PWM-Scheme and Current ripple of Switching Power Amplifiers

Abstract

In this work current ripple caused by switching power amplifiers is analysed for the conventional PWM (pulsewidth modulation) scheme and three-level PWM-scheme. Simplified models for estimation are introduced for the selection of the switching frequency and the inductance in series for a certain dc motor.

Introduction

Up to now dc servo motors are widely applied in small power drives because of their simple control strategy and high performance. In order to keep small power losses in the power circuit, switching power amplifiers based on power MOSFETs are often used in such a system. However, switching amplifiers produce current ripple, which is strongly associated with the PWM-scheme used in the power amplifier, the switching frequency and the inductance in the circuit. This current ripple will cause power losses in the winding and eddy current losses in the iron core. A large current ripple may cause commutation problems and even shorten the life time of a motor. For these reasons the amplitude of the current ripple must be limited in a dc motor to an acceptable value (typically < 10 % I_n).

In general, the amplitude of current ripple is reduced by increasing the indutance in the main circuit or by increasing the switching frequency of the power amplifier. It is also possible to adopt a proper PWM-scheme. In this work the current ripple will be discussed for two mainly used PWM schemes: conventional bipolar PWM and three-level PWM.

Power circuit

The switching power amplifier often used is a dc chopper as shown in *Fig. 1*. In the circuit the load of a dc motor is described with *R-L-E*, in which *E* is the induced voltage of the rotor winding, proportional to the speed of rotation. *R* and *L* are the resistance and inductance of the rotor winding, respectively.

According to the flow path of the current in the circuit, there are eight different working states, which are shown in *Fig. 2 (a)-(h)*.

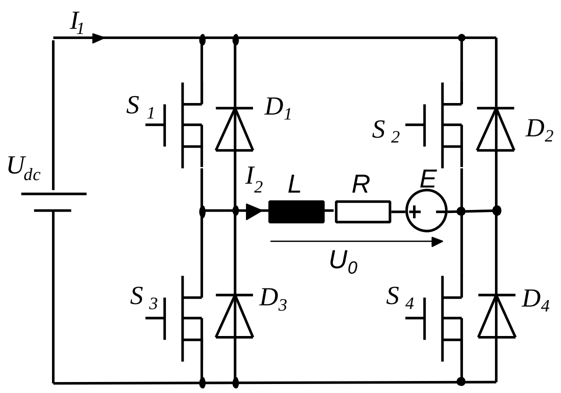


Fig. 1 DC chopper of dc drives

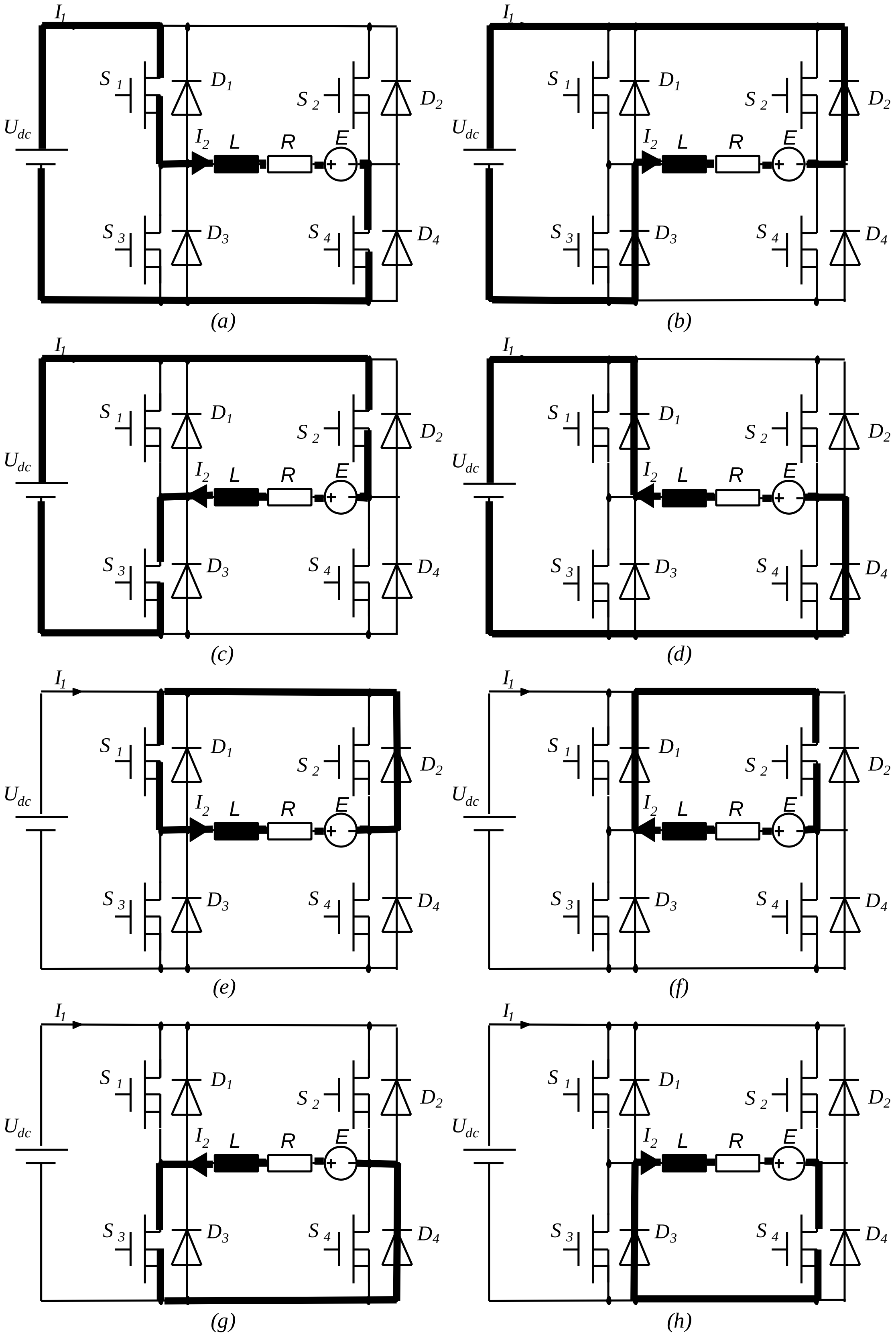


Fig. 2 Operating states of a chopper for dc servo-drive

Pulsewidth Modulation

The basic philosophy of switching amplifiers is that of pulse control in which the durations of positive, negative and zero pulses are controlled to obtain the desired average output. There are several basic principles: pulsewidth modulation (PWM), pulse frequency modulation (PFM) and current hysteresis control (CHC). For the dc servo drive, the first principle is mainly used. According to the waveform of the output voltage, there are two-level PWM-schemes and three-level PWM-schemes. A conventional scheme is a two-level PWM. In this scheme the two power switches S_1, S_4 in *Fig. 1* are controlled with the same switching signal S and the other two switches S_2, S_3 are controlled inversely with $/S$, as shown in *Fig. 3*. The switching signal is produced by the crossing points between the required voltage U_r and the triangle reference voltage U_D. The switching frequency of the power switches is constant, and equals the frequency of the triangle voltage signal. The output voltage U_0 is either U_{dc} or $-U_{dc}$. The power circuit works in the four different states as shown in *Fig. 2 (a), (b), (c),* and *(d).*

With a three-level PWM-scheme, the output voltage is switched among U_{dc}, 0, and $-U_{dc}$. A commonly used three-level PWM-scheme is the complementary PWM-scheme. As shown in *Fig. 4*, two triangle reference signals U_{D1}, U_{D2} in inverse phase are used to generate switching signals $S_{1/3}$ and $S_{4/2}$. In the power circuit, S_1 is controlled with $S_{1/3}$ and S_3 is controlled inversely with $S_{1/3}$. S_4 is controlled with $S_{4/2}$ and S_2 is inversely controlled. With this PWM-scheme, the eight states of the power circuit in *Fig. 2* are fully used to control the current. In *Fig. 4*, the conduction states of the circuit are indicated with *a...h*, corresponding to the states in *Fig. 2 (a)...(h)*. The output voltage appears at three levels, $+U_{dc}$, 0, and $-U_{dc}$. The pulsewidth of U_{dc} or $-U_{dc}$ is proportional to the required voltage U_r, and in the rest of the period, the output voltage is zero. The advantages of this PWM-scheme are: the amplitude of current ripple is very small when the induced voltage near zero (the motor works at a low speed), and the switching frequency is only half the frequency of the output voltage pulses. With the application of the complementary PWM-scheme, therefore, the switching frequency as well as the amplitude of current ripple in the motor can be reduced compared to the conventional PWM discussed above.

There are also other three-level PWM-schemes for the dc chopper. For instance the PWM-scheme used in MINIPOS is also a three-level PWM, in which the switches S_2 and S_4 are controlled according to the polarity of the required voltage, and S_1 and S_3 are switched by a PWM signal produced by a microcontroller. The switching frequency of S_1 and S_3, however, is the same as the frequency of the output voltage pulses.

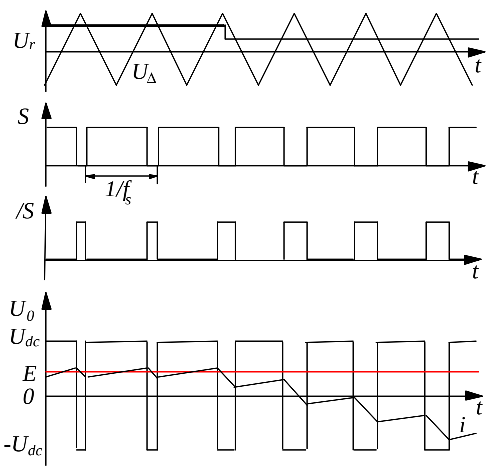


Fig. 3 Waveform of conventional two-level PWM-scheme

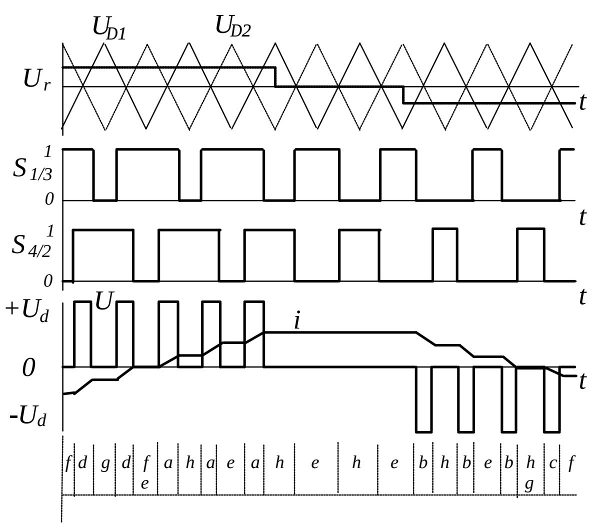


Fig. 4 Waveform of complementary PWM-scheme

Current ripple of a dc chopper

Different from an linear power amplifier, the switching power amplifier is a discontinuous element. The output voltage is a pulse series producing current distortion or current ripple in an inductive load. These current harmonics are sesponsible for extra power losses in the power circuit and the rotor winding, and eddy current losses in the iron core of the dc motor. A large current ripple may even cause commutation difficulties in the motor and overcurrent fault in the power amplifier although the dc current component is below the rating value. In this section the relation between the amplitude and switching frequency as well as the working point of a dc motor will be studied and formulas will be given for the calculation of current ripple in the conventional PWM-scheme and the complementary PWM-scheme.

As discussed above, the output voltage is only two-level when the conventional PWM-scheme is used to control the power amplifier. The current waveform in the rotor winding is shown in *Fig. 5*. When the voltage drop across power elements is neglected, the current equations are given as follows,

$$I_M = I_m e^{-\frac{t_0}{\tau}} + \frac{U_{dc} - E}{R}(1 - e^{-\frac{t_0}{\tau}}) \qquad (1)$$

$$I_m = I_M e^{-\frac{T_s - t_0}{\tau}} - \frac{U_{dc} + E}{R}(1 - e^{-\frac{T_s - t_0}{\tau}}) \qquad (2)$$

where T_s is the switching period $T_s = 1/f_s$, E is the induced voltage, R, L are the resistance and inductance of the rotor coil, respectively. t is the electrical time constant $t=L/R$. t_0 is the pulsewidth of the positive voltage

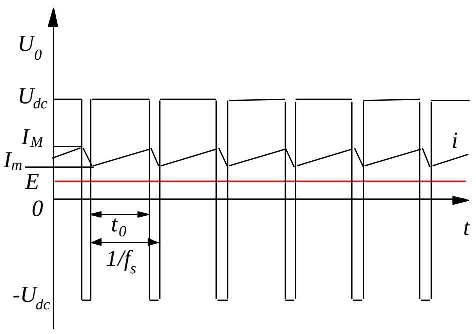


Fig. 5 Current waveform with the conventional PWM-scheme

$$t_0 = \frac{T_s}{2}(1 + \frac{I \cdot R + E}{U_{dc}})$$

and $I = \frac{1}{2}(I_M + I_m)$ is the dc component of the current.

From *Eq. (1)* and *(2)*, the peak-peak value of the current ripple is given as,

$$\Delta I_{pp} = I_M - I_m$$

$$= \frac{\frac{U_{dc}}{R}(2 - e^{-\frac{t_0}{\tau}} - e^{-\frac{T_s - t_0}{\tau}}) + I(e^{-\frac{t_0}{\tau}} - e^{-\frac{T_s - t_0}{\tau}}) + \frac{E}{R}(e^{-\frac{t_0}{\tau}} - e^{-\frac{T_s - t_0}{\tau}})}{1 + 0.5(e^{-\frac{t_0}{\tau}} + e^{-\frac{T_s - t_0}{\tau}})} \qquad (3)$$

In the steady case, the maximum current ripple appears at 50% duty cycle ($t_0=T_s/2$, $E=0$, $I=0$), and *Eq. (3)* reduces to,

$$\Delta I_{ppM} = \frac{2U_{dc}}{R} \cdot \frac{1 - e^{-\frac{T_s}{2\tau}}}{1 + e^{-\frac{T_s}{2\tau}}} \qquad (4)$$

In the same way, the equation of the current ripple (*Fig. 6*) can be given for a three-level PWM-scheme. When the voltage drop across the power elements is neglected, the current ripple is,

$$\Delta I_{pp} = I_M - I_m$$
$$= \frac{\frac{U_{dc}}{R}(1-e^{-\frac{t_0}{\tau}}) + I(e^{-\frac{t_0}{\tau}} - e^{-\frac{T-t_0}{\tau}}) + \frac{E}{R}(e^{-\frac{t_0}{\tau}} - e^{-\frac{T-t_0}{\tau}})}{1+0.5(e^{-\frac{t_0}{\tau}} + e^{-\frac{T-t_0}{\tau}})} \qquad (5)$$

where T is the pulse period of the output voltage. In the complementary PWM-scheme, $T=1/(2f_s)$, i.e. only half the value of T_s, and

$$t_0 = T\frac{I \cdot R + E}{U_{dc}}$$

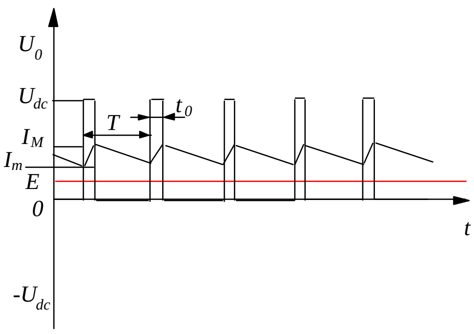


Fig. 6 Current waveform with a three-level PWM-scheme

In the steady case, the maximum current ripple appears also at 50% duty cycle ($t_0=T/2$, $E=U_{dc}/2$, $I=0$), and *Eq. (5)* reduces to,

$$\Delta I_{ppM} = \frac{U_{dc}}{R} \cdot \frac{1-e^{-\frac{T}{2\tau}}}{1+e^{-\frac{T}{2\tau}}} \qquad (6)$$

In *Fig. 7* a comparison of current ripple as a function of the induced voltage of the rotor coil between conventional PWM and three-level PWM is given at the same pulse frequency of the output voltage. It shows that the maximum current ripple in the conventional PWM-scheme is at E=0 and twice as high as in the three-level PWM-scheme where the maximum

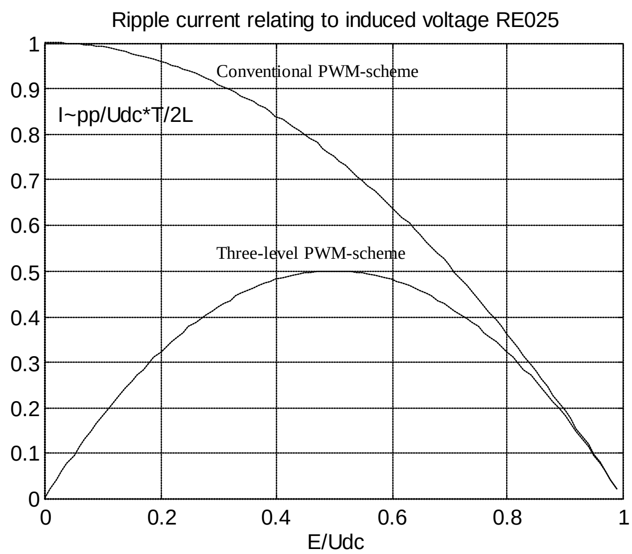


Fig. 7 Comparison of current ripple between the conventional PWM-scheme and a three-level PWM-scheme

lies at $E=U_{dc}/2$. As a result, for a given amplitude of current ripple, the switching frequency of the complementary PWM-scheme is only one fourth of that of the conventional PWM-scheme. Furthermore, since the current ripple of a three-level PWM-scheme is very small at low speeds, it is particularly appropriate for the positionning tasks of a dc motor servo drive.

Practical estimation of the maximum current ripple

In the engineering design of dc servo drives, it is required to determine the switching frequency and the inductance in order to reduce the maximum current ripple below an acceptable limit. For this purpose, it is possible to use a simple equation to approximate the current ripple. Here, two approximate equations are given for the conventional PWM-scheme and the three-level PWM-scheme, respectively.

In the conventional PWM-scheme, the maximum current ripple is given by *Eq. (4).* When the pulse period T_s of the output voltage is much smaller than the electrical time constant ($T_s << t$), *Eq. (4)* can be simplified to,

$$\Delta I_{ppM} = \frac{U_{dc}T_s}{2L} = \frac{U_{dc}}{2Lf_s} \qquad (7)$$

The error of the estimation is associated with the ratio $T_s/2t$. An error simulation is given in *Fig. 8*. The error of the estimation will be smaller than 2% when $T_s < t$.

To a three-level PWM-scheme, the maximum current ripple can also estimated with a simplified equation from *Eq. (6)*, when T is small enough.

$$\Delta I_{ppM} = \frac{U_{dc}T}{4L} \qquad (8)$$

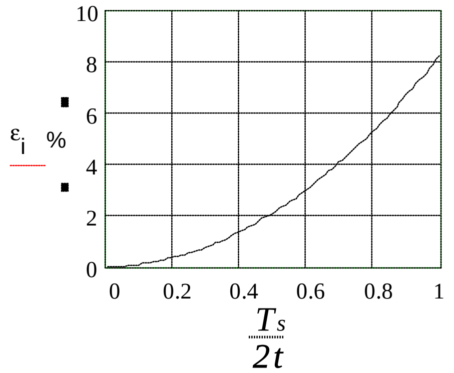


Fig. 8 Relative error of the estimation

The result of the error analysis above is valid also for *Eq. (8)*.
From *Eq. (7)* and *(8)*, the maximum current ripple is mainly dependent on the dc-link voltage, the switching frequency, and the inductance of the circuit.

Practical application of estimation formulas

In practical application, we can use the two approximate formulas to estimate the current ripple in a dc servo motor. On another hand, we can also use them to calculate the inductance required in the circuit in order to limit the amplitude of current ripple. The condition to use *Eq. (7)* and *(8)* is that the pulse period must be smaller than the electrical time constant of the rotor coil. For MAXON dc motors, the electrical time constant ranges from 0.08 ms to 0.25 ms, corresponding to a switching frequency range from 4 to 12.5 kHz. Usually, the switching frequency of the amplifiers used for the MAXON motors is higher than 20 kHz. Therefore, the estimation of the amplitude of current ripple with the equations *Eq. (7)* and *(8)* is reliable. Two calculation examples and a practical measurement to test the calculated results are given in appendix.

Conclusion

In this work, the current ripple produced by switching amplifiers for DC MAXON motors are studied in detail. For practical purposes two simple formulas can be used to estimate the maximum current ripple and determine the inductance required to limit the current ripple.

Appendix

Example 1:

The calculation of the current ripple in a DC motor supplied with a switching power amplifier, conventional PWM-scheme

Motor 2260.885-73.216-200

Motor parameter: U_n= 24 V, I_n=3.3 A, L_m= 560 mH, R_m=1.35 Ω

Power amplifier: 4-Q Servo amplifier, MMC-QR060058-05PD00A, Serie Nr. 0633, U_{dc}=48 V, f_s=26.7 kHz

Additional external inductance: L_s=600 µH

Electrical time constant: $\tau = \dfrac{L_m + L_s}{R_m} = 0.86\,ms$

The maximum current ripple is given by *Eq. (4)*,

$$\Delta I_{ppM} = \frac{2U_{dc}}{R} \cdot \frac{1 - e^{-\frac{1}{2f_s\tau}}}{1 + e^{-\frac{1}{2f_s\tau}}} = \frac{2 \times 48V}{1.35\Omega} \cdot \frac{1 - e^{-\frac{1}{2 \times 26.7\,kHz \times 0.86\,ms}}}{1 + e^{-\frac{1}{2 \times 26.7\,kHz \times 0.86\,ms}}} = 0.775\,A$$

or the maximum current ripple can be estimated with *Eq. (7)*

$$\Delta I_{ppM} = \frac{U_{dc}}{2Lf_s} = \frac{48V}{2 \times 1.16\,mH \times 26.7\,kHz} = 0.775\,A$$

The results from *Eq. (4)* and *Eq. (7)* are the same.

If the current ripple is to be limited to below 10% of I_n, the total inductance in the circuit according to *Eq. (7)* should be,

$$L = \frac{U_{dc}}{2\Delta I_{ppM} f_s} = \frac{48V}{2 \times 10\% \times 3.3\,A \times 26.7\,kHz} = 2.724\,mH$$

Hence, the extra inductance in series with the rotor coil is now (instead of 0.6 mH),

$$L_s = L - L_m = 2.724\,mH - 0.56\,mH = 2.164\,mH$$

Or the switching frequency should be increased without any other inductance in series, according to *Eq. (7)*,

$$f_s = \frac{U_{dc}}{2 \times \Delta I_{ppM} \times L_m} = \frac{48V}{2 \times 10\% \times 3.3\,A \times 0.56\,mH} = 130\,kHz$$

Example 2:
The calculation of the current ripple in a DC motor supplied with a switching power amplifier, three-level PWM-scheme
Motor RE025-055-35EBA201A
Motor parameter: U_n= 24 V, I_n=1.22 A, L_m= 240 mH, R_m=2.34 Ω
Power amplifier: MINIPOS, U_{dc}=24 V, f_s=60 kHz
Additional external inductance: L_s=400 µH
Electrical time constant: $\tau = \dfrac{L_m + L_s}{R_m} = 0.274ms$
The maximum current ripple is given by *Eq. (6)*,

$$\Delta I_{ppM} = \frac{U_{dc}}{R} \cdot \frac{1 - e^{-\frac{1}{2 f_s \tau}}}{1 + e^{-\frac{1}{2 f_s \tau}}} = \frac{24V}{2.34\Omega} \cdot \frac{1 - e^{-\frac{1}{2 \times 60\,kHz \times 0.274ms}}}{1 + e^{-\frac{1}{2 \times 60\,kHz \times 0.274ms}}} = 0.156\,A$$

or the maximum current ripple can be estimated with *Eq. (8)*

$$\Delta I_{ppM} = \frac{U_{dc}}{4Lf_s} = \frac{24V}{4 \times 0.64mH \times 60kHz} = 0.156\,A$$

If the current ripple is to be limited below 10% of I_n, the total inductance in the circuit according to *Eq. (8)* should be,

$$L = \frac{U_{dc}}{4\Delta I_{ppM} f_s} = \frac{24V}{4 \times 10\% \times 1.22\,A \times 60kHz} = 0.82mH$$

Then, the inductance in series is now (instead of 0.4 mH),

$$L_s = L - L_m = 0.82mH - 0.24mH = 0.58mH$$

Or the switching frequency should be increased without any other inductances in series, according to *Eq. (7)*,

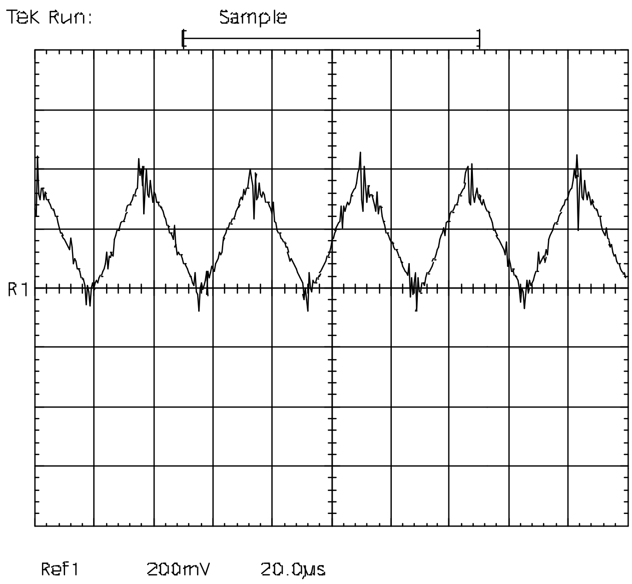
$$f_s = \frac{U_{dc}}{4 \times \Delta I_{ppM} \times L_m} = \frac{24V}{4 \times 10\% \times 1.22\,A \times 0.24mH} = 205kHz$$

Note: in this example, the dc-link voltage U_{dc} equals the rating voltage of the dc motor U_n. In order to keep the dynamical performance of the motor near the rating operating point, the dc-link voltage should be about twice the rating voltage. However, U_{DS} (drain-source)of the MOSFETs in MINIPOS is 50 V, and therefore, the dc-link voltage should be kept below 25 V. For this reason, MINIPOS power stage is not appropriate for dc motors with a rating voltage higher than 16 V.

Practical measurement of current ripple
Motor: 2260.885-73 216-200,
Power amplifier: 4-Q Servo amplifier, MMC-QR060058-05PD00A, Serie Nr. 0633
f_s= 26.7 kHz, U_{dc} = 48 V, L_s = 600 µH

In the following, two current waveforms of an oscilloscope are shown.
Unit of current measurement: 2.2 A/V

1) U_{nsoll} = 0 V, zero speed



2) U_{nsoll} = 9V, high speed