

Tunable white LED drivers with BCR601 and BCR602

How to design cost effective, highly efficient LED drivers of highest light quality



About this document

Scope and purpose

LED lighting has become a mature technology, with stringent requirements for light quality. The requirements regarding input power and light quality can be fulfilled most easily by two-stage topologies. A combination of high power factor flyback as the first stage with a buck as the second stage has turned out to offer flexibility and high performance.

This application note shows that the buck on the secondary can be replaced with a linear regulator with active headroom control (AHC) without sacrificing efficiency. At the same time the linear regulator leads to the highest possible light quality with a very cost effective approach. The cost advantage compared to buck topology becomes most obvious in tunable white and multichannel drivers.

Intended audience

LED driver design engineers, field application engineers.

Table of contents

About this document.....	1
Table of contents.....	1
1 Linear constant current regulators in LED drivers	3
1.1 Linear regulator with active headroom control	3
1.2 LED current setting and dimming circuit	5
1.3 Efficiency optimization	6
1.4 Setting V_{HR} for best efficiency	8
1.5 MOSFET and BJT selection	9
1.6 Overvoltage protection	10
1.7 Overtemperature and hot-plug protection	10
1.8 Output short-circuit protection	11
2 Multichannel LED driver with BCR601 plus BCR602	14
2.1 Setting V_{HR} in multichannel drivers.....	17
2.2 Dimming, overvoltage, short-circuit and hot-plug protection in multichannel drivers.....	17
3 Tunable white reference design with BCR601 and BCR602	18
3.1 Board setup	18
3.2 Board specifications.....	18
3.3 Schematics and circuit description	19
3.4 Electrical performance.....	23
3.4.1 Light quality.....	23
3.5 LED current stability and regulator efficiency.....	23

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Linear constant current regulators in LED drivers

3.6	Light quality.....	24
3.7	System performance	26
3.8	Bill of materials.....	28
4	References	30
	Revision history.....	31

1 Linear constant current regulators in LED drivers

As the importance of light quality continues to increase, the popularity of LED driver topologies that provide an almost perfect DC current at the output is increasing as well. This is because any low-frequency fluctuation of the LED current will show up as flicker or cause stroboscopic effects, while high frequencies may lead to interference with other equipment such as cameras or scanners. This results in a preference for two-stage topologies, especially a single-stage power factor correction (PFC) flyback on the primary, followed by a buck stage at the secondary side[2].

Designers typically do not consider linear regulators as second stage, since it is difficult to achieve high efficiency with these. On the other hand, linear regulators can control the LED current to be perfect DC, resulting in the best light quality possible, at least as far as the driver is concerned. Moreover, with AHC, a linear regulator can achieve efficiencies on par with a buck stage.

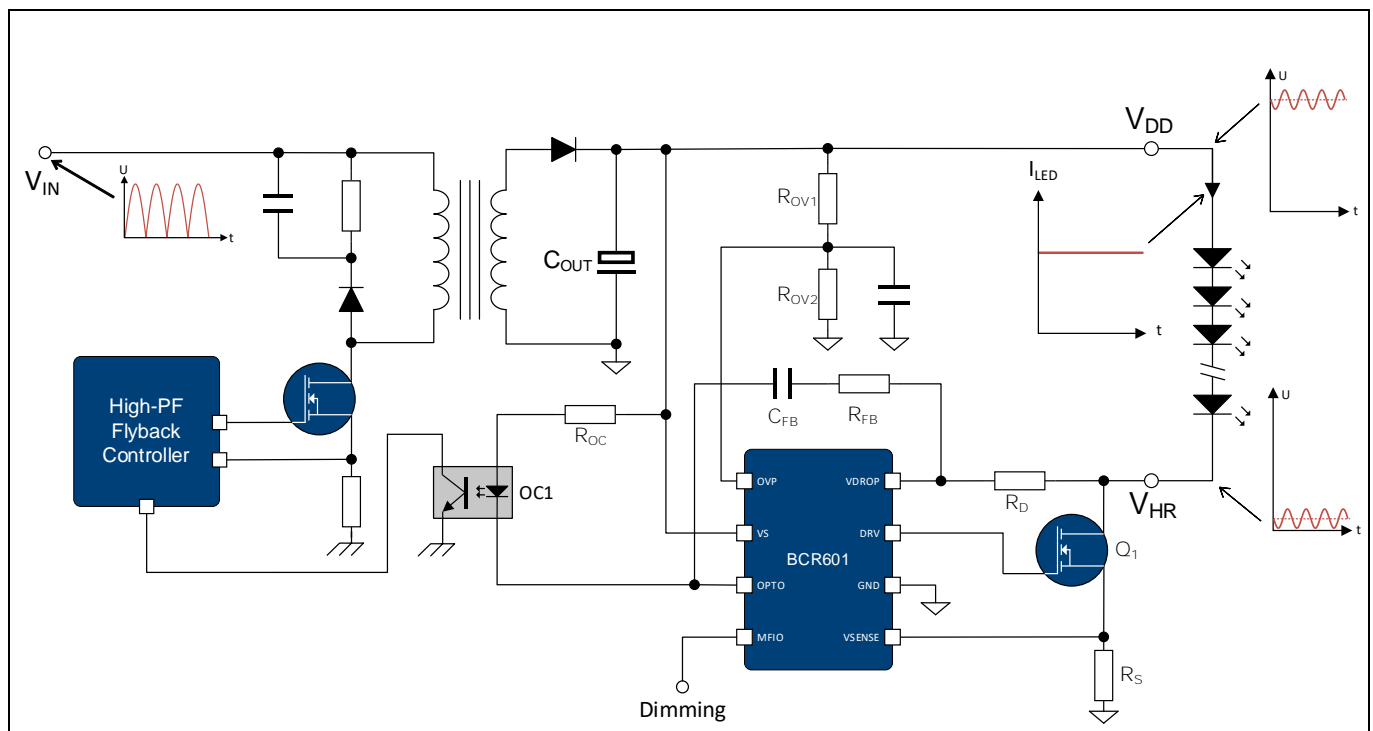


Figure 1 LED driver with linear LED current regulator and AHC

1.1 Linear regulator with active headroom control

A system with AHC contains two regulation loops. The first loop, in **Figure 1** consisting of Q_1 , R_S and an internal regulator of **BCR601**, stabilizes the LED current to a value determined by the value of R_S and the internal reference of 400 mV in undimmed state. At constant forward current and for a given temperature, the LED forward voltage is constant as well. The difference between the output voltage V_{DD} of the first stage and the LED string forward voltage V_{LED} is the drain voltage of Q_1 or headroom voltage V_{HR} .

The system described so far contains several parameters that vary a lot with component tolerances and operating conditions. The output voltage of the first stage for example has a certain tolerance, say ± 3 percent, and an additional ripple with an amplitude that depends on the actual value of C_{OUT} (which means on its tolerance). At the same time V_{LED} has a high tolerance and varies greatly with junction temperature. Consequently, the value of Q_1 drain voltage can easily vary by several volts over all possible operating conditions. Additionally, in order to have enough headroom to regulate these variations, the average drain

Linear constant current regulators in LED drivers

voltage of Q1 needs to be relatively high. Depending on the value of the LED current, this can result in a huge power dissipation of Q1.

Now the second feedback loop is used to reduce this overhead voltage close to the minimum possible. To achieve this, the drain voltage is fed via resistor R_D to an error amplifier that controls V_{DD} , the output voltage of the first stage, via the optocoupler OC1. The RC network $R_{FB}-C_{FB}$ makes this loop slow enough to regulate the average drain voltage only, and not to react to the ripple voltage of double the line frequency.

With this approach, the drain voltage essentially consists of the output ripple of the first stage with a certain DC offset (V_{HR} in [Figure 1](#)).

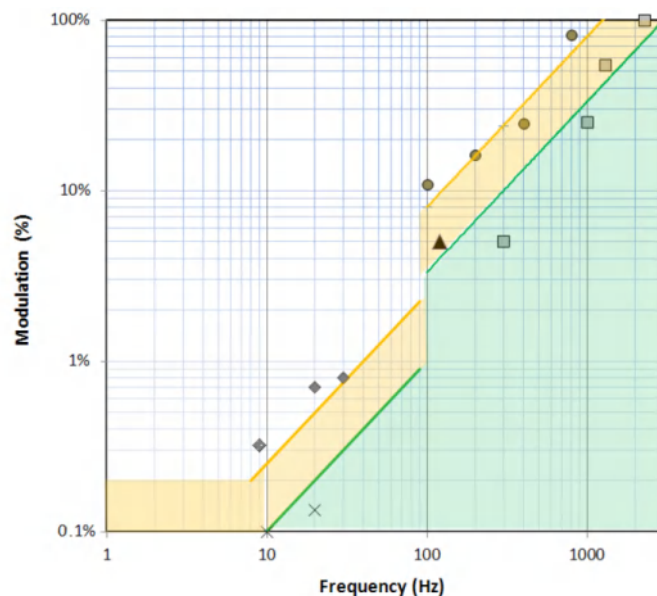
The power loss of Q_1 has two main contributors: conduction losses due to $R_{DS(on)}$ and the loss due to the inherent output voltage ripple V_{RIP} of the first stage, that has double the line frequency. Actually, the loss due to V_{RIP} is typically the largest contribution, while conduction loss can easily be reduced with lower $R_{DS(on)}$ MOSFET.

Without going into too much detail, it can be said that V_{RIP} is determined by the physics of the PFC function and so Eq. 1 is valid for virtually any PFC stage that delivers very high power factor.

$$V_{RIP} = \frac{I_{LED}}{4 \cdot \pi \cdot f_{LINE} \cdot C_{OUT}} \quad \text{Eq. 1}$$

That means there is no way to reduce V_{RIP} other than by using a higher capacitance value for C_{OUT} .

As will be shown in section 1.3, it is possible to achieve efficiencies of the linear regulator stage above 95 percent if C_{OUT} is chosen properly. Then the losses and in turn the efficiency of the linear regulator are on par with switching regulators while still giving a considerable cost advantage.



Note: Modulation (%) is defined as the difference between maximum and minimum luminance divided by the sum of maximum and minimum luminance (multiplied by 100).

Figure 2 Recommended modulation (%) vs. frequency (from [1]). Yellow area: low-risk, green area: no observable effect.

Linear constant current regulators in LED drivers

Because the LED current provided by a properly designed linear regulator is a pure DC, the light is free of any flicker, stroboscopic effects or similar. The effects of light modulation have been the subject of extensive investigations, and led for example to the recommendations of IEEE 1789-2015 [1]. Not only are the requirements for 100/120 Hz, a modulation depth smaller than 8 percent, difficult to fulfill with other topologies, but the latter is the biggest challenge for single-stage topologies, while dual-stage topologies excel with 100/120 Hz suppression. But for both single- and dual-stage topologies the light modulation under dimming conditions can be a challenge. Deep analog dimming, which means without PWM-like modulation of the output current, is very difficult to achieve with both. The dimming method of BCR601/602 is fully analog down to 3 percent. That means the absence of any AC component, apart from some random noise in the light, persists over the whole dimming range.

It also means that the linear regulator topology is one of the few in the market, if not the only one, that can fulfill the requirement of IEEE1789-2015 under all operating conditions.

1.2 LED current setting and dimming circuit

The regulator for the LED current has a non-dimmed typical reference voltage of 400 mV. For a given maximum LED current R_S is then determined by:

$$R_S = \frac{0.4 \text{ V}}{I_{LED}} \quad \text{Eq. 2}$$

The nominal LED current can be reduced through dimming pin MFIO. As long as the voltage at this pin is equal to or higher than 3.3 V, the full LED current is regulated. The 400 mV setpoint, and with that the LED current, is reduced linearly from 100 percent to 3 percent of nominal as V_{MFIO} goes from 3.3 V down to 0.2 V.

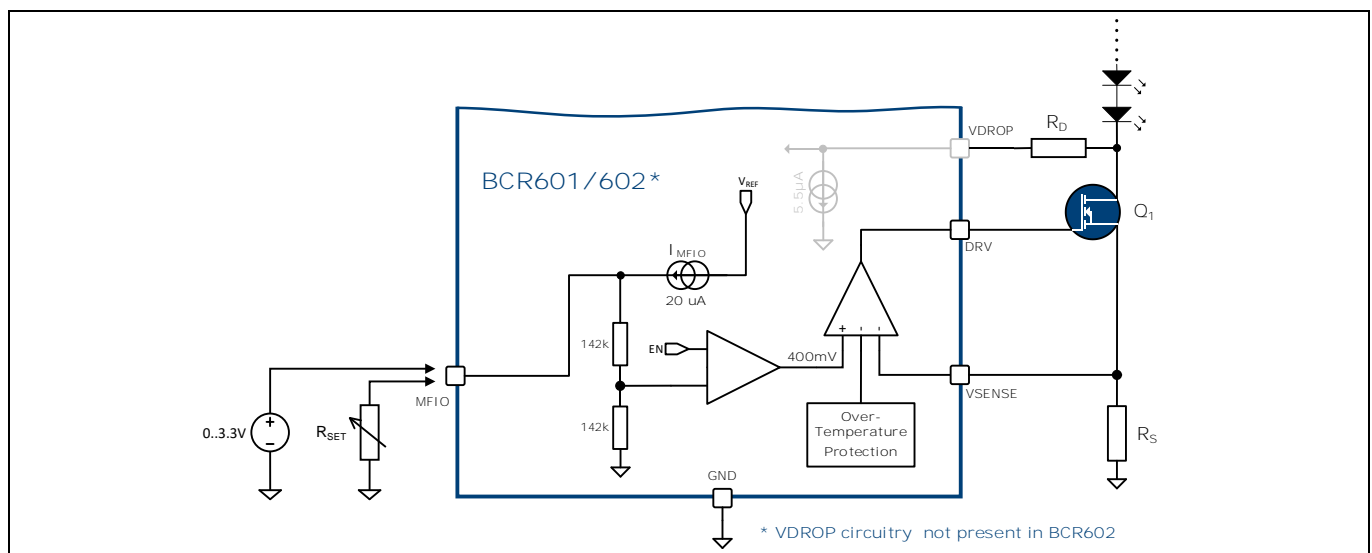


Figure 3 BCR601/602 LED current regulator and dimming circuit – dimming by applying a DC voltage or by a variable resistor R_{SET} connected to MFIO

When V_{MFIO} is lowered further the current stays at 3 percent until V_{MFIO} reaches 0.1 V. Below 0.1 V the LEDs are completely off. To prevent flicker the circuit has a built-in hysteresis of 0.1 V, i.e. raising V_{MFIO} from below 0.1 V will turn on LED current at 0.2 V only.

Linear constant current regulators in LED drivers

Dimming is also possible by applying a variable resistor R_{SET} to MFIO. **BCR601** has an internal current source driving a typical $20\ \mu\text{A}$ out of MFIO. This current causes a voltage drop across the external resistor, which means that the resistor value determines V_{MFIO} and in turn the dimming level. Due to the internal resistance of $285\ \text{k}\Omega$ the dimming level vs. external resistor has non-linear characteristics.

$$V_{MFIO} = \frac{R_{SET} \cdot 285\ \text{k}\Omega}{R_{SET} + 285\ \text{k}\Omega} \cdot 20\ \mu\text{A} \quad \text{Eq. 3}$$

In non-dimmable LED drivers with configurable output current, a fixed R_{SET} can be used for configuration instead of altering R_S . The output current selected by R_{SET} can be overridden at any time by a voltage source connected to MFIO.

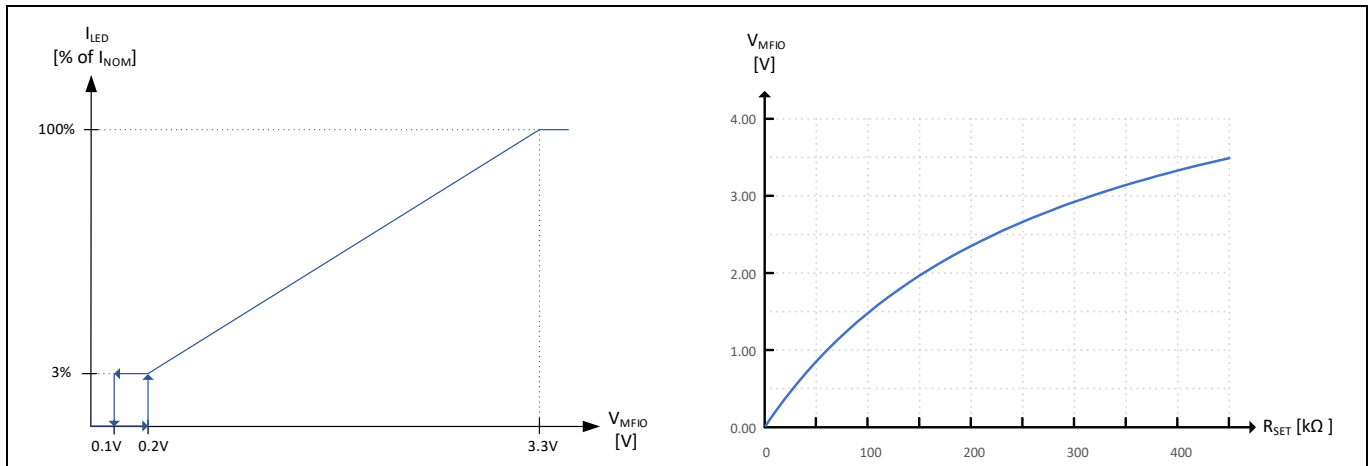


Figure 4 Output current I_{LED} vs. V_{MFIO} (left) and V_{MFIO} vs. R_{SET} (right)

1.3 Efficiency optimization

The total power loss of the LED current regulator with AHC is given by the losses in the sense resistor P_{loss,R_S} and the losses in Q1 are at least:

$$P_{loss,tot} \geq P_{loss,Q1} + P_{loss,R_S} = \frac{I_{LED}^2}{4 \cdot \pi \cdot f_{line} \cdot C_{out}} + I_{LED}^2 \cdot R_{DS(o)n,Q1} + V_{RS} \cdot I_{LED} \quad \text{Eq. 4}$$

That means that all losses in the power stage increase with the current, and those in the transistor are even quadratic. The voltage V_{RS} across the sense resistor is typically 400 mV at full current, becoming smaller as the LEDs are dimmed. One observation in Eq. 4 is that neither input nor output voltage occur in the losses. That leads to the recommendation to use as high an LED voltage as possible for a given output power in order to maximize efficiency. This is illustrated in **Figure 5** for a constant output power of about 20 W. The total current drawn from the primary side is decreasing from about 0.61 A to 0.42 A as the output voltage rises from about 33 V (12 LED) to 49 V (18 LED). This alone gives an efficiency increase from 93 percent to almost 94.6 percent.

The biggest contribution to the power losses nevertheless comes from the output voltage ripple of the primary side. Consequently, reducing this ripple is the most effective way to increase the efficiency of a linear regulator with AHC. **Figure 6** shows this very clearly.

Linear constant current regulators in LED drivers

For the measurement of the curves in this figure the output power has been held constant while the value of C_{OUT} has been varied from 220 μF to 1100 μF in five steps. It is clearly visible that the peak-to-peak output ripple voltage of the first stage is proportional to C_{OUT}^{-1} . For each ripple value the proper RD has been selected in order to adjust the average drain voltage to the value mentioned earlier: $V_{HR} \geq \frac{V_{RIP,PP}}{2} + R_{DS(on)} \cdot I_{LED} + 400\text{mV}$.

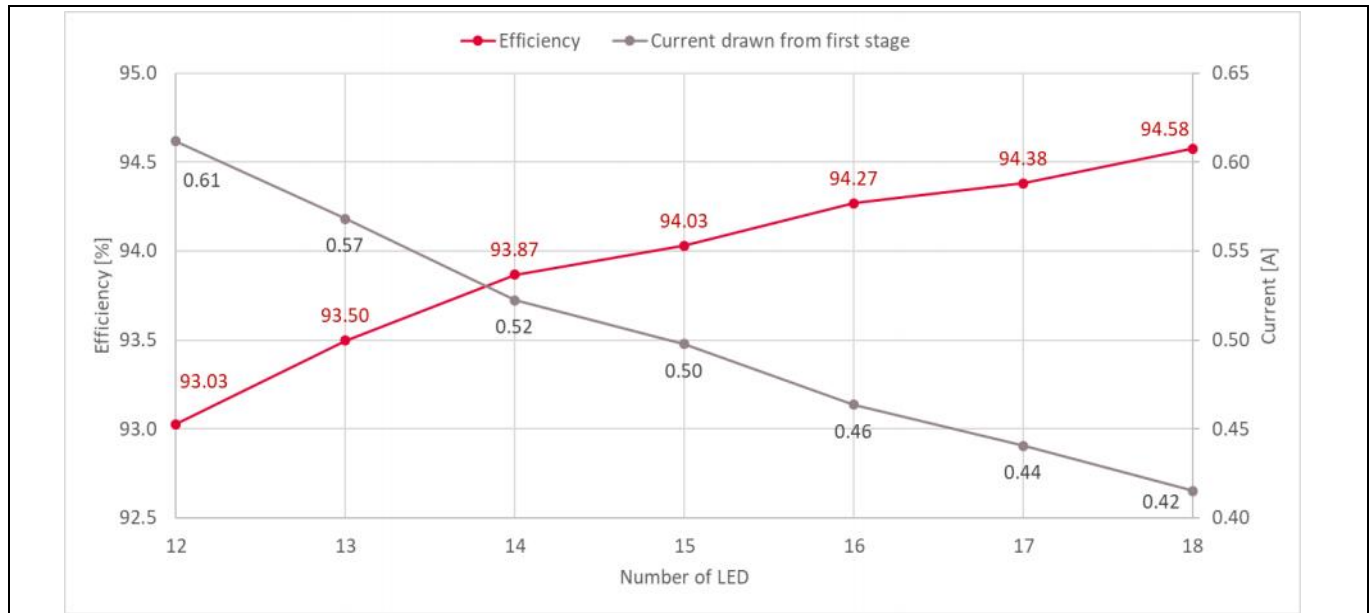


Figure 5 Efficiency vs. LED voltage for a constant output power of 20 W

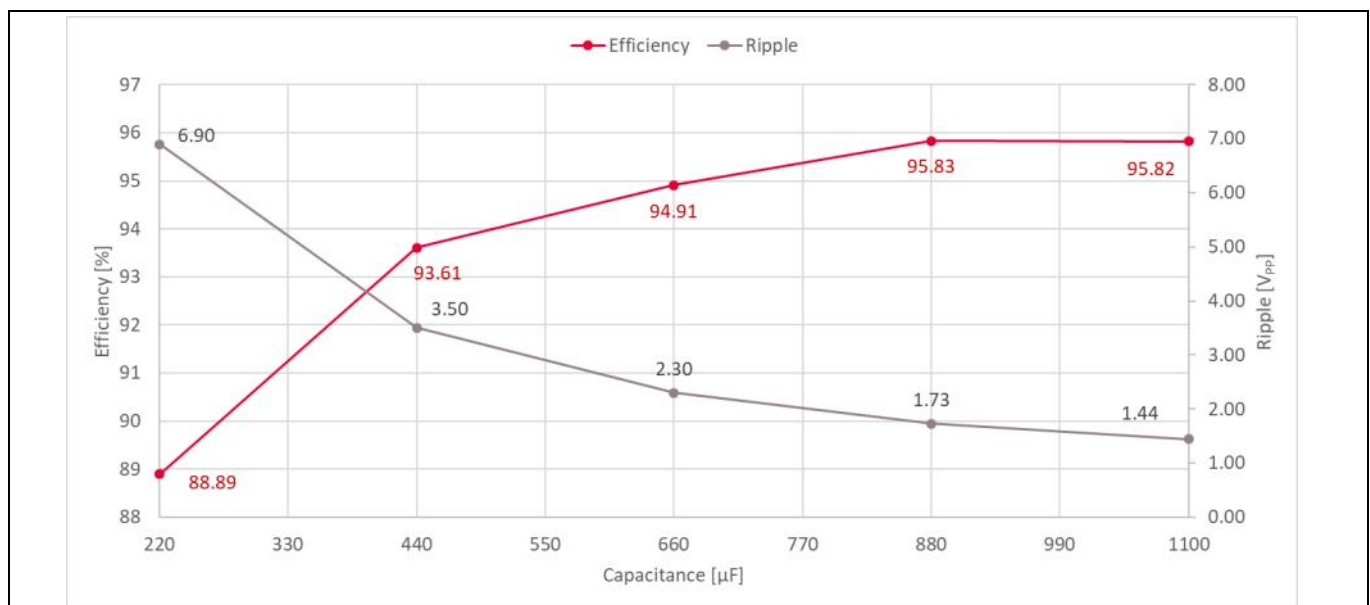


Figure 6 Efficiency vs. output capacitance C_{OUT} of first stage for a constant output power of 20 W

With such reduced ripple voltages, efficiencies above 95 percent are achieved.

Consequently, it is recommended to design the high power factor flyback stage to have a small output ripple amplitude.

Finally, combining both loss reduction methods, i.e. the highest possible LED voltage and a small ripple of around 2 V_{PP} , leads to an efficiency of the tunable white stage of 97 percent.

Linear constant current regulators in LED drivers

1.4 Setting V_{HR} for best efficiency

Key for high efficiency is a proper setting of V_{HR} .

The output voltage of a high power factor flyback is a DC voltage V_{DD} with a superimposed AC ripple of $\sin(x)^2$ shape and an amplitude V_{RIP} (Figure 7, right-hand side).

It is easy to see that the minimum of the output voltage of the first stage, which is $V_{DD} - V_{RIP}$, must be higher than the sum of the voltage drops of LED string, regulator and sensing voltage:

$$V_{DD} - V_{RIP} \geq n \cdot V_{LED} + I_{LED} \cdot R_{DS(on)} + V_{SENSE} \quad \text{Eq. 5}$$

The minimum possible voltage across the regulator is $V_{HR,MIN} = I_{LED} \cdot R_{DS(on)} + V_{SENSE}$. With ripple it varies periodically from $V_{HR,MIN}$ to $V_{HR,MIN} + V_{RIP,PP}$ with $V_{RIP,PP} = 2 \cdot V_{RIP}$.

Consequently, the optimum average drain voltage, which is regulated by the AHC, has the value:

$$V_{HR} = V_{RIP} + I_{LED} \cdot R_{DS(on)} + V_{SENSE} \quad \text{Eq. 6}$$

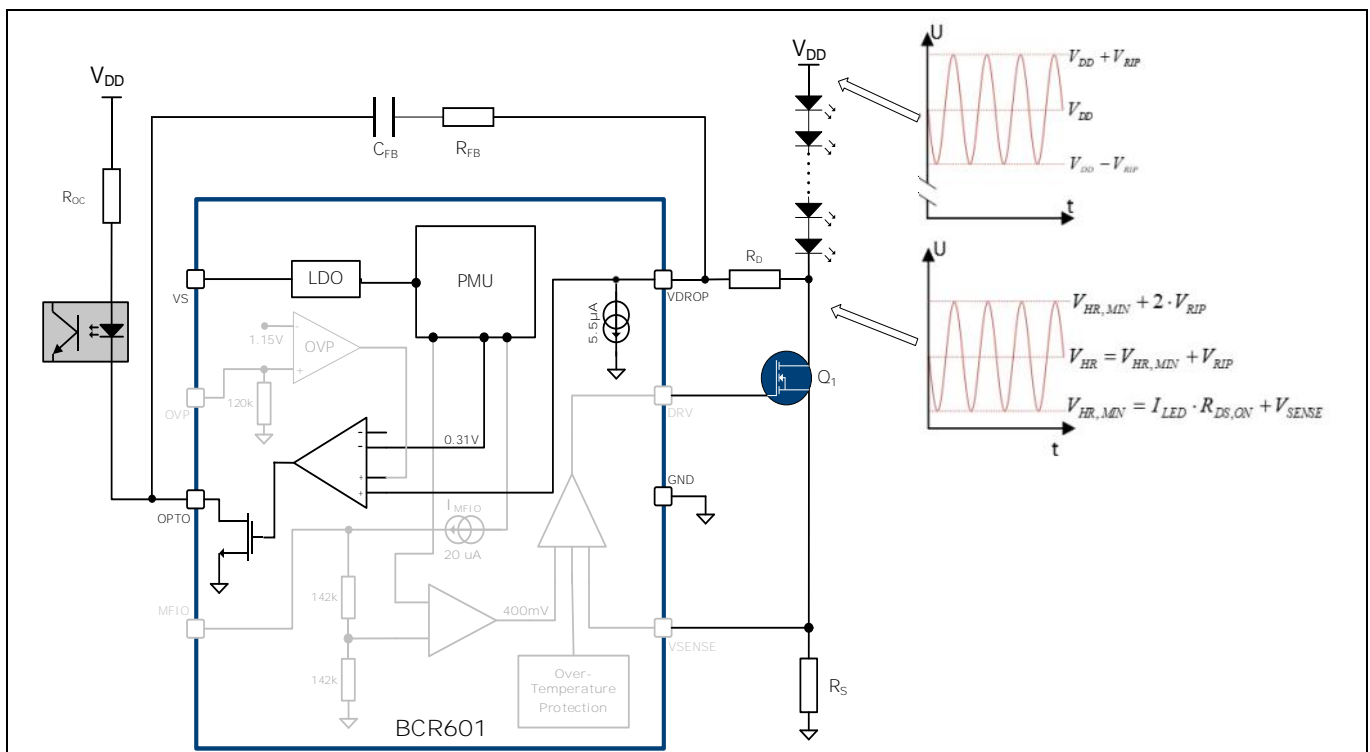


Figure 7 Block diagram of BCR601, AHC circuit emphasized

In steady-state the headroom voltage according to Eq. 6 is equal to the drop across R_D plus the setpoint of typically 0.31 V. **BCR601** has an internal current sink of typically 5.5 μA connected to pin VDROP, and with that the value of R_D is determined.

$$R_D = \frac{V_{RIP} + I_{LED} \cdot R_{DS(on)} + V_{SENSE} - 0.31 V}{5.5 \mu A} \quad \text{Eq. 7}$$

Linear constant current regulators in LED drivers

Because of the tolerance of ± 25 percent of the internal current sink it is recommended to add a resistor from VDROP to GND (R_{SINK} in **Figure 8**) that draws current that is much higher than the internal one, reducing the overall tolerance of V_{HR} . Working without R_{SINK} is recommended only for low-current LEDs that start to glow at very low current levels, and where at the same time the power dissipation is low.

In steady-state R_{SINK} draws an additional current of $0.31 \text{ V}/R_{SINK}$. In the final implementation of the multichannel driver (**Figure 17**) the value of R_{SINK} (R10) is $10 \text{ k}\Omega$ and the additional current is $31 \mu\text{A}$. The accuracy of the current through R_D then becomes ± 5 percent, which is sufficient for all purposes. Higher currents and accuracies are possible, but since the overall impedance level is lowered, C_{FB} may become too large.

Finally, Eq. 14 gives the expression for R_D if R_{SINK} is present.

$$R_D = \frac{(V_{RIP} + I_{LED} \cdot R_{DS(on)} - 0.09 \text{ V}) \cdot R_{SINK}}{5.5 \mu\text{A} \cdot R_{SINK} + 0.31 \text{ V}} \quad \text{Eq. 8}$$

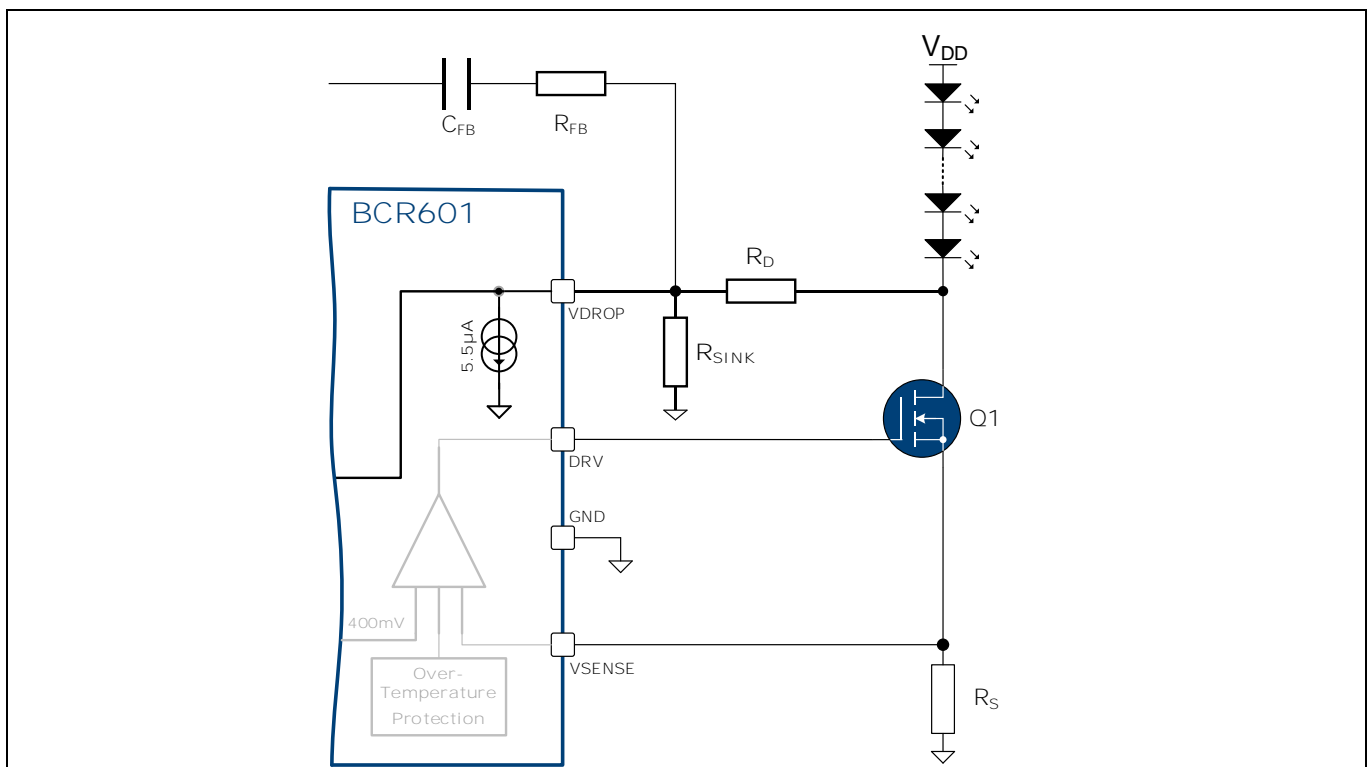


Figure 8 Improved AHC with reduced tolerance of V_{HR} due to R_{SINK}

The proper values for C_{FB} , R_{FB} and R_{OC} depend on the gain of the first stage and need to be chosen so that the total loop gain has a crossover frequency of about 10 Hz.

1.5 MOSFET and BJT selection

A very important parameter for the MOSFET selection is the gate threshold voltage $V_{GS(th)}$. Since the worst-case drive voltage of **BCR601** and **BCR602** is 4.5 V, the selected MOSFET must be fully on at this level. Consequently, a logic-level MOSFET is the best solution. Apart from that, MOSFET selection is very straightforward. Obviously the breakdown voltage $V_{DS(max)}$ of the MOSFET must be higher than the maximum input voltage V_{DD} , maximum drain current $I_{DS(max)}$ higher than targeted LED current, and power dissipation $P_{D(max)}$ at maximum ambient temperature higher than the one determined with Eq. 4. If the calculated loss is too high, a MOSFET with lower $R_{DS(on)}$ may help, or reducing ripple voltage on V_{DD} .

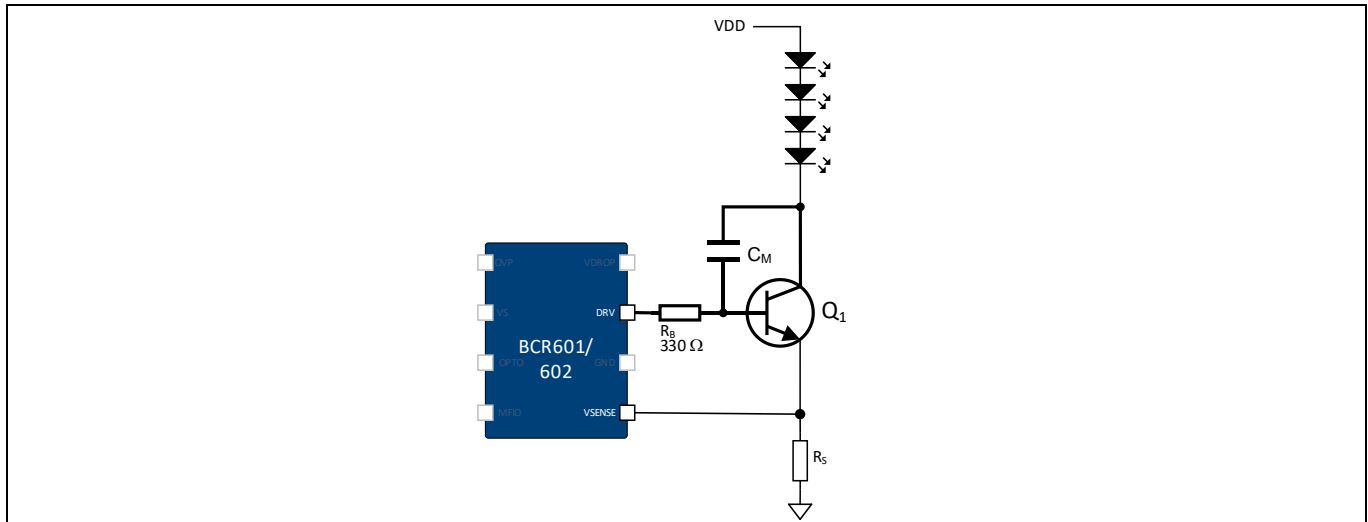


Figure 9 Using a BJT as an active element with BCR601/602

With a guaranteed drive current of 10 mA, **BCR601** and **BCR602** are able to drive a BJT as well. The same selection criteria apply as for the MOSFET. In addition, it is important that the current gain h_{FE} is as high as possible in order to drive the BJT into saturation at targeted current. Darlington BJTs can also be used. But these have the disadvantage of higher saturation voltage $V_{CE(sat)}$ and therefore don't support highest efficiency. High- h_{FE} BJTs can cause high-frequency oscillations of the current loop that need to be suppressed by a Miller capacitor C_M connected between collector and base. The correct value must be found by experimentation. Typically, a few tens of nF do the job. BJTs are a low-cost alternative for moderate current levels of a few hundred mA.

1.6 Overvoltage protection

BCR601 has a built-in overvoltage protection (OVP) stage. As soon as the supply voltage reaches the OVP level $V_{DD,OVP}$, which means the OVP pin reaches the internal setpoint of typically 1.15 V, the OVP stage takes control of the current through the optocoupler and in turn regulates V_{DD} to that level.

The actual OVP level is determined with the divider ratio of R_{OVP1} and R_{OVP2} . **BCR601** contains an internal resistor of typically 120 kΩ from the OVP pin to GND, i.e. in parallel to R_{OVP2} , influencing the divider ratio.

The design procedure is such that a value of R_{OVP2} in the range of a few kΩ is chosen and R_{OVP1} is calculated with:

$$R_{OV1} = \frac{V_{DD,OVP} - 1.15 \text{ V}}{1.15 \text{ V}} \cdot \frac{R_{OV2} \cdot 120 \text{ k}\Omega}{R_{OV2} + 120 \text{ k}\Omega} \quad \text{Eq. 9}$$

Since the setpoint has a certain tolerance, the minimum and maximum possible values must be considered as well. An Excel-based calculation sheet is available [1] that allows a quick calculation of R_{OV1} based on standard resistor series as well as the resulting minimum, typical and maximum overvoltage levels.

1.7 Overtemperature and hot-plug protection

Besides the OVP **BCR601** and **BCR602** provide overtemperature and hot-plug protection as well.

For the overtemperature protection (OTP) an IC internal sensor is used, and no further configuration steps are required. When a junction temperature T_J of 140°C is detected, I_{LED} is reduced to one-third of its nominal value. Normal operation is resumed when T_J has lowered to 120°C. Be aware that this protection mainly protects the chip, and is not able to safely protect the MOSFET or even the whole application from overtemperature.

Linear constant current regulators in LED drivers

When an LED string is connected to an active driver, for example when replacing defective LED modules, high current spikes with the potential for LED destruction might occur. Hot-plug protection reduces these spikes to an absolute minimum. This protection is activated whenever the voltage at pin VSENSE is below 8 mV. During this time I_{LED} is limited to 4 percent of its nominal value by reduction of the VSENSE reference voltage from 400 mV to 16 mV.

If the LED string fails with high resistance or is removed during operation, the resulting voltage at VDROPP will drop to zero. Consequently, the feedback loop will increase V_{DD} until the OVP-level $V_{DD,OV}$ is reached and then reduce power delivery from the primary side to an absolute minimum, in order to keep this level. Reconnection of an LED string will lead to a VSENSE level of 16 mV, and hot-plug protection is disabled with a delay of 2 μ s. After deactivation of the protection I_{LED} will go back to its nominal value while V_{DD} is still high. Since the feedback loop is slow and in a state of minimum power delivered from the primary side, the bulk capacitor will be discharged by I_{LED} down to the forward voltage V_{LED} of the connected LED. This leads to an additional power dissipation $P_{Q1,Hot-Plug}$ with an average power of:

$$P_{Q1,Hot-Plug} = \frac{1}{2} \cdot (V_{DD,OV} - V_{LED}) \cdot I_{LED} \quad \text{Eq. 10}$$

Duration $\Delta t_{Hot-Plug}$ of the additional dissipation is:

$$\Delta t_{Hot-Plug} = \frac{(V_{DD,OV} - V_{LED}) \cdot C_{OUT}}{I_{LED}} \quad \text{Eq. 11}$$

And typically in the range of some 10 ms. The drain voltage V_{HR} of Q_1 drops linearly from $V_{DD,OV} - V_{LED}$ to 0 V within that time. It must be ensured that the point ($V_{DD,OV} - V_{LED}$, I_{LED}) is below the 1 ms limit of the SOA of Q_1 and ($1/3 [V_{DD,OV} - V_{LED}]$, I_{LED}) below the 10 ms line.

After discharge of C_{OUT} the actual LED current drops to very small values until the feedback loop has settled and normal operation is resumed.

1.8 Output short-circuit protection

If the driver is integrated into a luminaire where LED modules can't be replaced, the likelihood of output short is extremely small. This is not the case when LED modules can be replaced, or in case of a separate driver.

Short-circuit seems not to be an issue, because AHC will reduce output voltage of the first stage as much as needed to keep the voltage across the MOSFET at its setpoint. At the same time the current regulation loop limits the output current to its nominal value.

An issue occurs when the first stage has a minimum output voltage that is higher than the AHC setpoint, which is typically a few volts. Depending on the behavior of the first stage, the corresponding regulator MOSFET might be overloaded. This is because its drain voltage immediately steps to the momentary V_{OUT} while it is still conducting the nominal LED current.

Overload under short-circuit conditions can be prevented by the additional network consisting of R_1 , ZD_1 , R_{17} and R_{18} in [Figure 10](#). If a short-circuit occurs, the drain voltage of the corresponding MOSFET will jump immediately to V_{DD} . ZD_1 starts to conduct then, and a current flow through R_1 , R_{18} and R_s starts, raising the voltage at the VS pin above 400 mV. The current regulator, which is very fast, immediately shuts down the MOSFET because the increased voltage level at V_s is interpreted as overcurrent. R_{17} limits the current into the VS pin in case of spikes exceeding the maximum voltage of that pin.

Linear constant current regulators in LED drivers

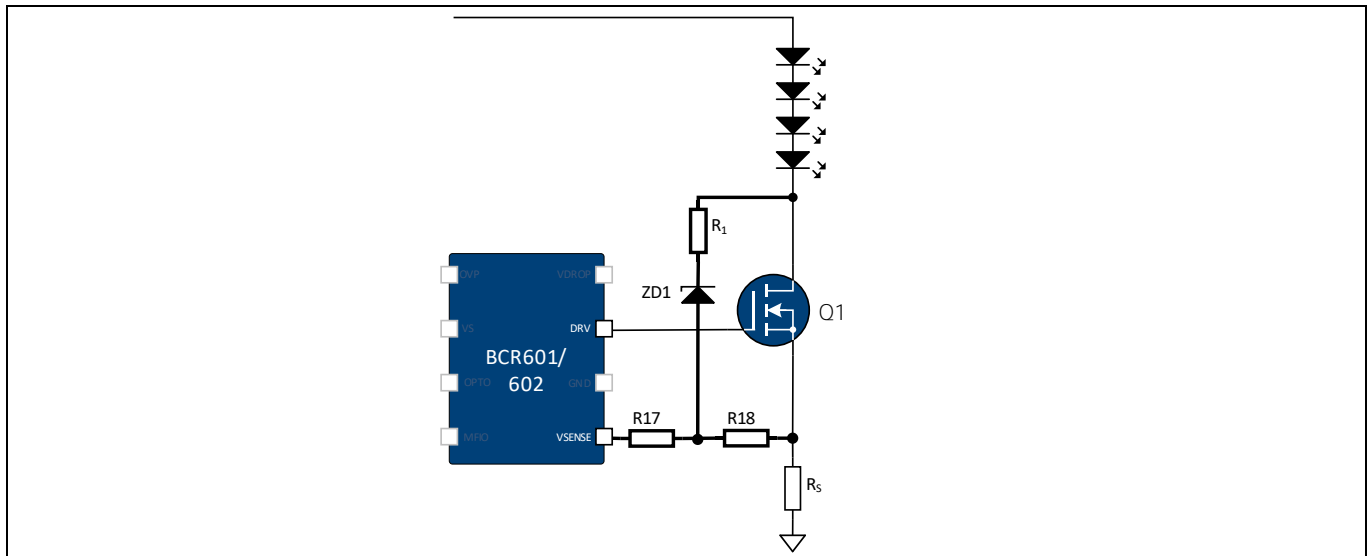


Figure 10 Adding output short-circuit protection to BCR601/602

The loop regulating the headroom voltage will at the same time recognize a far too high V_{HR} and reduce V_{DD} in order to bring V_{HR} back to its setpoint. Since this regulation loop is slow, it will typically take several hundred ms to bring V_{DD} down to the minimum output voltage of the primary side. This is not an issue, as the MOSFET is off. Of course $V_{DS(max)}$ of the MOSFET needs to be higher than the maximum possible V_{DD} .

The breakdown voltage of ZD1 must be lower than $V_{DD,min}$ in order to keep the MOSFET off under all conditions but higher than the peak of the regulated V_{HR} in order to prevent mis-triggering.

R1 and R18 limit the current and power dissipation of ZD1 when V_{DD} is at its maximum when a short-circuit occurs. The value of R_5 is small compared to R_1 and R_{18} and can be ignored here.

$$R_1 + R_{18} \geq \frac{V_{OUT,max} - V_{ZD1,BR}}{I_{ZD1,max}} = \frac{V_{BR,ZD1} \cdot (V_{OUT,max} - V_{ZD1,BR})}{P_{ZD1,max}} \quad \text{Eq. 12}$$

The maximum power dissipation of the two resistors also needs to be considered. Typically, resistors are not sensitive to short overload, and therefore it may be possible to use smaller resistors than the peak power dissipation suggests. This is because V_{DD} drops to $V_{DD,MIN}$ after the short-circuit has occurred, and P_{R1+R18} will drop accordingly. If V_{DD} drops fast enough, small SMD resistors are fine for R_1 and R_{18} . But this must be verified with the final hardware.

The divider ratio $R_{18}/(R_1+R_{18})$ must be selected so that the maximum voltage of 3.6 V at the VS pin is not exceeded.

$$R_{18} \leq \frac{3.6 \text{ V} \cdot (R_1 + R_{18})}{V_{OUT,max} - V_{ZD1,BR}} \quad \text{Eq. 13}$$

Consider the [XDPL8219](#) reference design as an example for the primary side. The output voltage range is from 32 V to 54 V typical. In this case for ZD1 a breakdown voltage $V_{ZD1,BR}$ of 27 V is a very reasonable value, providing 5 V margin to $V_{DD,min}$. Entering these data into Eq. 12 results in a minimum value of 1458 Ω for the sum of R_1 and R_{18} . The maximum value of R_{18} is determined with Eq. 13 to be 200 Ω . Finally, R_{18} is chosen to have 180 Ω and R_1 has a value of 1.3 k Ω .

Linear constant current regulators in LED drivers

With these values the Zener diode is well protected against electrical overstress. The peak power dissipation in R1 and R18 will be $(27\text{ V})^2/1480\ \Omega$, which equates to a dissipation of 0.49 W. The majority of the power will be dissipated in R1 while the loss in R18 can be ignored.

If the above dissipation is not acceptable, increase the resistance of $R_1 + R_{18}$ proportionally. Multiply both values by five, for example, and the dissipation will drop by the same factor. This also reduces the current through the Zener diode. Do not increase the value of the resistors too much, say by a factor of 10, because the current levels might become too small. The breakdown voltage of a Zener diode is not well defined at low current levels. At the same time the system will become more sensitive against leakage currents due to the higher resistance levels.

2 Multichannel LED driver with BCR601 plus BCR602

The full beauty of the linear regulator approach is that it can be easily extended to more than one output channel. Additional channels can be implemented by adding **BCR602** stages as shown in **Figure 11**.

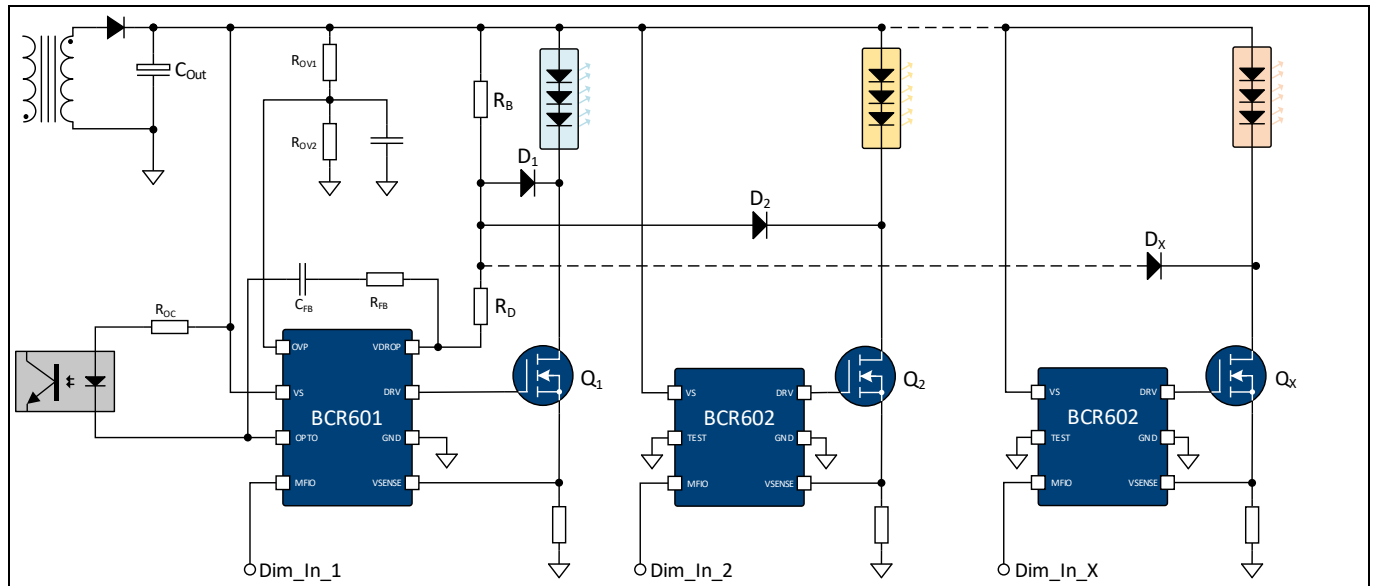


Figure 11 Linear regulator approach

BCR602 is a device very similar to **BCR601**, but without AHC and therefore in a smaller SOT-23-6 package. Consequently, the headroom control for all channels needs to be accomplished by **BCR601**. Apart from that all other parameters are identical to **BCR601**: current regulator, dimming range and dimming characteristics, etc. are identical. Therefore Eq. 2 and Eq. 3 for example are valid for **BCR602** as well.

At first, it seems to be a good strategy to implement a network that always automatically uses the channel with the highest V_{HR} to regulate V_{DD} . More detailed analysis quickly shows that this approach doesn't work well. If the highest drain voltage is used to regulate the headroom, it may happen that the resulting V_{DD} is too low to supply the other stages. This can be understood with the help of **Figure 12**.

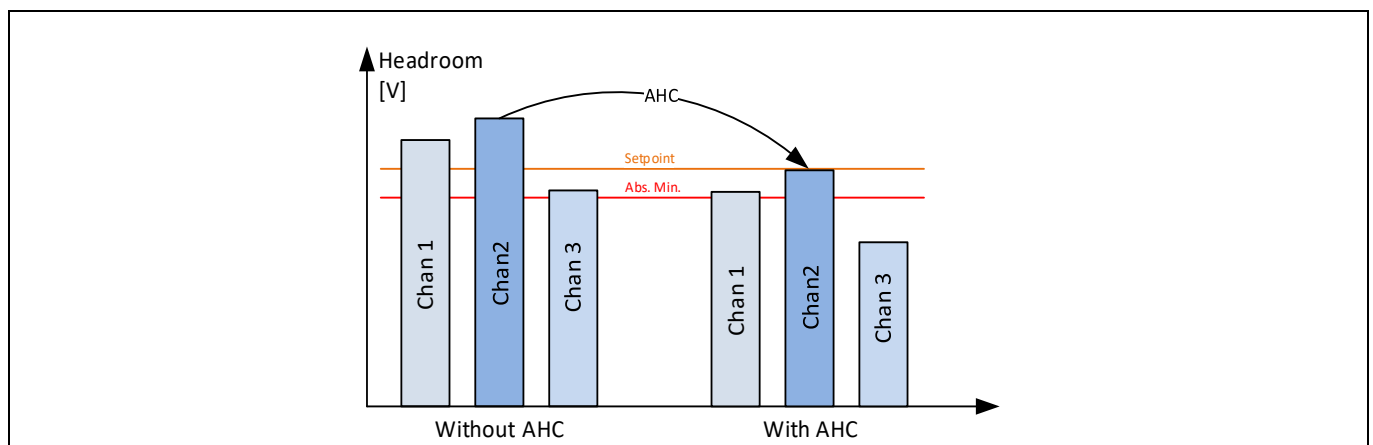


Figure 12 AHC regulating the channel with maximum headroom voltage

Multichannel LED driver with BCR601 plus BCR602

The left-hand side of **Figure 12** shows an arbitrary distribution of headroom voltages of a system with three output channels and without AHC. In this case channel 2 shall have the highest value. Now assume that AHC is turned on and regulates the highest of the three voltages to the desired setpoint. Because this is achieved by lowering V_{DD} by a proper amount, the headroom voltages of the other two channels would be reduced by exactly the same amount. In this example, the headroom voltage of channel 3 would fall below the absolute minimum that is necessary to allow current flow through the LEDs. In other words, channel 3 would be dark.

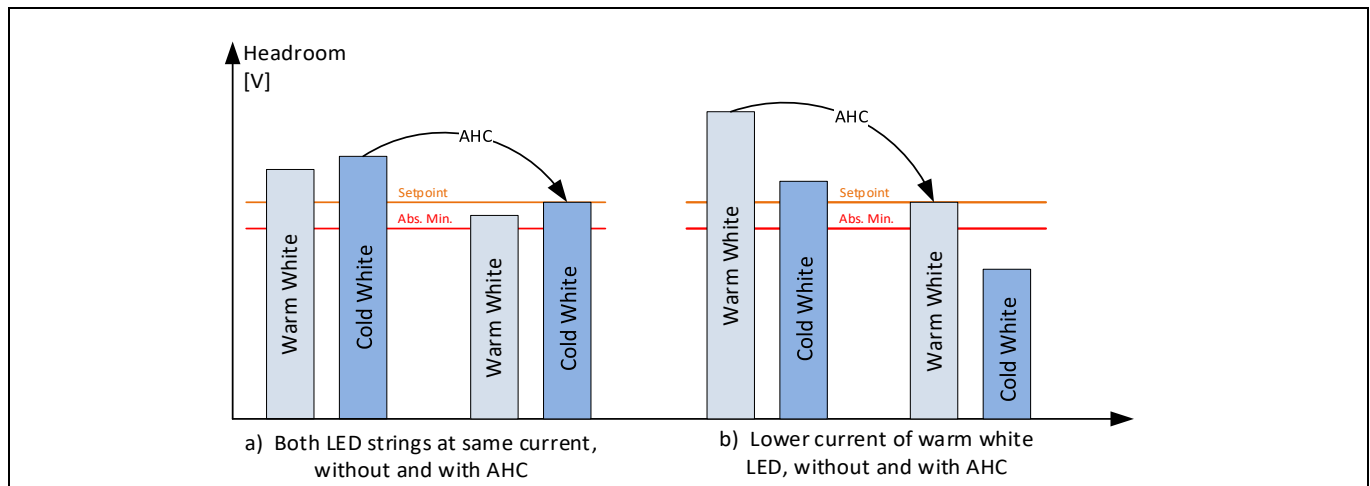


Figure 13 Hypothetical tunable white system with AHC regulating the maximum headroom voltage

For tunable white applications this seems at first to be a minor problem, at least as long as the warm white and cold white LEDs are from the same family. In that case the forward voltages of the two strings are close enough that a proper selection of the setpoint allows both channels to work as intended (**Figure 13 a**). This is the case when both LED strings are driven more or less with the same current. But in tunable white applications, the currents of both strings vary over a wide range. If for example a higher comparative color temperature (CCT) is desired, the current in the warm white string is reduced, while the one in the cold white string is increased. That means that the forward voltage of the warm white LED string is decreasing and the headroom of that channel is increasing by the same amount. In the cold white string exactly the opposite happens: increase of forward voltage, decrease of headroom. An AHC that always regulates the maximum would now use the warm white channel for regulation and reduce V_{SUP} so much that the cold white string would be dark (**Figure 13 b**).

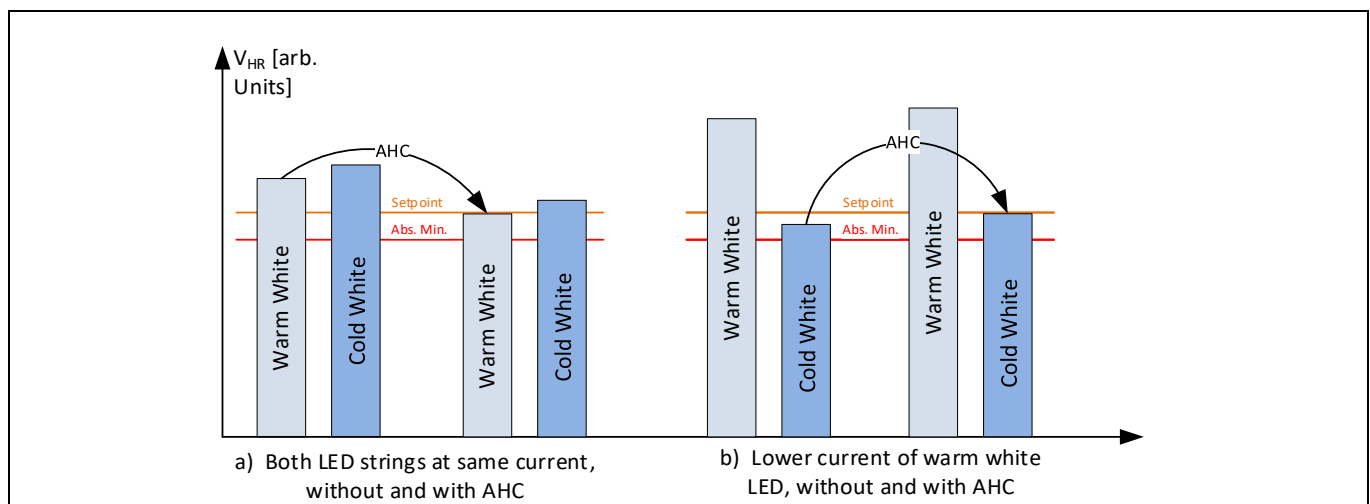


Figure 14 As in Figure 13, but AHC regulating the channel with minimum headroom voltage

Multichannel LED driver with BCR601 plus BCR602

It therefore follows that the best strategy for multichannel systems is to always use the channel with lowest V_{HR} for regulation, as shown in **Figure 14**.

With this strategy for regulation the other channel will have an increased headroom voltage. This is not a problem, because that is the channel with the lower current. Consequently, overall loss and efficiency are hardly affected by this approach.

The network that enables regulation of the V_{HR} of the channel having the lowest V_{HR} simply consists of one resistor (R_B in **Figure 11**) and a small signal diode per channel ($D_1 \dots D_X$). Since the resistor is connected to V_{SUP} it simulates a relatively high headroom voltage. By means of the diodes it gets clamped to the smallest V_{HR} of all channels plus one diode forward voltage.

Figure 15 shows the measurements of a tunable white driver that uses the above-mentioned regulation scheme. The inputs are controlled by two 180-degree shifted triangular signals leading to a periodically changing CCT. CH3 (magenta) and CH4 (light blue) are the actual LED currents, while M1 (red) and M2 (blue) show the low-pass filtered drain voltages of the corresponding regulator MOSFET (V_{HR}). The cut-off frequency of the low-pass is selected to be low enough (10 Hz) to remove the 100 Hz ripple so that the displayed curves are essentially the average drain voltages.

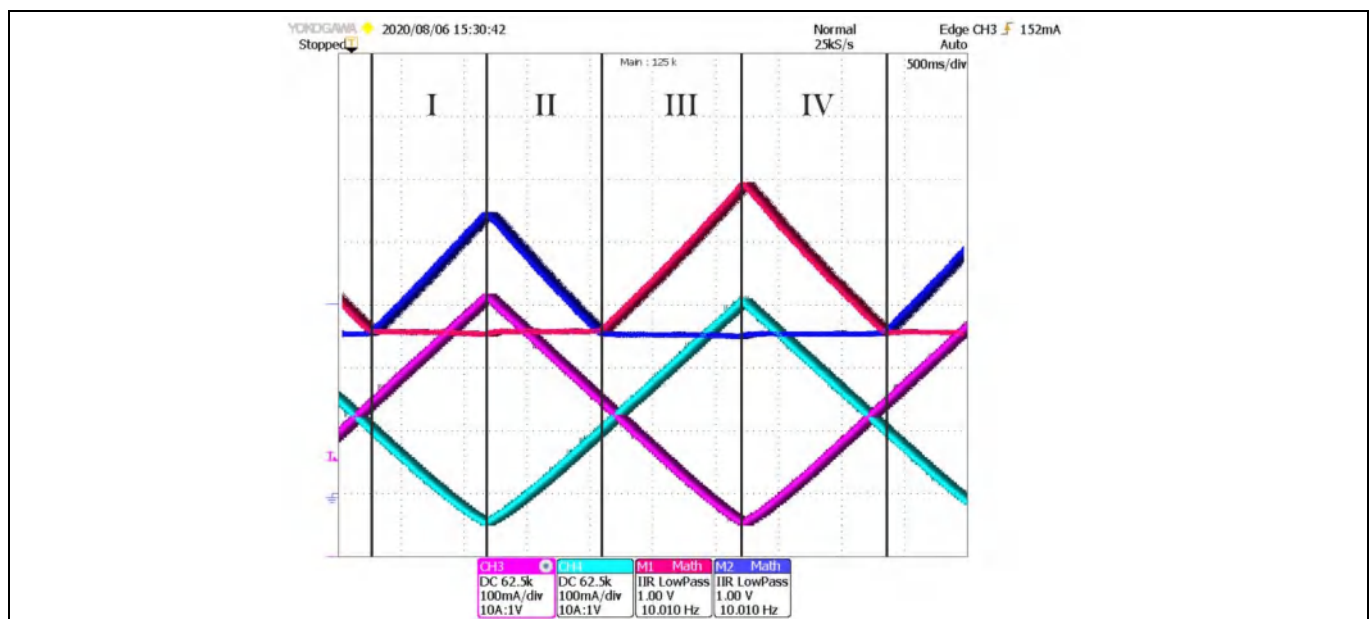


Figure 15 LED currents and average drain voltages of MOSFETs in a tunable white application.

CH3: I_{LED} of warm white LED, CH4: I_{LED} of cold white LED

M1: Average drain voltage of Q2, M2: Average drain voltage of Q22

In area I, the current through the warm white LED string is higher than the one through cold white, causing the voltage drop of warm white to be higher than the one of cold white. Higher drop across the LED would mean lower headroom voltage of the corresponding MOSFET. Consequently, this headroom voltage is regulated to a constant value determined by R_D and R_G (red, M1).

The voltage drop of the cold white LED goes down in the same way the current drops, leading to increased headroom voltage of the corresponding MOSFET (blue, M2).

In area II, warm white current is falling and cold white current is rising again. As the voltage drop of cold white is increasing the headroom voltage of the corresponding MOSFET Q_{22} is falling until it becomes smaller than the drop across Q_2 . Then, in area III and IV, the V_{HR} of Q_{22} is regulated while V_{HR} of Q_2 behaves in a similar way as V_{HR} of Q_{22} did in areas I and II.

Multichannel LED driver with BCR601 plus BCR602

An interesting point is that the takeover of the regulation from Q2 to Q22, i.e. the transition from II to III, doesn't occur at equal LED currents. This shows that the $V(I)$ characteristics of the warm white and cold white LEDs are not identical, although both are from the same family and are at the same temperature.

2.1 Setting V_{HR} in multichannel drivers

There is no big difference between the V_{HR} settings in single- and multichannel applications. Assuming that R_{SINK} is used for increased accuracy of the V_{HR} setting, the only difference is the additional forward voltage drop V_{FWD} of one the (identical) diodes $D_1 \dots D_X$ of [Figure 11](#). Adding this drop to Eq. 7 leads to Eq. 14 for R_D in multichannel drivers:

$$R_D = \frac{(V_{RIP} + I_{LED} \cdot R_{DS(on)} + V_{FWD} - 0.09 V) \cdot R_{SINK}}{5.5 \mu A \cdot R_{SINK} + 0.31 V} \quad \text{Eq. 14}$$

It is noteworthy that the forward drop of the diode enters the equation for R_D but doesn't increase drain voltage and losses. Finally, select R_B to provide a higher current than the one through R_D at minimum output voltage.

2.2 Dimming, overvoltage, short-circuit and hot-plug protection in multichannel drivers

Regarding dimming behavior, there is no difference between the single- and multichannel applications.

OVP is handled by [BCR601](#) for the whole application.

[BCR602](#) has identical current regulator and dimming circuits as [BCR601](#), and therefore behaves exactly the same. That means that the dimming behavior and performance is, within the accuracy of the components, the same.

Regarding short-circuit protection (SCP), there is a slight difference between single- and multichannel. Due to the diodes $D_1 \dots D_X$ [BCR601](#) is not able to detect the excessive V_{HR} caused by a short-circuit. Consequently, V_{DD} will essentially stay at the level it was before short-circuit. This is no issue, since there is no current flow in the respective MOSFET. But it is important to pay attention to the maximum allowed power dissipation of the Zener diode and its series resistor.

Hot-plug detection works in a similar way as in the single-channel driver. The actual protection is handled by the respective control IC. A major difference to the single-channel application is that the remaining channels are still working and drawing power from the primary side. At the same time the feedback loop increases V_{DD} due to loss of regulation signal. This leads to the situation that the remaining channels have to deal with input voltages up to V_{OVP} while potentially still at full current. This can cause breakdown of the active element if the hot-plug event takes too long.

Note: Consequently, the multichannel application doesn't support hot-plug of LEDs.

3 Tunable white reference design with BCR601 and BCR602

3.1 Board setup

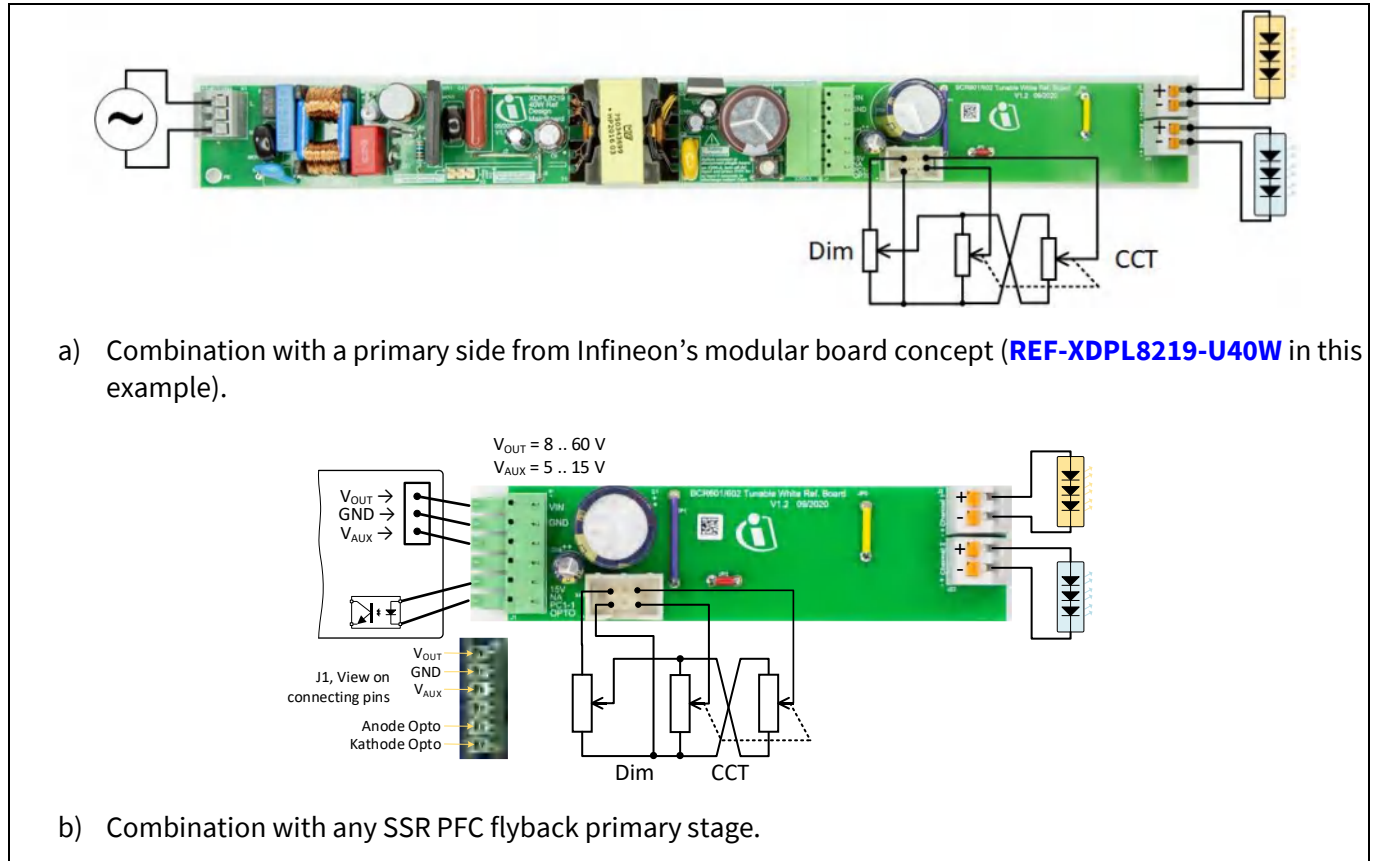


Figure 16 Board setup

3.2 Board specifications

Table 1 lists the key electrical specifications of the reference design. The board as delivered is ready to work with [REF-XDPL8219-U40W](#) as primary side, and the below-mentioned performance data are determined in combination with that board.

Table 1 Electrical specifications for tunable white reference design board

Parameters	Symbol	Min.	Typ.	Max.	Unit	Remarks
Input voltage range	V_{DD}	33 ⁽¹⁾		60	V	Input voltage is regulated by AHC
Targeted LED voltage	V_{LED}	33 ⁽¹⁾		53	V	SCP is for V_{DD} greater than 30 V
Oversvoltage protection	V_{OVP}	54		57.9	V	
Regulated headroom	V_{HR}		1.8		V	Average of drain voltage
LED current CH1	I_{LED}	15		470	mA	Depending on dimming and CCT
LED current CH2	I_{LED}	15		470	mA	Depending on dimming and CCT
Light flicker	$P_{st,lm}$			0.15	–	
Stroboscopic effect	SVM			0.1	–	
Efficiency	η			94.5	%	V_{DD} ripple = 3 V _{PP} , V_{LED} = 52 V, CH1 = CH2 = 235 mA

Table 2 System specifications for tunable white reference design with REF-XDPL8219-U40W

Parameters	Symbol	Min.	Typ.	Max.	Unit	Remarks
Input voltage range	V_{IN}	90		277	V _{RMS}	
Power factor	λ		0.99		–	110 V AC, full load
Input current THD	iTHD		6.5		%	
Efficiency	η		88.0		%	
Power factor	λ		0.99		–	230 V AC, full load
Input current THD	iTHD		8.2		%	
Efficiency	η		88.5		%	

Note: Dimensions: 100 mm x 26 mm x 26 mm

3.3 Schematics and circuit description

Figure 17 and **Figure 18** illustrate the schematics and layout of the board, respectively. Many parts of the schematic have already been described and can be recognized easily.

R12/R13 determine the maximum current of channel 1 and R32/R33 do the same for channel 2. With the given values of 1.5 Ω and 1.6 Ω the maximum LED current in each channel is 516 mA nominal. In the tunable white application, the sum of the currents of both channels will add up to this value.

The output SCP for each channel is implemented by means of D1, R1, R17, R18 and D21, R21, R37, R38 respectively.

R6 and R10 determine the headroom voltage of the MOSFET Q2 and Q22. With the given values the average V_{HR} is 1.8 V typical.

C7 and C27 generate a low-impedance supply for **BCR601** and **BCR602** respectively and reduce high-frequency noise that might be present on the supply voltage. Resistors R2 and R7 as well as R22 and R27 have a value of 0 Ω on the board as delivered. Use non-zero values if a reduction of power dissipation of the two ICs is desired.

¹ Minimum supply of BCR601/602 is 8 V but minimum regulated voltage of REF-XDPL8219-U40W is ~30 V and SCP is dimensioned for that.

Tunable white reference design with BCR601 and BCR602

If BJTs are used instead of MOSFETs and the maximum driver output current of 10 mA is needed, this may be the case. For details, refer to the datasheets of [BCR601](#) and [BCR602](#).

In case the primary side you intend to use delivers an auxiliary voltage in the range of 8 V to 15 V, you may use that to supply the control IC as well as the optocoupler circuit. To do this, you need to remove R20 and solder a suitable value for R14. If the auxiliary voltage is well below 15 V, a few tens of Ω are acceptable for R14. If the voltage is higher, a higher value for R14 is needed to prevent D15 from overloading. If you want to supply the control IC as well, you need to remove either R2 or R7 and solder a wire from the cathode of D15 to pin VS of U1. In the second channel you have to remove R22 or R22a and solder a 0 Ω resistor for R23 in this case. Supplying the IC and optocoupler from an auxiliary voltage further improves efficiency.

The voltage divider R3/R4 determines the OVP level, which is 57.7 V typically.

R20, D15 and C15 generate a stable and noise-free supply for the optocoupler circuit. The optocoupler itself is located on the board of the first stage. The values of R15 and R19 define the maximum optocoupler current and the selected values are optimized for operation with [REF-XDPL8219-U40W](#). For different primary sides these values need to be adjusted as well as potentially C2 and R8, which determine the frequency response of the feedback loop.

C5 and C25 reduce the susceptibility to oscillations in case of long LED connection cables and their inductance. Stability of current regulators in general is reduced with inductive load, and capacitance at the regulator output helps.

Finally, there is another regulator G41 that delivers a stable 3.3 V output from a 15 V input. This can be used as reference voltage for a passive tunable white controller or as the supply voltage for an active one.



Tunable white LED drivers with BCR601 and BCR602

How to design cost effective, highly efficient LED drivers of highest light quality



Tunable white reference design with BCR601 and BCR602

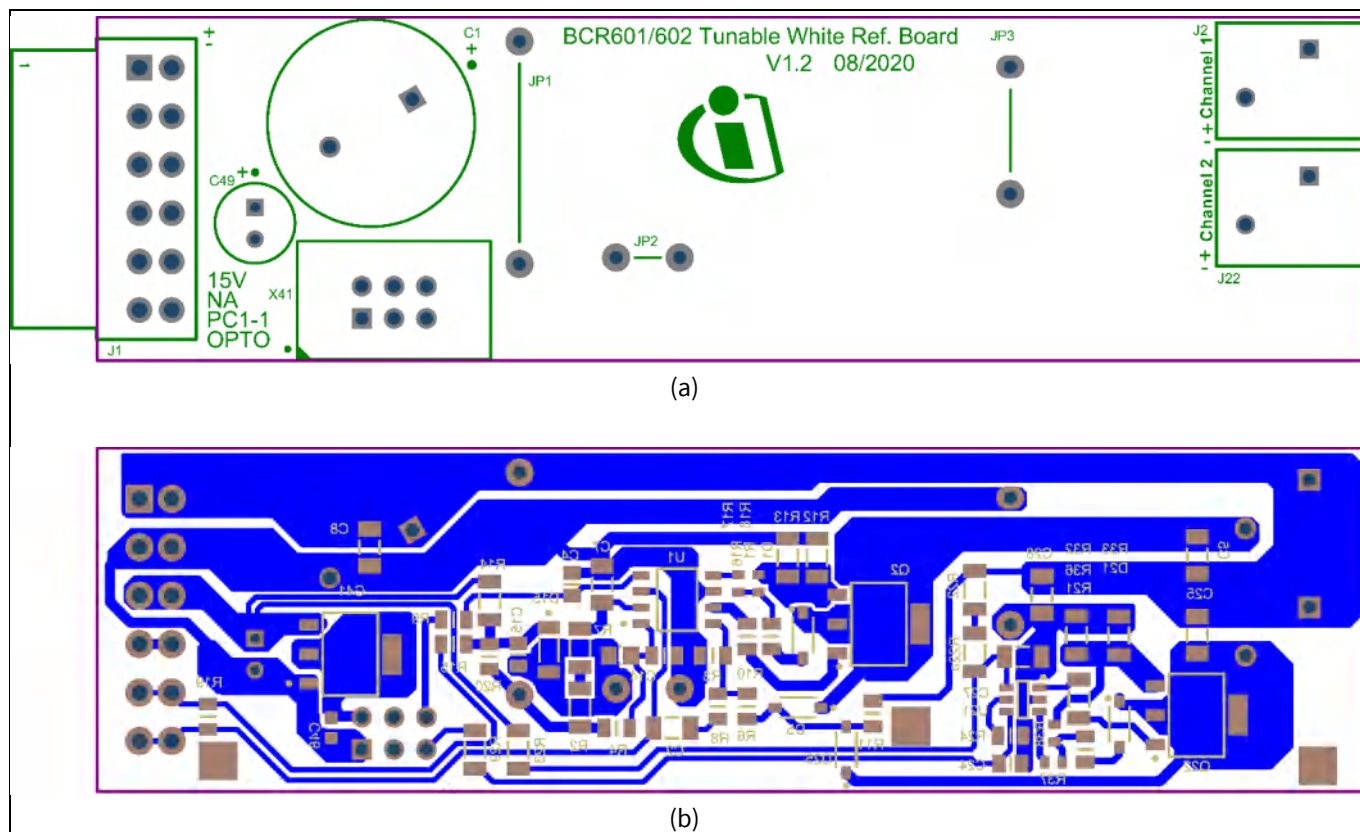


Figure 18 Board layout – (a) top side and (b) bottom side

3.4 Electrical performance

3.4.1 Light quality

This section describes those performance data of the BCR601-based regulator that are essentially independent of the primary stage used. In some case this is not totally true, because the performance of the first stage (e.g., output voltage ripple) will also influence **BCR601** performance.

3.5 LED current stability and regulator efficiency

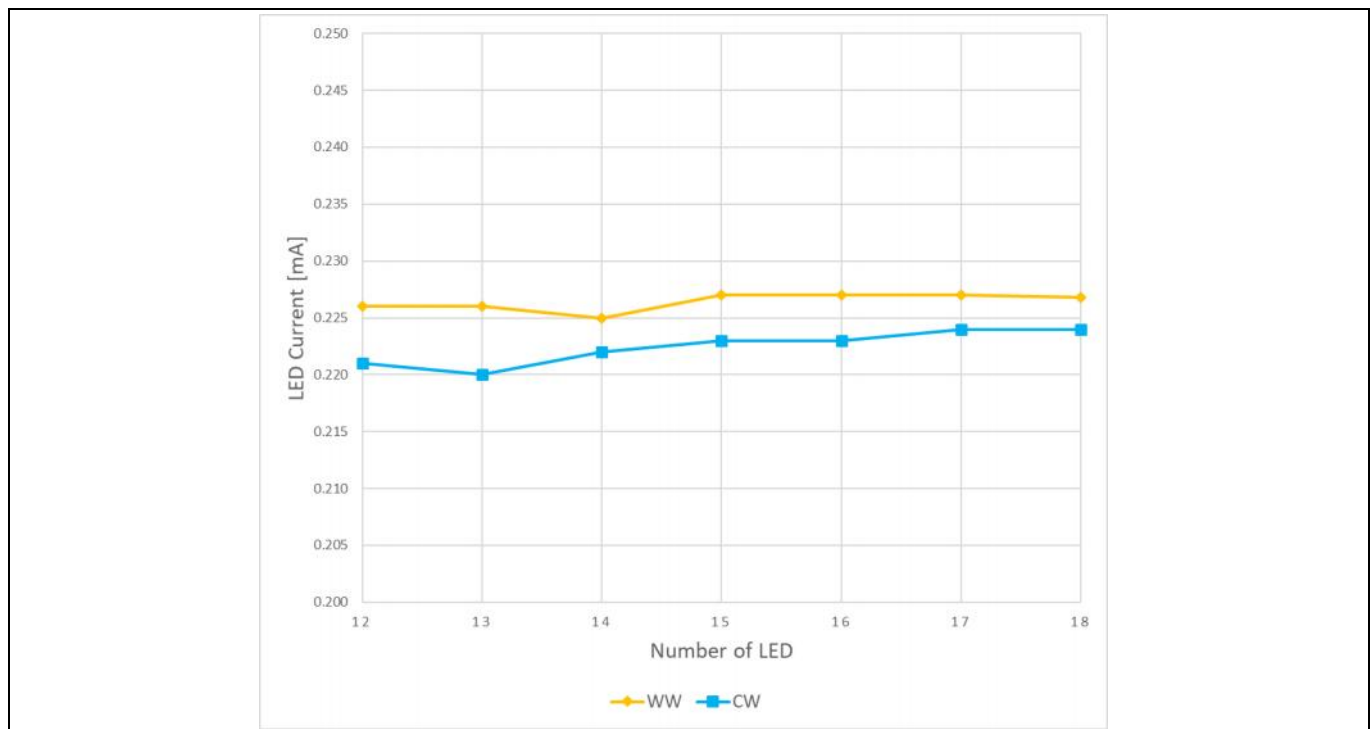


Figure 19 Undimmed LED current vs. number of connected LEDs

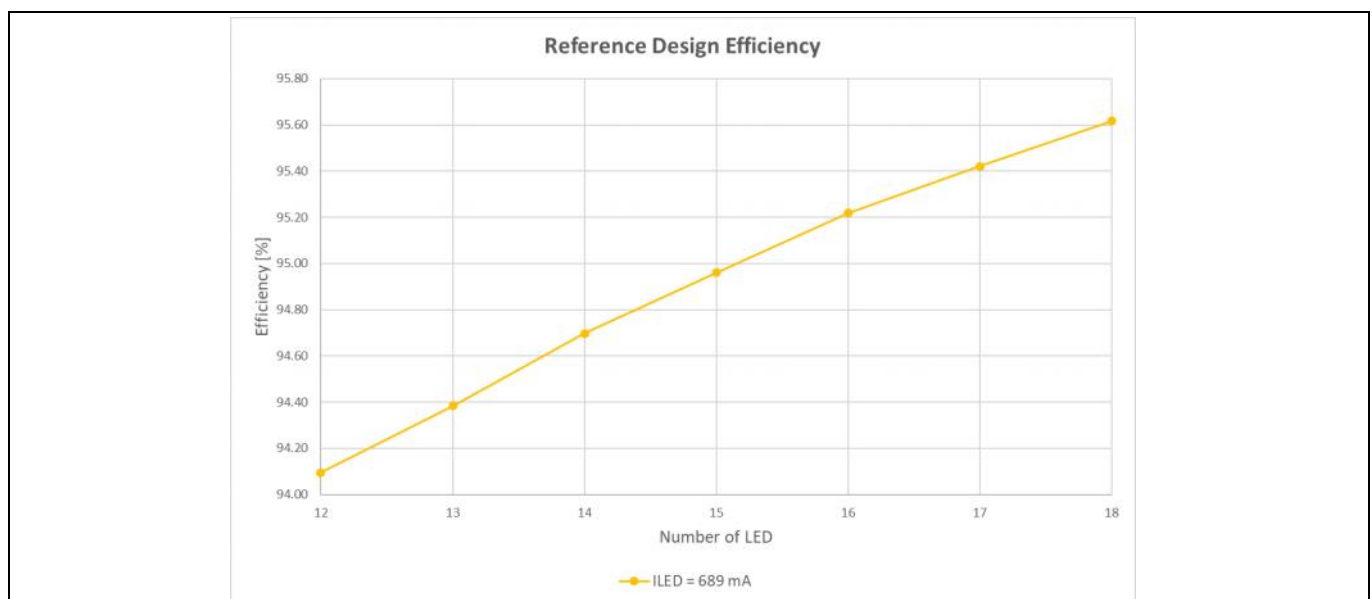


Figure 20 Efficiency of BCR601/602 reference design vs. number of LEDs ($V_{\text{Ripple}} = 2.5 V_{\text{pp}}$)

3.6 Light quality

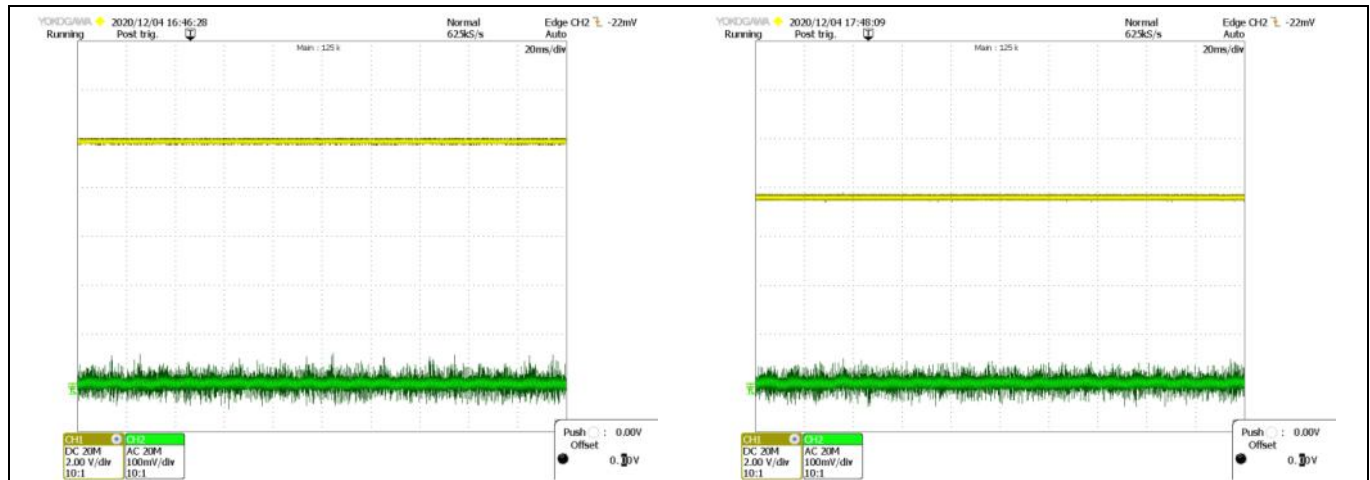


Figure 21 Output of light sensor. CH1 (yellow) is the DC signal, which is proportional to the light intensity. CH2 is AC content of the same signal at 20 times higher sensitivity. Left: $I_{LED} = 689$ mA, right: $I_{LED} = 515$ mA (dimmed to 75 percent).

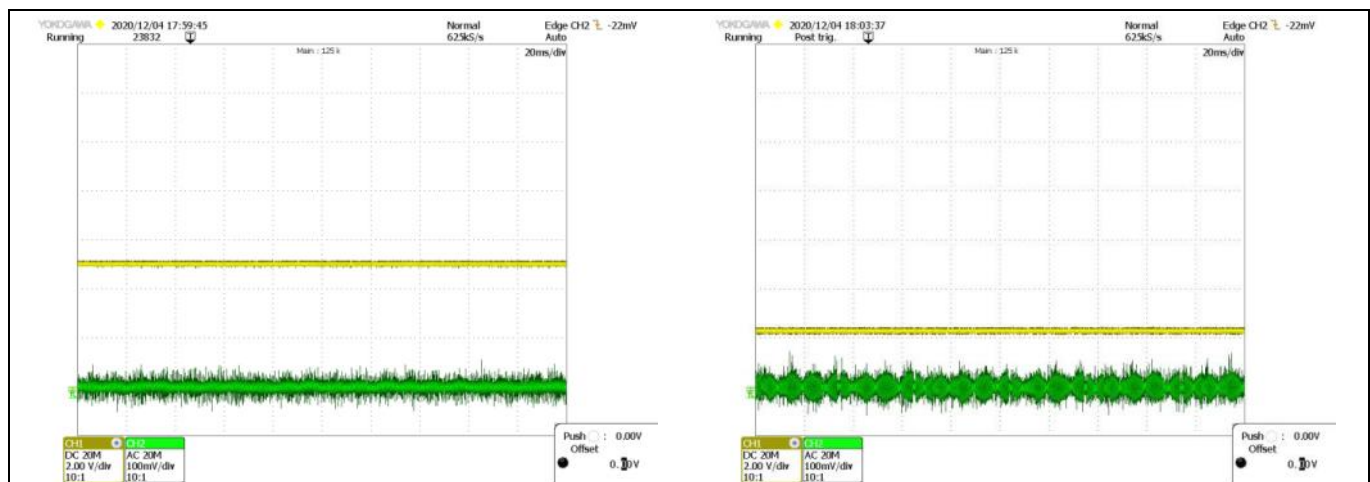


Figure 22 Same as Figure 21. Left: $I_{LED} = 320$ mA, right: $I_{LED} = 130$ mA.

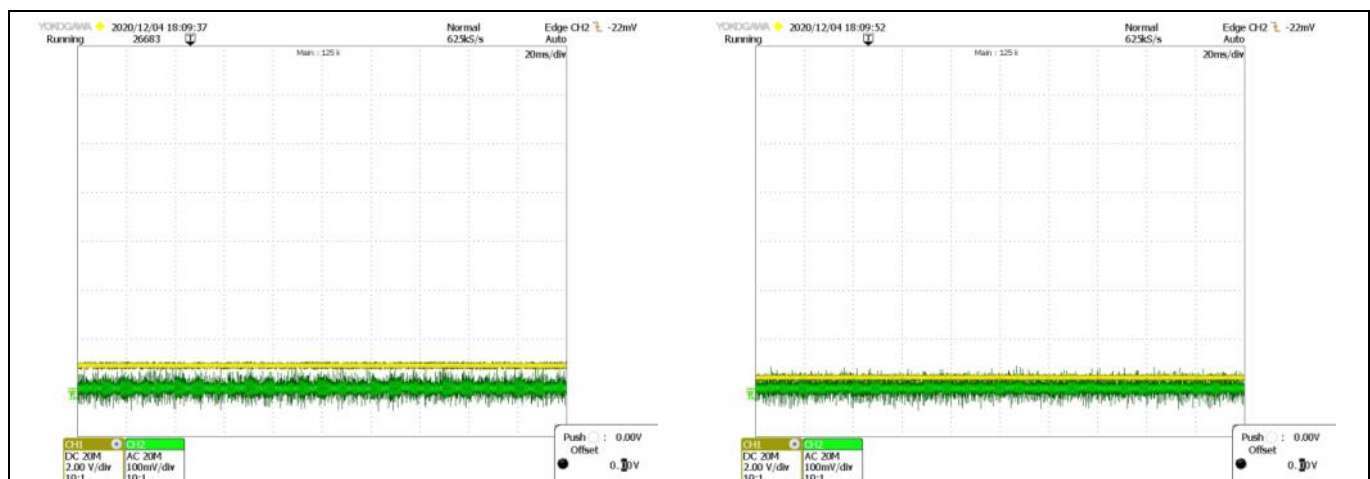


Figure 23 Same as Figure 21. Left: $I_{LED} = 56$ mA, right: $I_{LED} = 23$ mA.

Tunable white reference design with BCR601 and BCR602

Figure 19 and **Figure 20** give an impression of the excellent stability of the LED current and the high efficiency of the design.

Figure 21 to **Figure 23** show the output of a light sensor. The yellow trace is the sensor output, which is proportional to the light intensity, while the green trace is the AC content of the same signal at 20 times higher sensitivity. This AC component is a measure for the modulation of the light.

The output level of the light sensor at maximum I_{LED} is about 10 V while the noise level is less than 100 mV_{pp} (**Figure 21; left**). From that a modulation index of less than 0.1 percent can be determined. The AC content of the light sensors' signal doesn't vary with light output (**Figure 21; right**), which indicates that the AC signal is mainly determined by sensor noise. When the output current is reduced to 20 percent and below, the effect of primary-side burst mode becomes visible in the light sensor's signal (**Figure 22, right**).

The temporary lighting artifacts (TLA) have been determined with a BTS256-EF spectral light meter with flicker measurement function from Gigahertz-Optik. The results for different LED currents are summarized in **Table 3**.

		LED current					
		689 mA	515 mA	320 mA	130 mA	56 mA	23 mA
TLA measure	P_{st}^{LM}	0.0	0.0	0.0	0.042	0.066	0.122
	SVM	0.042	0.042	0.042	0.042	0.042	0.042
	Assist Mp	0.009	0.012	0.02	0.03	0.046	0.087

Table 3 Light quality vs. output current

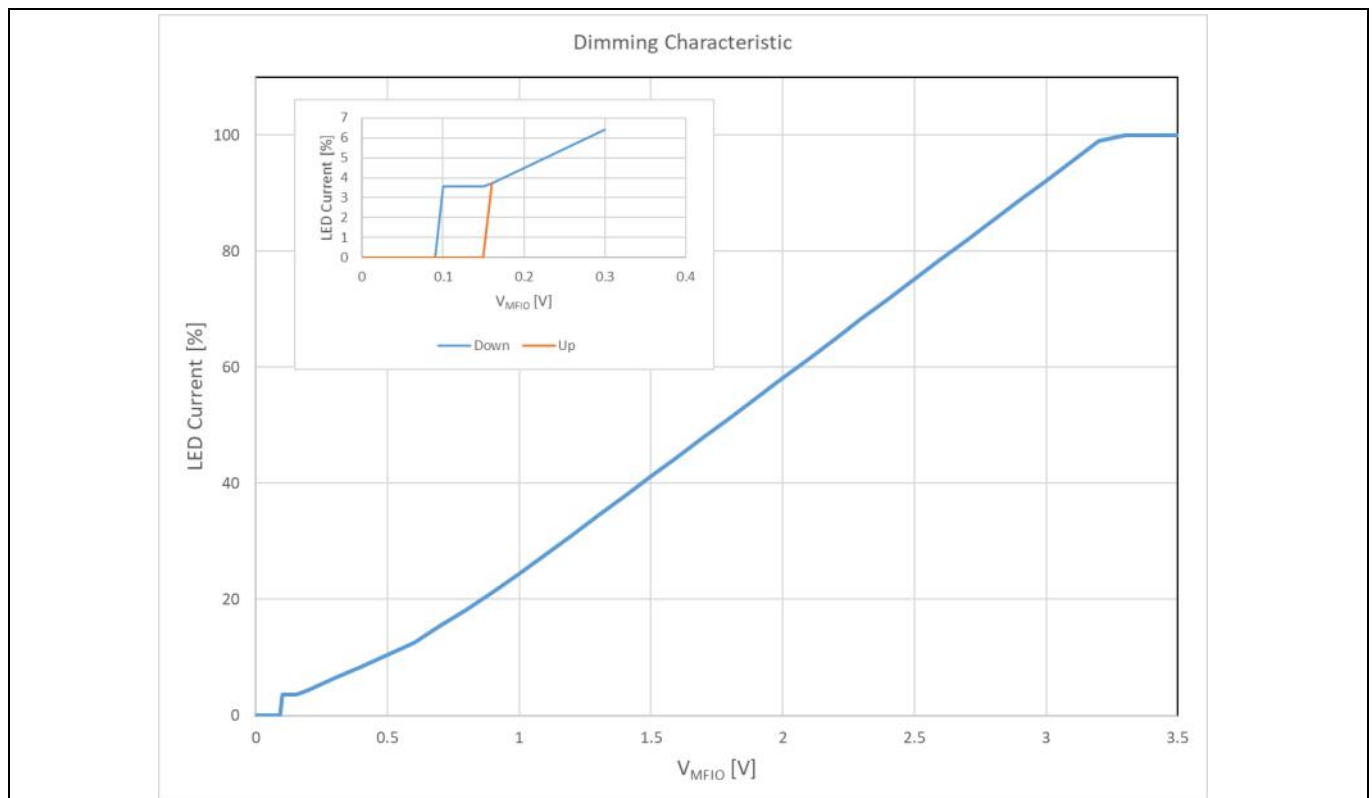


Figure 24 Dimming characteristic: relative LED current vs. V_{MFIO}

3.7 System performance

The board as delivered is ready to work with **REF-XDPL8219-U40W** as primary side, and the below-mentioned performance data are determined in combination with that board.

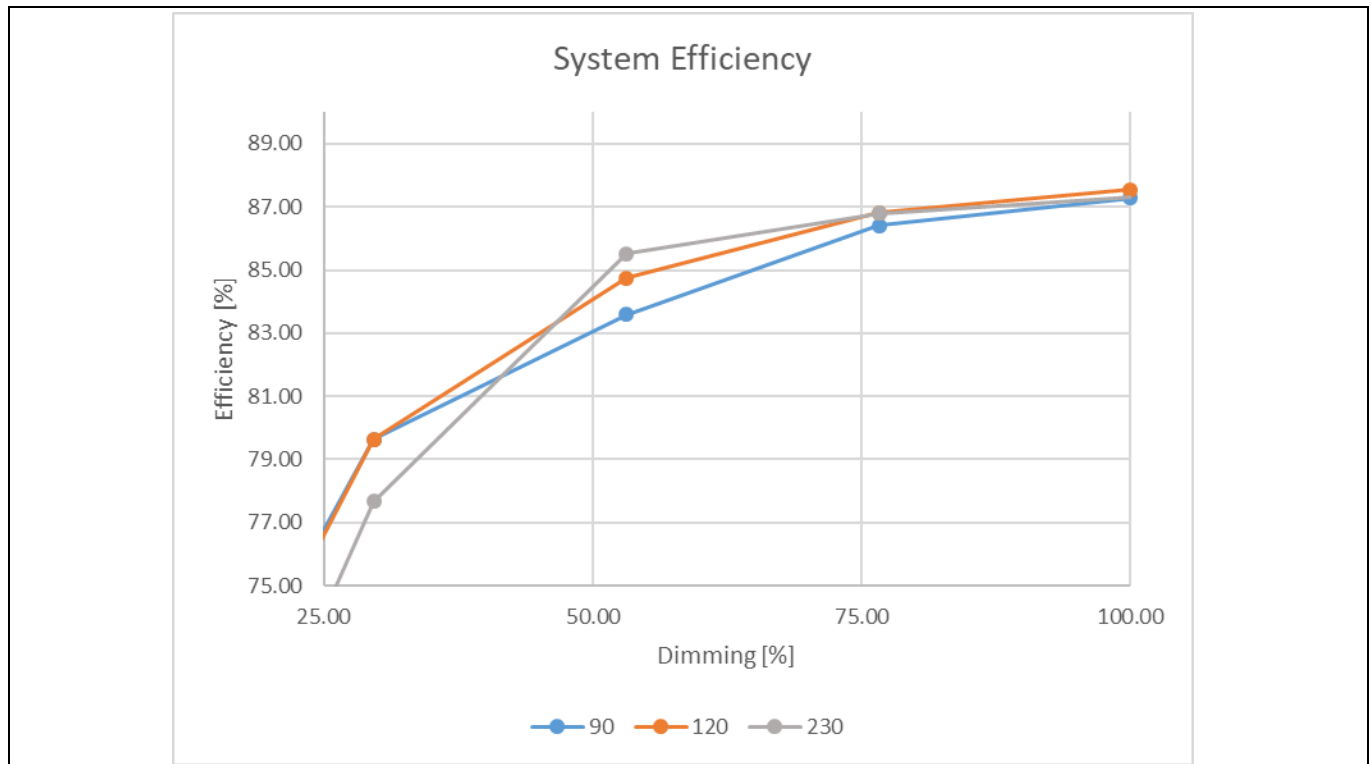


Figure 25 Total system efficiency of BCR601 reference design in combination with XDPL8219-U40W

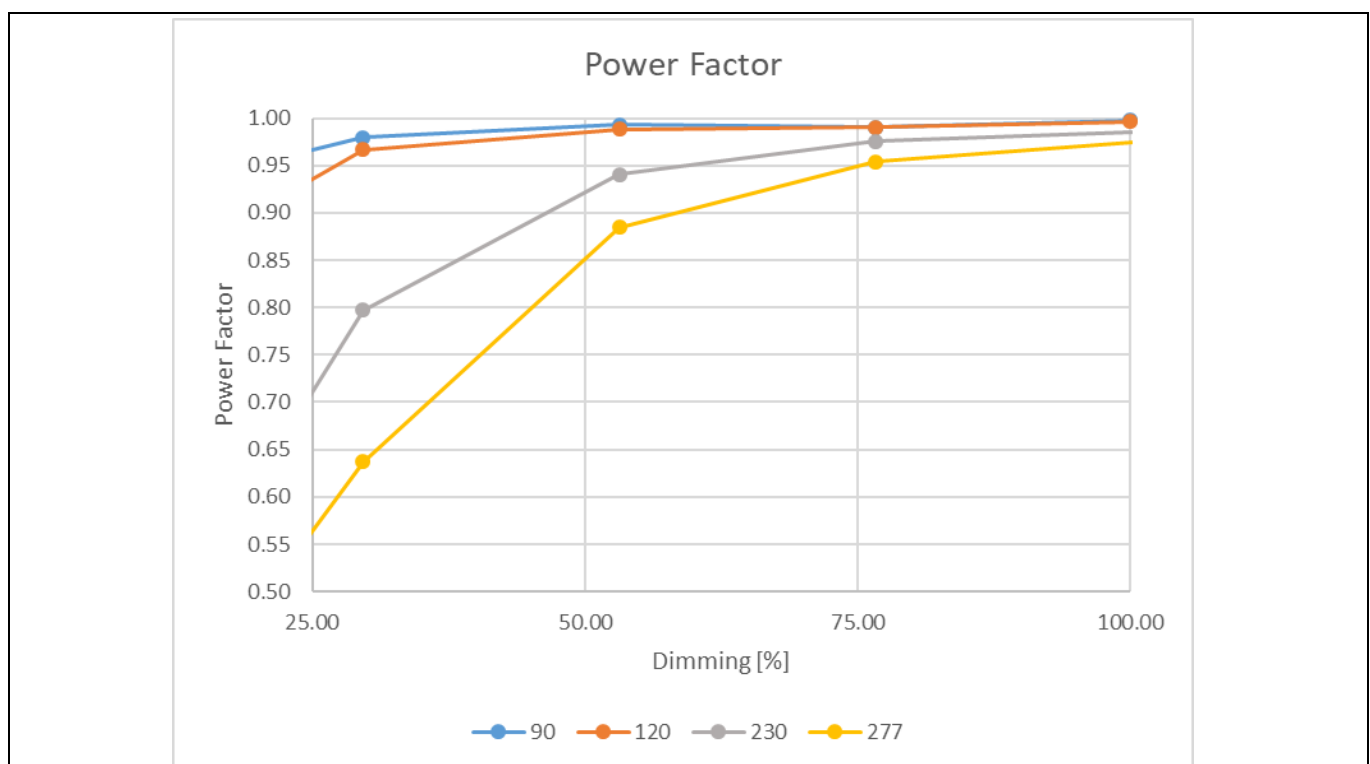


Figure 26 Power factor of BCR601 reference design in combination with XDPL8219-U40W

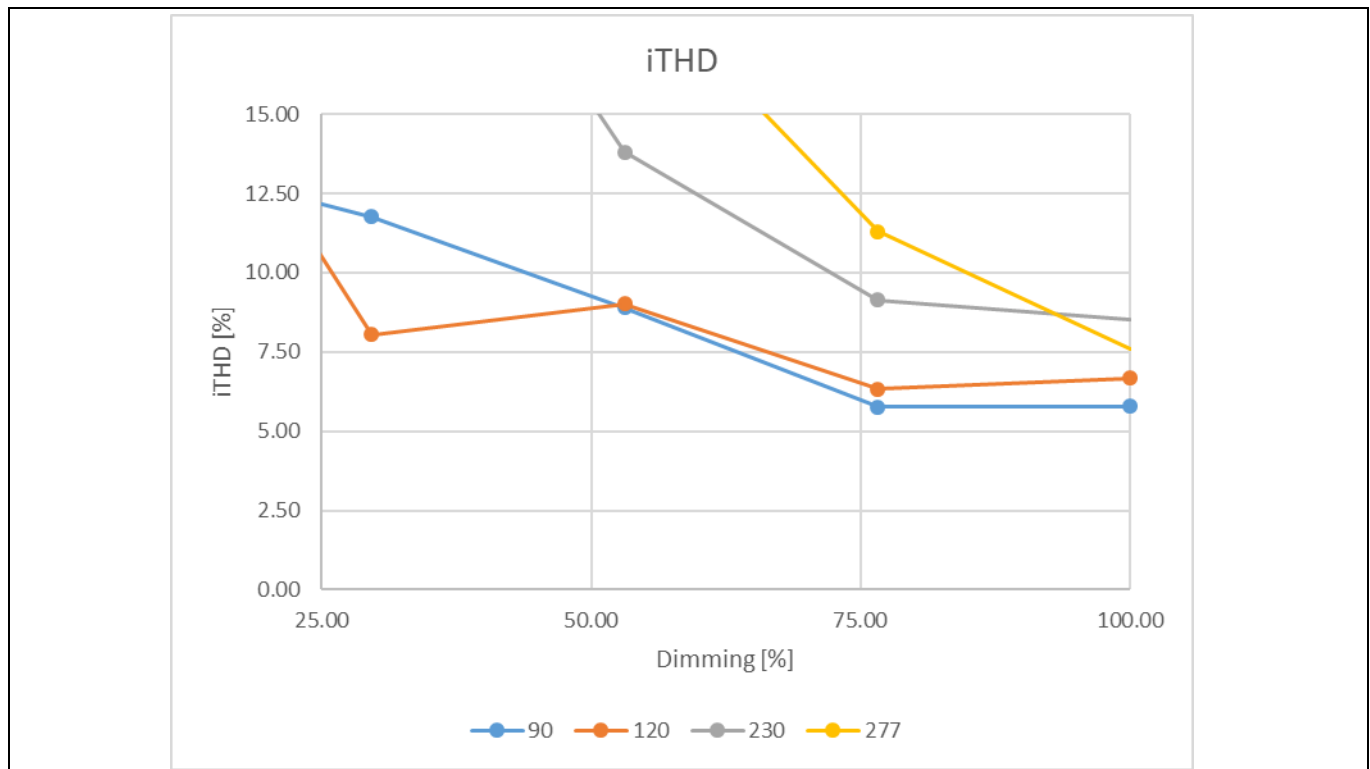


Figure 27 iTHD of BCR601 reference design in combination with XDPL8219-U40W

The full load system efficiency ([Figure 25](#)) is slightly higher than 87 percent for all input line voltages. This is an excellent result for a high-power flyback in combination with a linear second stage and is difficult to exceed even for more complex topologies.

Power factor and iTHD ([Figure 26](#) and [Figure 27](#)) are excellent over a wide dimming range.

To put these performance data into the right context, remember that the reference system's full output power of about 25 W is only 63 percent of the rated power of 40 W of the first stage. If the latter is optimized for 25 W, performance will be improved further.

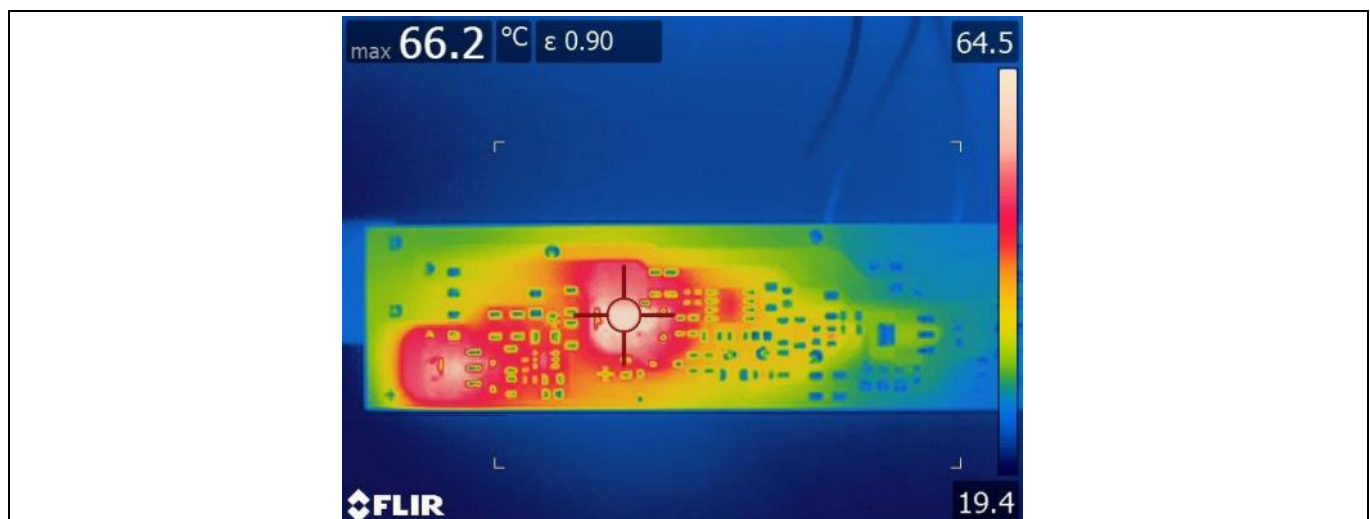


Figure 28 Thermal performance of reference design under full load conditions

3.8 Bill of materials

No.	Quantity	Designator	Description	Manufacturer	Part number
1	1	C1	470 μ F/63 V/20%	Panasonic	EEU-FC1J471
2	1	C2	2.2 μ F/25 V/1206/X7R/10%	Murata	GRM31MR71E225KA93
3	2	C4, C24	1 nF/25 V/0805/X7R/5%	Kemet	C0805C102K3RACTU
4	4	C5, C7, C8, C25	100 nF/100 V/1206/X7R/10%	Murata	GRM319R72A104KA01
5	1	C15	100 nF/50 V/0805/X7R/5%	Murata	GRM21BR71H104JA01
6	1	C27	100 nF/100 V/1206/X7R/10%	AVX	12061C104KAZ2A
7	1	C48	100 nF/50 V/603/X7R/10%	AVX	06035C104K4Z2A
8	1	C49	100 μ F/25 V/20%	Panasonic	EEUFC1E101S
9	2	D1, D21	Zener diode 27 V/SOD-123	N/A	MMSZ5254
10	2	D5, D25	BAT46W-7-F/SOD-123	N/A	BAT46W-7-F
11	1	D15	Zener diode 15 V/SOD-80C	N/A	BZV55-C15
12	1	G41	IFX1117ME V33/PG-SOT-223	Infineon Technologies	IFX1117ME V33
13	1	J1	Connector 6-pole	Würth Elektronik	691309310006
14	2	J2, J22	Connector 2-pole	Würth Elektronik	691412320002
15	2	Q2, Q22	BSP716N	Infineon Technologies	BSP716N H6327
16	3	R1, R3, R21	2.7 k/150 V/0805/1%	N/A	N/A
17	5	R2, R7, R22, R22a, R36	0 R/1206	N/A	N/A
18	1	R4	130 k/150 V/0805/1%	N/A	N/A
19	1	R6	51 k/150 V/0805/1%	N/A	N/A
20	2	R8, R16	0 R/0805	N/A	N/A
21	1	R9	1 k/150 V/0805/1%	N/A	N/A
22	1	R10	10 k/150 V/0805/1%	N/A	N/A
23	1	R11	390 k/150 V/0805/1%	N/A	N/A
24	2	R12, R32	1.6 R/200 V/1206/1%	N/A	N/A
25	2	R13, R33	1.5 R/200 V/1206/1%	N/A	N/A
26	1	R15	15 k/150 V/0805/1%	N/A	N/A
27	2	R17, R37	1.1 k/75 V/0603/1%	N/A	N/A
28	2	R18, R38	300 R/75 V/0603/1%	N/A	N/A
29	1	R19	3.3 k/150 V/0805/1%	N/A	N/A
30	1	R20	18 k/150 V/0805/1%	N/A	N/A

Tunable white LED drivers with BCR601 and BCR602

How to design cost effective, highly efficient LED drivers of highest light quality



Tunable white reference design with BCR601 and BCR602

31	1	R24	1 MEG/150 V/0805/1%	N/A	N/A
32	1	R29	1 k/200 V/1206/1%	N/A	N/A
33	1	U1	BCR601	Infineon Technologies	BCR601
34	1	U21	BCR602	Infineon Technologies	BCR602
35	1	X41	Connector 6-pole	Harting	09 18 506 7324

References

4 References

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- [2] LED professional Review (LpR), 78, S. 76-79. Luger Research e.U., ISSN 1993-890X
- [3] [Excel sheet](#)

References

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Document version	Date of release	Description of changes
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