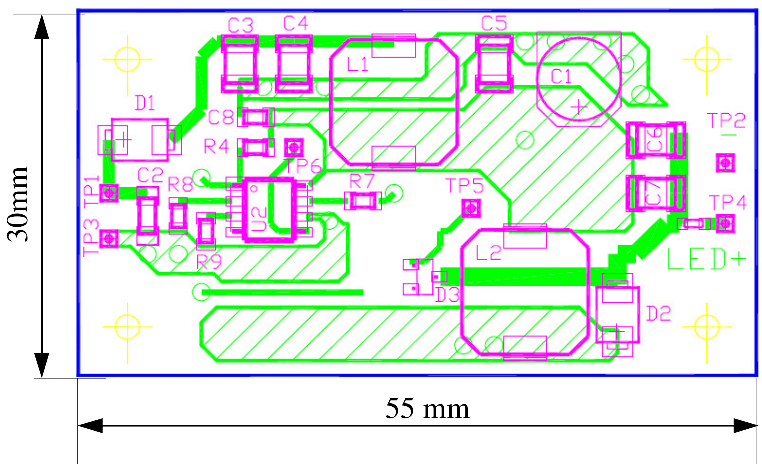
3 Table of contents

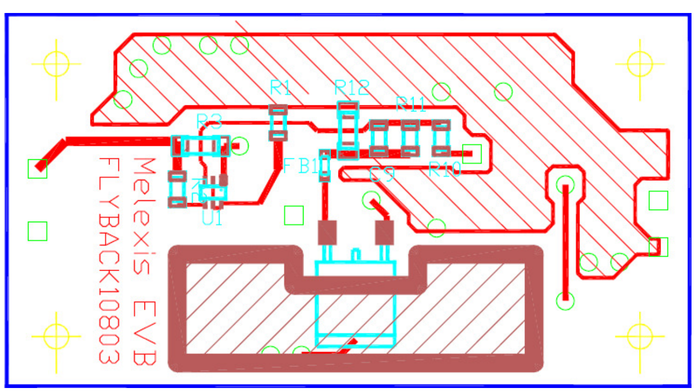
4 Evaluation board EVB10803_5



5 PCB Layout

EVB10803_5 has 2 layers, 1mm thick FR-4.





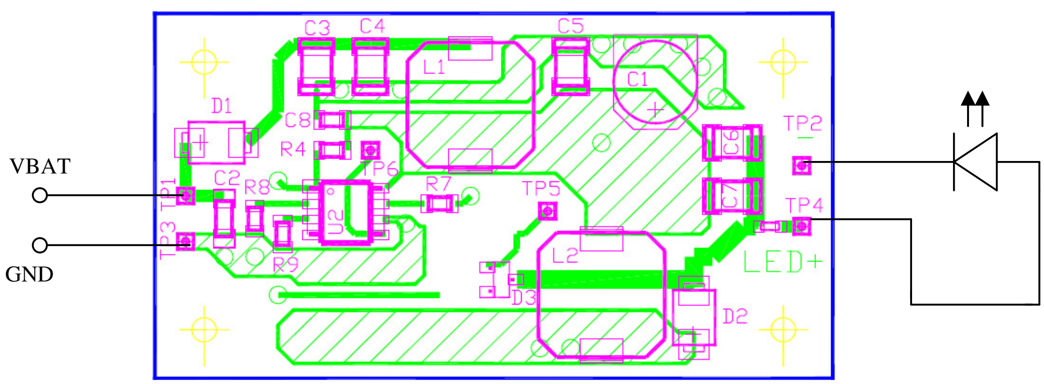
Components R5, R6, Q1 and C9 are not populated
TP5 and TP6 are shorted by a wire.

6 Connector Pin definition

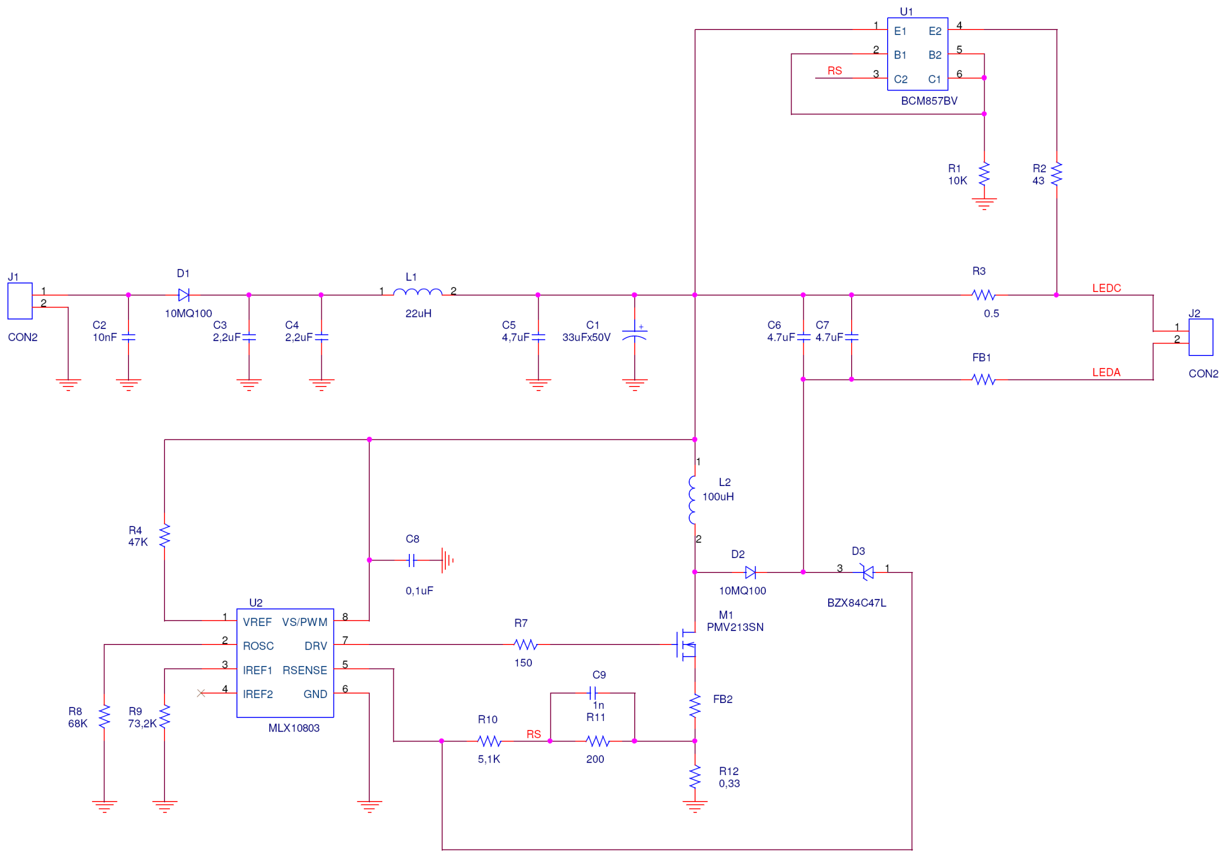
Connector J1	Signal	Connection
Pin 1	VBAT	power supply [6..32]V
Pin 2	GND	ground
Connector J2	**Signal**	**Connection**
Pin 1	LED-	LED cathode
Pin 2	LED+	LED anode

7 Minimal board connection

The Supply pins on connector J1 can be directly connected to the battery. The LED's anode has to be connected to the pin LED+ of connector J2 and the LED's cathode has to be connected to the pin LED-.



8 Application circuit



Reference Design Application Schematic:
25V forward voltage, 150mA LED current

9 BOM

Q-ty	Ref.	Part	Part Spec.	Package	Manuf.
1	C1	33uFx35V	Auminium electr. Cap.	D-size	
1	C2	10nF	X7R,+/-10%,50V	1206	
2	C3,C4	2,2uF	X7R,+/-10%,50V	1210	
3	C5, C6, C7	4,7uF	X7R,+/-10%,50V	1210	
1	C8	100nF	X7R,+/-10%,50V	805	
1	C9	0,1uF	X7R,+/-10%,50V	805	
2	D1,D2	10MQ100	Shottky diode		
1	D3	BZX84C47V	Zener diode	SOT-23	
2	FB1, FB2	BLM18PG331	ferrite bead	603	Murata
1	L1	22uH	CDRH8D28N-220N		Sumida
1	L2	100uH	CDRH104R-101N		Sumida

1	M1	PMV213SN	100V, MOSFET	SOT-23	Philips
1	R1	10K	1/8W,+/- 5%,50V	805	
1	R2	43	1/8W,+/- 1%,50V	805	
1	R3	0,5	1/4W,+/- 1%,50V	1206	
1	R4	47K	1/8W,+/- 5%,50V	805	
1	R5		Not populated		
1	R6		Not populated		
1	R7	150	1/8W,+/- 5%,50V	805	
1	R8	68K	1/8W,+/- 1%,50V	805	
1	R9	73,2K	1/8W,+/- 1%,50V	805	
1	R10	5,1K	1/8W,+/- 5%,50V	805	
1	R11	200	1/8W,+/- 1%,50V	805	
1	R12	0.33	1/4W,+/- 1%,50V	1206	
1	U1	BCM857BS	matched pair PNP	SOT-363	Philips
1	U2	MLX10803	LED driver	SO-8	Melexis

10 Battery Voltage operating range

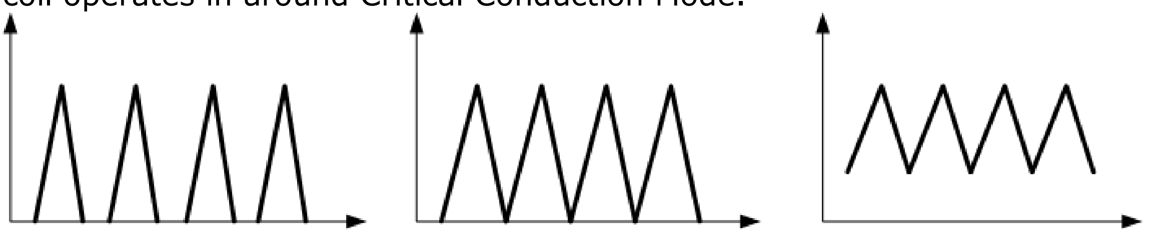
In this application note we will consider 2 operating ranges:

- Normal operating range: 9V to 16V
- Reduced Operating range: 6V to 19V outside of the Normal operating range.

11 Component selection

11.1 Imax Polynomial

In first order we neglect the losses in the driver, the Diode D2 and shunt resistor R3. To minimize switching losses and FET heating we assume the coil operates in around Critical Conduction Mode.



Power equation for Critical Conduction Mode: $LI_{max}^2 / 2*(Ton+Toff) = Pled$

where

$Pled$ - power dissipated in the LED string
$\mathrm{Im}ax$ - max. coil current , value determined by resistor R9 (IREF1)
$Toff$ - monoflop time , value determined by resistor R8 (Rosc)
Ton - coil charge time to reach Imax.

The Ton time depends on the input voltage (Vin)

$$Ton = \frac{L*\mathrm{Im}ax}{Vin}$$

This leads to the following **polynomial** equation:

$$\mathrm{Im}ax^2 - \frac{2*Pled*\mathrm{Im}ax}{Vin} - \frac{2*Pled*Toff}{L} = 0$$

11.2 Coil selection, Rosc selection

Assumptions for the reference board design:

- ⇨ Total forward voltage 25V at 150mA => Pled = 3.75W
- ⇨ For best EMC performance according to CISPR25 the switching frequency is kept below 150kHz (See EMC review at the end of this application note). Therefore we will operate the coil as close as possible to critical conduction mode (CCM). Remark that the Imax polynomial is only valid when the coil is not in discontinuous mode.
- ⇨ By minimizing Imax cheaper coils can be used. According to the polynomial this can be done by:
  - o Minimizing Toff
  - o Maximizing L (but this adds cost)

  In this reference design we have chosen available components with some safety margin by using
  - o Toff = 4us
  - o L = 100uH

The maximum coil peak current (Vbatmin = 8 to 9V) can then be calculated from the polynomial to be: **Imax ~ 1.2A**

The applied coil is a Sumida CDRH104R-101N with Isat (65%) = 1.35 A

Remark when choosing a smaller inductance value:

- Reducing L increases the peak current through the coil. However the specified saturation current for smaller L values typically rises faster than the increase in required peak current.
  For instance the 68uH CRH104R0680N has Isat = 1.5A, while the Imax only increases to 1.3A.
- A smaller inductance value has a steeper slope dI/dt. When L is chosen too small for a given monoflop time the coil current will reach zero before the monoflop time is over. Therefore for the same Toff time the coil may move into discontinuous mode.
  When operating in strong discontinuous mode the above polynomial is no longer applicable.

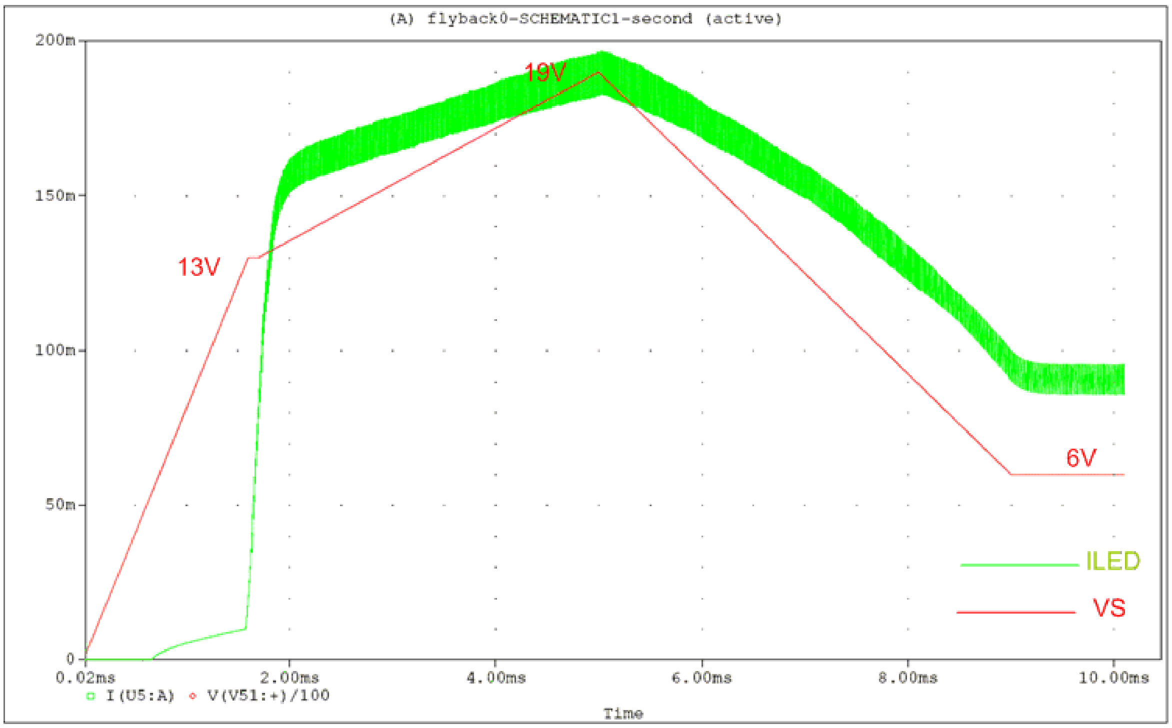
11.3 Rref selection

The selection of Rref is linked to the feed forward compensation network that will be discussed in the next paragraph.
In order to allow maximum freedom for designing the feed forward compensation network RREF = 73.2kOhm is selected.
This value gives an input voltage as close as possible to the 3.6V maximum input voltage on IREF.

12 Compensation network for VBAT variations

12.1 Intro

In a Buck Boost topology the LED current is largely dependent on the input voltage (Vbat).



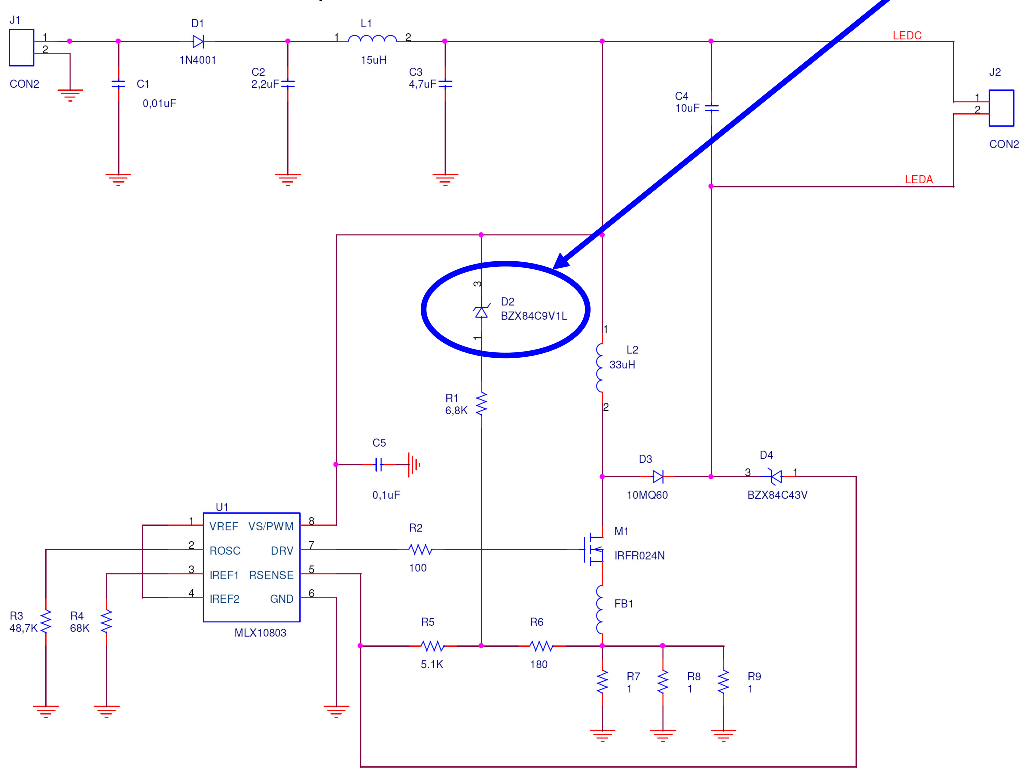
Vbat = VS : Red line (scale 1/100).
Uncompensated LED current = green line.

By influencing the input voltage on the RSENSE input pin, a feed forward compensation can be realized.

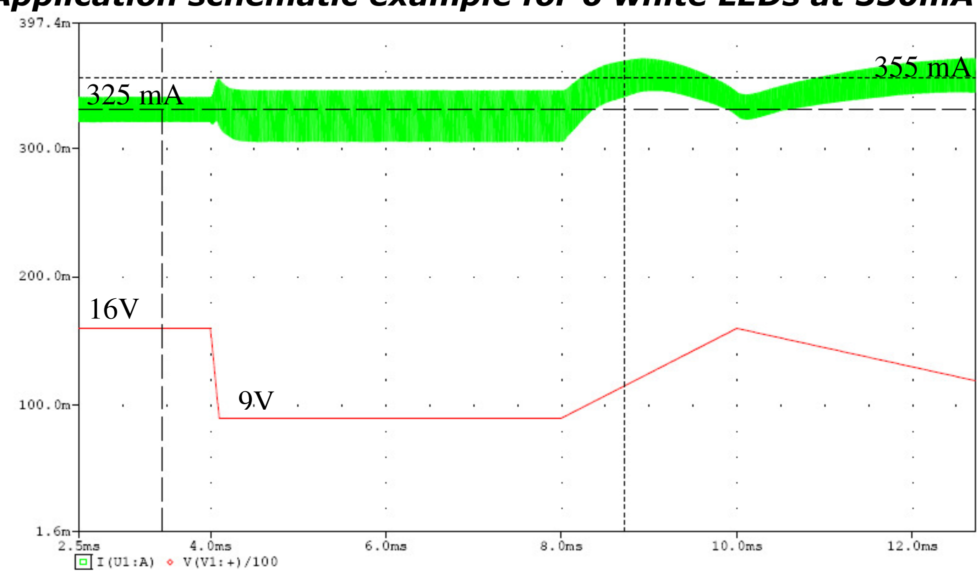
To allow maximum compensation span, the switching point is set to ~ 700mV, with 300 to 400mV drop over the shunt, and the rest over the feed forward compensation resistor.

12.2 Zener diode compensation

A low cost feed forward compensation can be realized using a (9V) Zener diode that injects additional current onto R8 as VBAT increases above 9V as shown in the below example schematic.

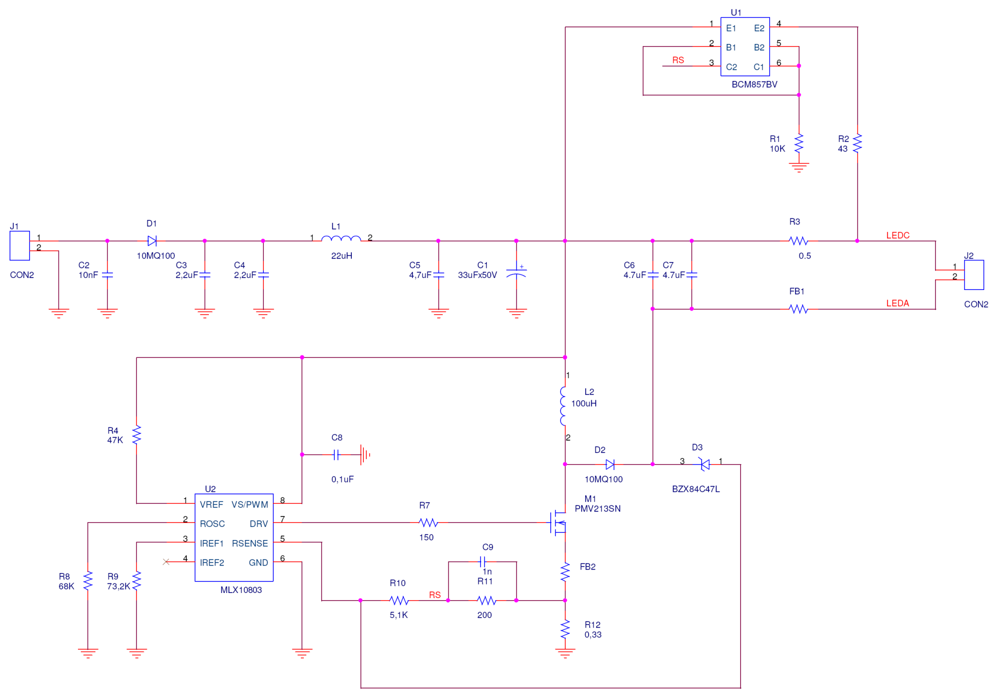


Application schematic example for 6 white LEDs at 350mA



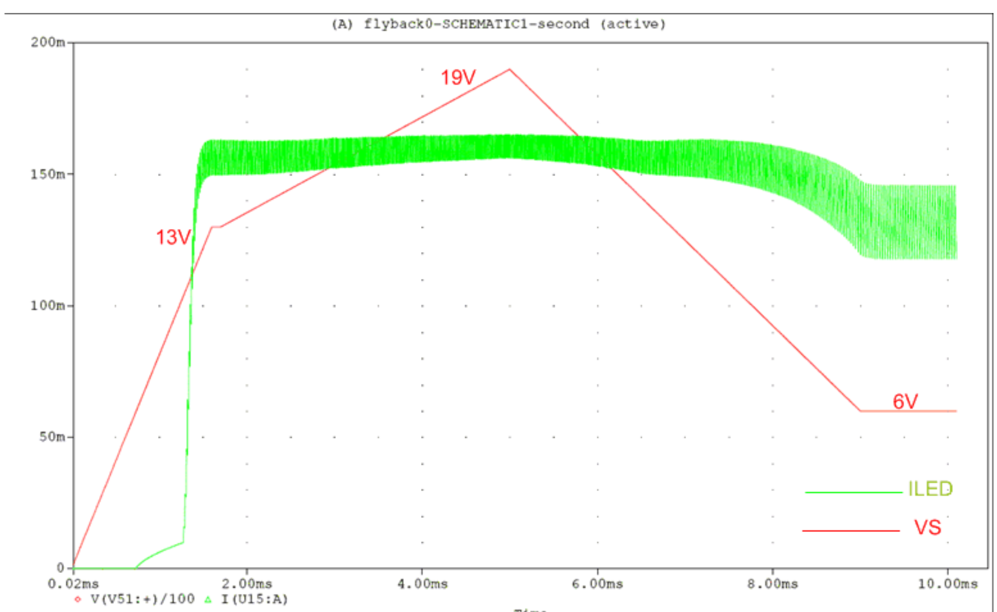
LED current(green) varies+/- 3.5% as a function of VBAT
355mA @13V,   330mA @9V and @16V

12.3 Current mirror compensation



Reference design Application schematic (same as p5)

In the reference design board a current mirror compensation network has been implemented, further reducing the LED current tolerance to less than +/-1.5% (see perform. measurement results at the end of the appl. note.)



Vbat = VS : Red line (scale 1/100):
Resulting LED current = green line.

The compensation network will modify the duty cycle as the inverse function of Vin.

Duty cycle: D = Ton / ($Toff$ + Ton),

where

- the monoflop time $Toff$ is fixed by R8 (Rosc).
- Ton is the time needed for the voltage on the Rsense pin to reach the peak value set by R8 (IREF1)
  The voltage on the Rsense pin is the sum of the voltages at R11 and R12

As Vbat increases, the LED current will have the tendency to increase. This increased current is copied onto R11 (+R12), and will reduce Ton.

This compensation network is realized using

- a low cost SOT363 current mirror with PNP transistors (BCM857)
- Resistor R1 determines the working point of the current mirror at the [0.8 .. 1.5]mA level and depends upon the input voltage (Vin).
- Shunt resistor R3 should be as small as possible to minimize the power dissipation. Here we target a total drop of 75mV.
  For Iled = 150mA, this gives R3 = 0.5 Ohm.

The current mirror copies the current through the shunt resistor R3 onto the resistor (R11+R12) with scale factor Iled*R3/R2
This creates an additional drop voltage V = Iled * (R11+R12) * R3/R2.

12.4 Coil Calc for Buck boost with Current mirror

BuckBoost_CM_Coilcalc.xls calculates the starter values for the external component values.

INPUT DATA		OUTPUT DATA	
Min. supply, VS	9	I coil_peak (VSmin) , A	1.16
Nom. supply, VS	13	I coil_peak (VSnom) , A	0.95
Max. supply, VS	16	I coil_peak (VSmax) , A	0.87
LED Forward Voltage, VF	2.2	Rosc, Kohm	68
Number of LEDs in series	12	Rsense, Ohm	0.33
Average current: LED , mA	150	Rshunt_CM, Ohm	0.50
Tmonoflop, uS	4.1	Min. switching frequency, kHz	59
L (the Coil), uH	100	Nom. switching frequency, kHz	88
		Max. switching frequency, kHz	105
		Avg. current FET (Vsmin), A	0.76
		Nom Output Capacitor Co, uF	6.7

13 Reduced operating range

13.1 Operation below 6V

The MLX10803 operation is specified down to 6V.
In order to operate the IC down to Vbat = 6V, a PMOS reverse polarity protection can be applied instead of diode D1 which has a drop voltage of 0.6V at 0.5A.
For VS < 6V (typ around 5.3V) the 50uA from the current sources on IREF are no longer guaranteed which may cause a flickering behavior as the voltage drops. The IC power down level is below 3V.
To avoid this flickering below 6V the MLX10803 can be shut down by pulling VREF to ground.

13.2 Operation from 6 to 19V

Below table shows measurement results from 6V to 19V, using a LED string with total forward voltage = 25V.

VS = Vbat [V]	**6**	**7**	**8**	**9**	**13**	**16**	**19**
Iled [mA]	110	130	140	148	153	151	144
Pled [W]	2.71	3.24	3.56	3.7	3.825	3.775	3.6
Icoilpeak [A]	1.23	1.23	1.21	1.15	0.99	0.9	0.81

Remark that VS = VBAT, meaning that the reverse polarity diode D1 has been removed (shorted) for these evaluations.

Remarks:
The feed-forward compensation network ensures that

- the LED current is independent from VBAT variations within the normal operating range [9, 16]V.
- the LED current drops in a monotonous way outside [9, 16]V range.
- the coil peak current does not increase significantly as Vbat reduces below 9V

14 Protection circuits

14.1 Reverse polarity protection

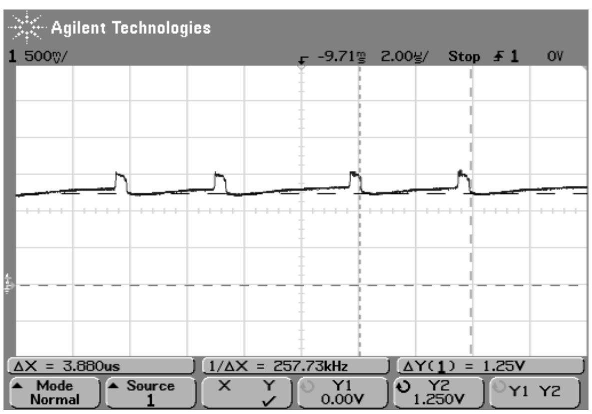
Reverse polarity protection on this reference pcb EVB10803-5 is realized using an input diode D1.
This is a low cost solution, but implies

- an additional voltage drop (~0.6V)
- reduction of overall efficiency (5% at Vbat = 13V)
- increased dissipation on the board

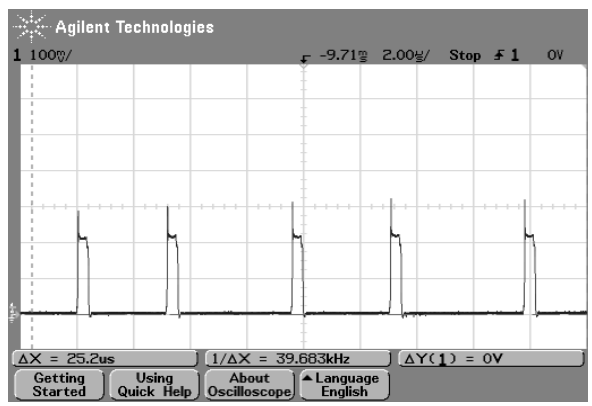
If one of these effects is not acceptable, a MOSFET should be used instead.

14.2 Shorted load condition for current mirror compensation network

In case the Load (CON2) is shorted (for instance both wires going from the driver pcb to a remote LED string are shorted), the supply caps C5 and C6 are shorted.
With only D2 as load, the output voltage is smaller than the input voltage, therefore the charge time of the coil will be shorter than the discharge time, which equals Toff.



plot 1. Voltage on Rsense input pin



plot. 2 Voltage over R12 (0.33Ohm)

However in case a current mirror compensation network is applied, this will copy any increase in LED current onto R11.
On scope plot 1 the 1.25V indicates 6.25mA over R11. The 200mV ripple is due to the additional drop over R12 (see plot 2.) while the FET is on.
Because the voltage on RSENSE is continuously larger than IREF/5 = 720mV, the debounce time defines the switch on time.
The switching frequency of ~230 kHz corresponds to the 4us monoflop time + 300ns debounce time.

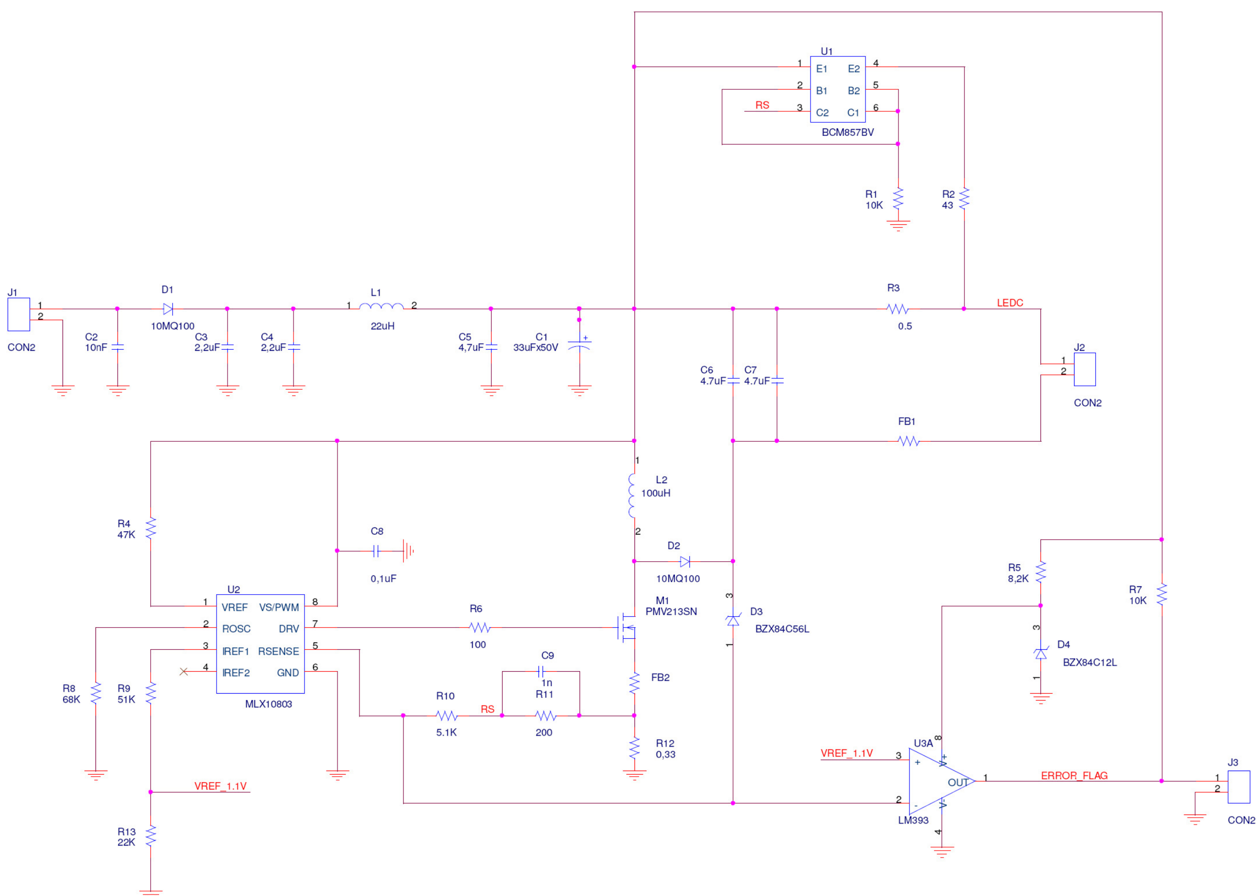
14.3 Open load detection

In case the load is an open circuit (load disconnected, or a failing LED that breaks the string) the Buck-Boost will pump up the voltage over the supply capacitors.
Zener D3 will start to conduct additional current over R10+R11+R12 pulling the voltage on RSENSE above 0.7V.

14.4 Diagnostics feedback

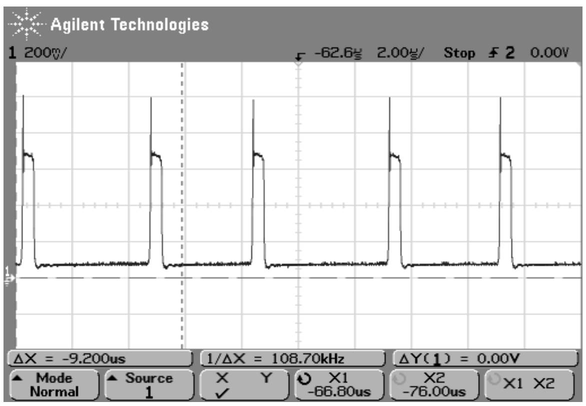
The MLX10803 continues switching at ~ 230kHz (randomized) during open load and shorted load conditions, and pulls small amounts of current from the supply line which are not critical for EMC behavior.

A simple method to provide diagnostic feedback for these conditions can be realized using a comparator that monitors the voltage on the RSENSE input pin.   This comparator is not foreseen on the reference design board.

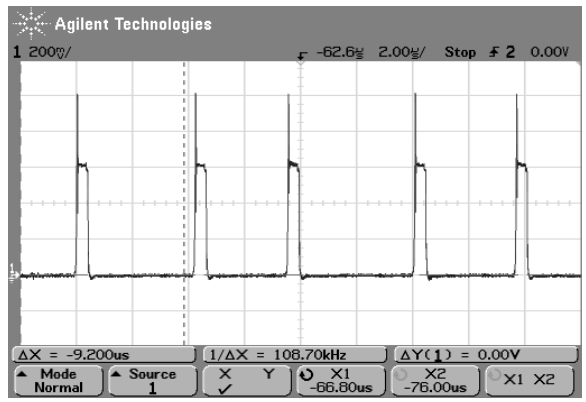


14.5 Shorted load condition for zener compensation network

Unlike with the current mirror, the zener compensation circuit gives no direct feedback from the LED current onto the RSENSE pin.
RSENSE will reach its peak voltage only by an increase of the current over the shunt R12.
From plot 2 we find that - instead of 200mV in case of the current mirror - the voltage over R12 is now 600mV, indicating that the energy taken from the supply line will be 3 times larger, leading to a small increase in conducted emissions compared to the current mirror solution.



plot 1. Voltage on Rsense input pin



plot 2 Voltage over R12 (0.33Ohm)

The switching frequency remains ~230 kHz corresponding to the 4us monoflop time + 300ns debounce time.

15 Filter stages

15.1 Low frequency input filter

To meet the EMC requirements in the low frequency band an LC-input filter is used (C1, C3, C4, L1). C4 is an aluminum (high ESR) capacitor to 'damp' the cut-off frequency of the LC filter.
An additional bypassing capacitor C2 is populated close to the input connector.

Remark that for several VM's with less stringent EMC requirements we can remove the input filter coil L1, which is saves significant cost.

15.2 Output filter

The output capacitors C6, C7 supply the LED current while the coil is charging. Their value influences the LED current ripple.

15.3 RF filter

The FB1, FB2 ferrite beads reduce the level of noise disturbance in the FM radio band [60 – 108]MHz.

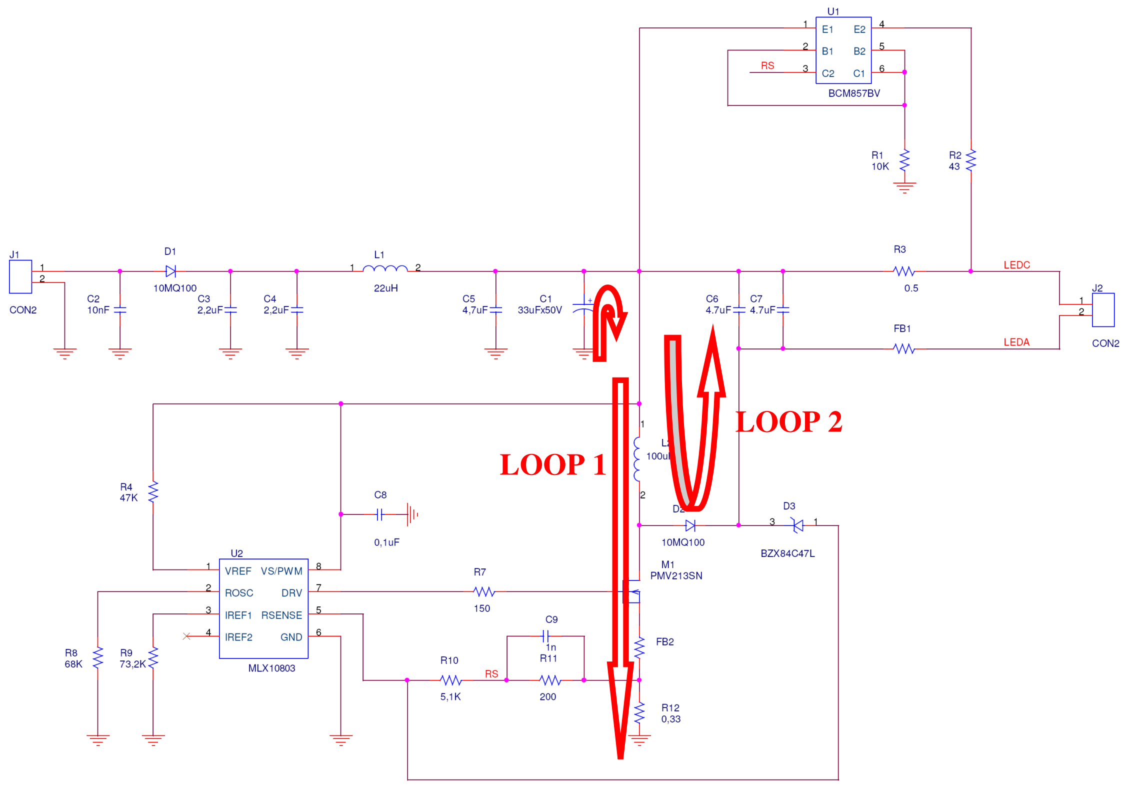
15.4 FET gate voltage slew rate

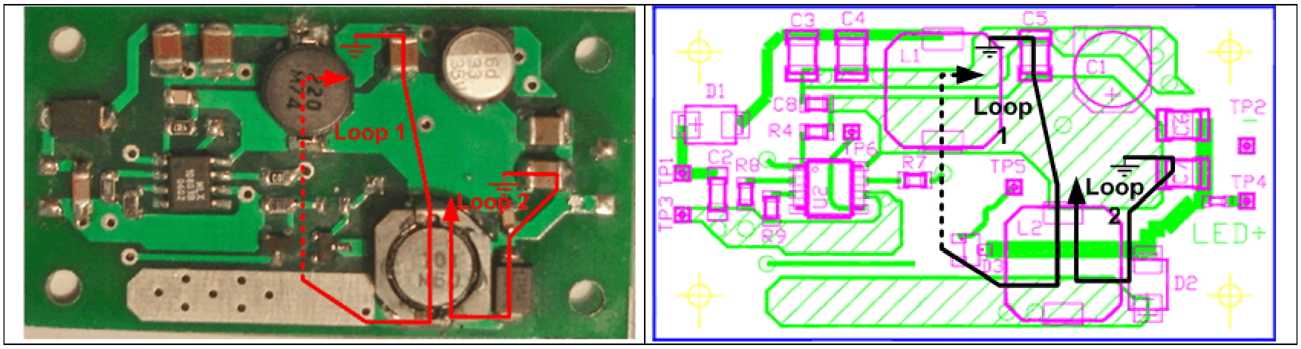
The FET gate voltage slope has a major impact on the FM radio band [60 – 108]MHz. Slow slopes are good for EMC, but increase the switching losses, and may overheat the FET.

16 Layout considerations

In order to minimize EMI the area of the current loops is minimized.

- Current loop 1: Coil charging
- Current loop 2: Coil discharge





The drain area of the FET is designed large to serve as a heatsink, because the original intention was to make a general purpose board up to 15W. This is also why a DPAK footprint is foreseen.
For optimal radiated emission performance, it is recommended to minimize this area and position it in the center of the pcb.

17 Measurement results: LED current tolerance + efficiency

17.1 Measurement setup

To evaluate the thermal performance of the driver electronics over temperature, independent from the LEDs, the driver electronics were placed in the oven, while the LED string remained under room temperature.

To evaluate the efficiency results of the driver electronics the protection diode D1 was shorted.

17.2 Measurement results

- Vin: battery voltage
- Iin: Average current consumption from battery
- Pled: Power dissipated by the LEDs string
- Iled: Average LED current
- Eff: Driver efficiency
- Tol: Current tolerance
- Vled: Total LED string forward voltage = 25V

Temperature	22C		100uH					
VIN(V)	9	10	11	12	13	14	15	16
IIN(mA)	470	436	396	365	335	313	289	268
PIN(W)	4,23	4,36	4,36	4,38	4,36	4,38	4,34	4,29
PLED(W)	3,65	3,73	3,79	3,82	3,82	3,84	3,81	3,78
ILED(mA)	148	150	152	153	153	153	152	151
eff	86%	86%	87%	87%	88%	88%	88%	88%
tol	1,96							

Temperature	-10C		100uH					
VIN(V)	9	10	11	12	13	14	15	16
IIN(mA)	490	452	413	380	349	324	300	279
PIN(W)	4,41	4,52	4,54	4,56	4,54	4,54	4,50	4,46
PLED(W)	3,78	3,86	3,92	3,95	3,95	3,95	3,93	3,89
ILED(mA)	153	156	158	158	158	158	157	156
eff	86%	85%	86%	87%	87%	87%	87%	87%
tol	1,90							

Temperature	-40C		100uH					
VIN(V)	9	10	11	12	13	14	15	16
IIN(mA)	511	471	427	390	357	328	303	281
PIN(W)	4,60	4,71	4,70	4,68	4,64	4,59	4,55	4,50
PLED(W)	3,95	4	4	4	4	4	3,97	3,91
ILED(mA)	160	161	161	161	160	159	158	156
eff	86%	85%	85%	85%	86%	87%	87%	87%
tol								

Temperature	+70C		100uH					

VIN(V)	9	10	11	12	13	14	15	16
IIN(mA)	461	425	389	359	331	306	285	266
PIN(W)	4,15	4,25	4,28	4,31	4,30	4,28	4,28	4,26
PLED(W)	3,51	3,6	3,66	3,7	3,72	3,72	3,71	3,7
ILED(mA)	143	145	147	148	148	147	146	145
eff	85%	85%	86%	86%	86%	87%	87%	87%
tol	1,35							

Summary of influence of the temperature change of the electronics on the LED current at VBAT = 13V:

Oven Temperature	-40C	-10C	22C	+70C
ILED(mA)	160	158	153	148

This Monotonous negative temperature dependence can be resolved, similar to compensating the LED current temperature dependence: by applying a PTC resistor in series with resistor R9.

18 EMI measurements

The EVB10803-5 is designed as a reference design board that allows customers to evaluate the EMI performance.
From experience radiated emissions have not been a problem; therefore we have only included the conducted emission results.
The measurements were executed

- according to CISPR25,
- for narrow band,
- with the Peak values mentioned.

The applied load was a LED string at 150mA with total forward voltage of 25V.

18.1 Conducted emission standards

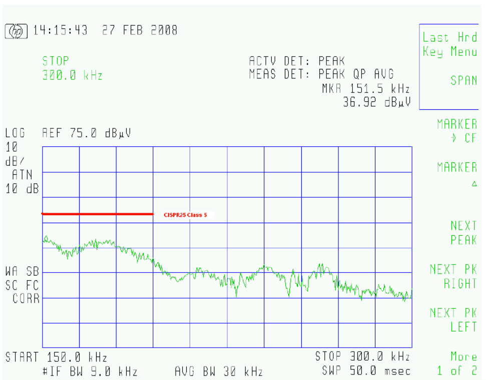
Table 7 of CISPR25 gives the test limits for conducted narrowband voltage emissions. Following is a copy of this table:

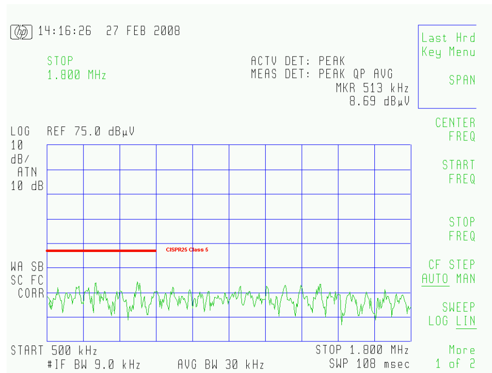
Table 7 – Limits for narrowband conducted disturbances (peak detector)

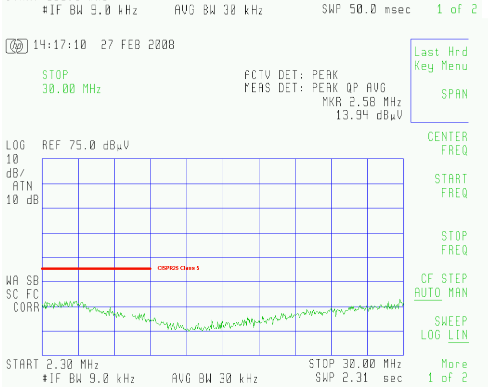
Class	Levels dB(µV)					
	0,15 MHz to 0,3 MHz	0,53 MHz to 2,0 MHz	5,9 MHz to 6,2 MHz	30 MHz to 54 MHz	68 MHz to 87 MHz Mobile services	76 MHz to 108 MHz Broadcast
1	90	66	57	52	42	48
2	80	58	51	46	36	42
3	70	50	45	40	30	36
4	60	42	39	34	24	30
5	50	34	33	28	18	24

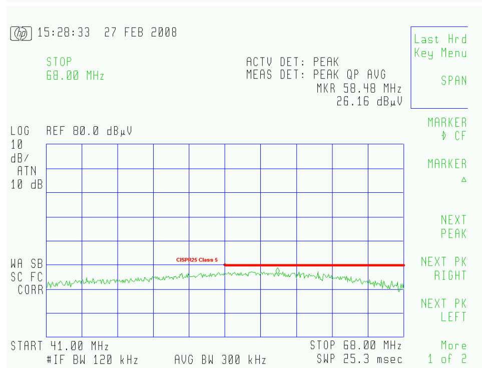
18.2 Measurement results

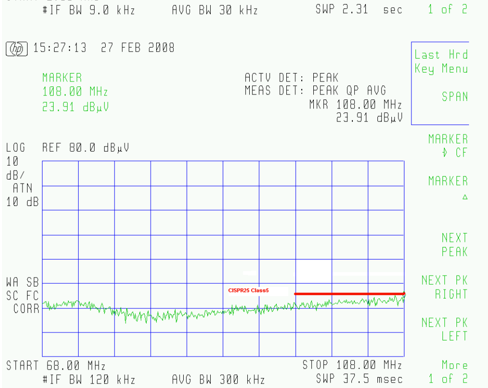
The following drawings show the results of the CE measurements for the EVB10803-5 board. RED line indicates Class 5 level.









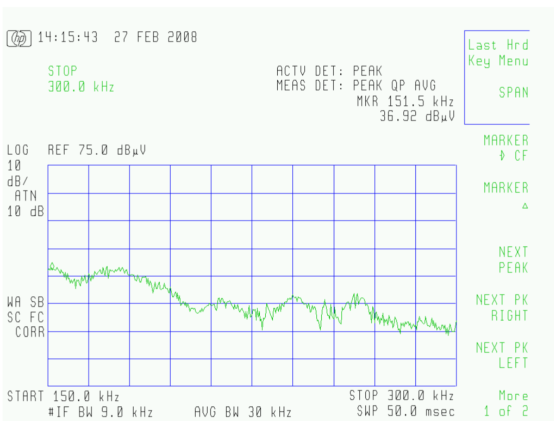


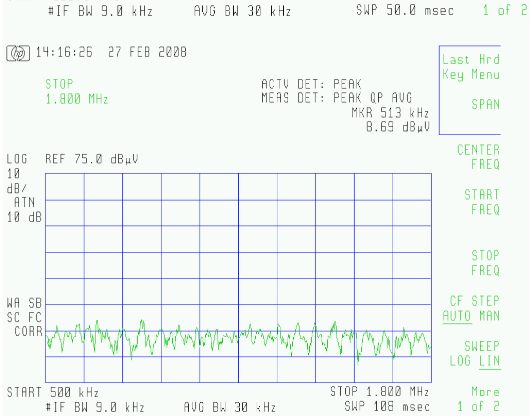
Remark:
The CE measurements for the EVB10803-5 board <u>with disconnected load</u> remained below the CISPR25 class 5 requirements.

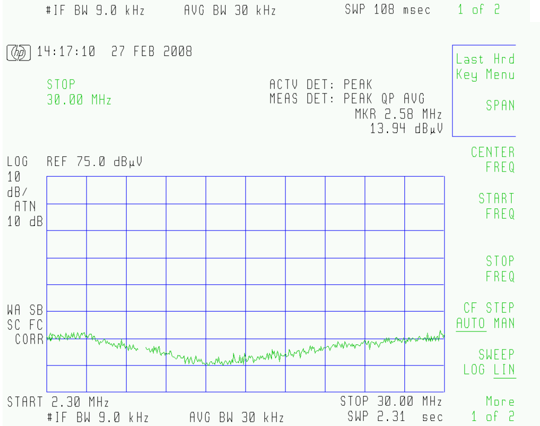
19 Comparison 4W application with/without input filter coil

For customers with >75 dBuV requirements in the lower frequency band, the input filter coil L1 can be removed.

With input filter coil L1







Without input filter coil L1