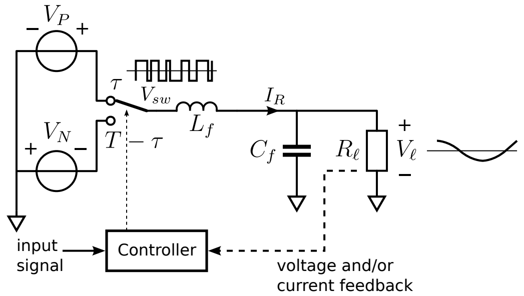
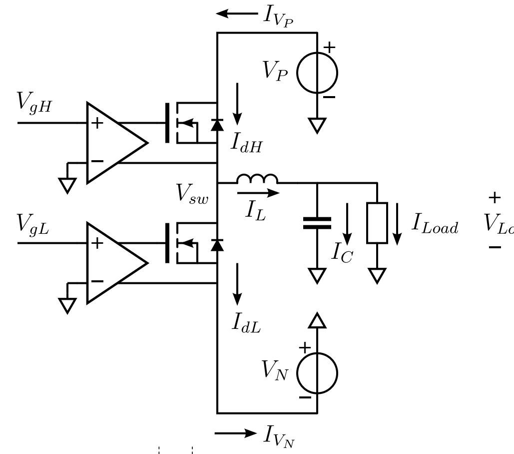
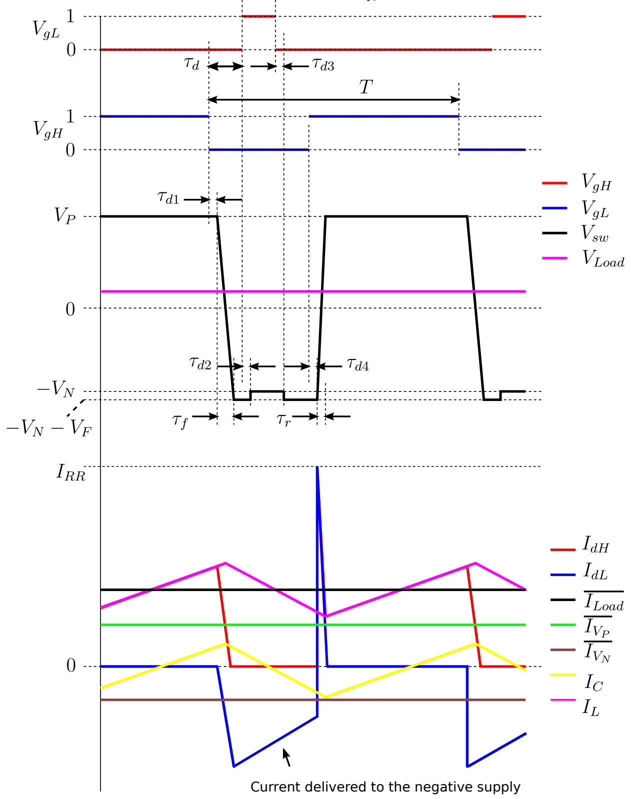
Class D amplifier design

Half-bridge operating principle



Half-bridge switching behavior and power losses



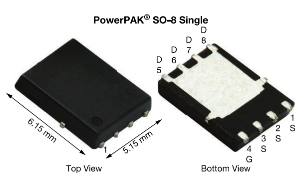


Si7454FDP

www.vishay.com

Vishay Siliconix

N-Channel 100 V (D-S) MOSFET



PRODUCT SUMMARY	
V_{DS} (V)	100
$R_{DS(on)}$ max. (Ω) at V_{GS} = 10 V	0.0295
$R_{DS(on)}$ max. (Ω) at V_{GS} = 4.5 V	0.034
Q_g typ. (nC)	8
I_D (A)	23.5
Configuration	Single

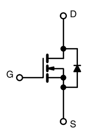
FEATURES

- TrenchFET® Gen IV power MOSFET
- Very low R_{DS} x Q_g figure-of-merit (FOM)
- Tuned for the lowest R_{DS} x Q_{oss} FOM
- 100 % R_g and UIS tested
- Material categorization: for definitions of compliance please see www.vishay.com/doc?99912

RoHS COMPLIANT HALOGEN FREE

APPLICATIONS

- Synchronous rectification
- Primary side switch
- DC/DC converters
- Power supplies
- Motor drive control



N-Channel MOSFET

Output voltage and ripple current

$$V_\ell = \frac{\tau}{T}(V_P+V_N) - V_N$$
$$V_S = V_P = V_N$$
$$\delta = \frac{\tau}{T}$$
$$V_\ell = V_S(2\delta - 1)$$
$$I_R = \frac{2V_s\delta(1-\delta)T}{L}$$

$I_R =$ peak-to-peak value of the ripple current.

Expressions for peak ripple current much smaller than load current

MOS conduction losses

$$P_{cM} \approx I_{load}^2 R_{dsOn}\frac{T-2\tau_d}{T}$$

$R_{dsOn} =$ MOS on resistance

Body diode conduction losses

$$P_{cD} \approx I_{load}V_F\frac{2\tau_d}{T}$$

$V_F =$ MOS body diode forward voltage at load current

Output charge losses

$$P_{oQ} \approx \frac{2}{T}Q_o(V_P+V_N)$$

$Q_o =$ MOS output charge

or

$$P_{oQ} \approx \frac{2}{T}C_o(V_P+V_N)^2$$

$C_o =$ MOS output capacitance

Reverse recovery losses

$$P_{RR} = \frac{1}{T}Q_{RR}(V_P+V_N)$$

$Q_{RR} =$ MOS body diode reverse recovery charge

Switching losses

$$P_{sw} = \frac{2}{T}\int_0^{\tau_{sw}}(V_P+V_N)\left(1-\frac{t}{\tau_{sw}}\right)I_{load}\frac{t}{\tau_{sw}}dt = \frac{\tau_{sw}I_{load}(V_N+V_P)}{3T}$$

$$\tau_{sw} \approx \frac{C_{gd}(V_P+V_N)}{I_G}$$

$C_{gd} =$ MOS gate-drain capacitance

$I_G =$ Gate driver current

Gate driver losses

$$P_G = \frac{1}{T}V_G^2C_{iss}$$

or

$$P_G = \frac{1}{T}V_GQ_G$$

$V_G =$ Gate driver voltage

$C_{iss} =$ MOS input capacitance

$Q_G =$ MOS gate charge

Low losses and low EMI during quiescent operation:
Resonant switching:

$$I_R\tau_d \approx 2(V_P+V_N)C_{oss}$$

$C_{oss} =$ MOS output capacitance

FOM power devices:
Combine a low charge storage with a low on resistance.

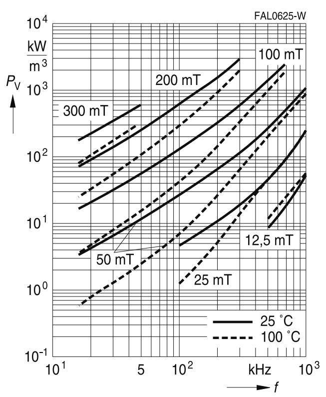
Core losses

- Manufacturer's calculation tools
- Material specification:

TDK

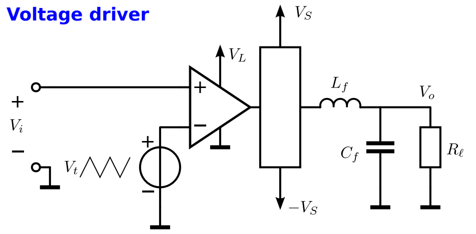
SIFERRIT m
N97

Relative core losses
versus frequency
(measured on R34 toroids)



Fixed frequency PWM

Voltage driver



$V_t =$ peak-to-peak value of triangular voltage

Transfer

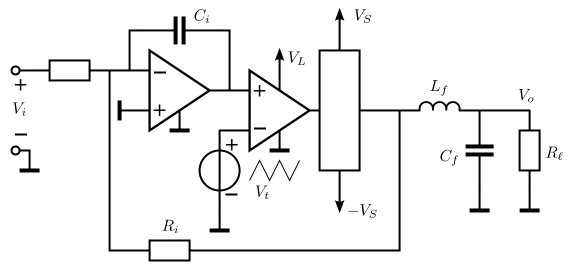
$$\frac{V_o}{V_i} = \frac{2V_S}{V_t}\frac{1}{1+\frac{s}{Q_\ell\omega_f}+\frac{s^2}{\omega_f^2}}$$

$Q_\ell =$ Quality factor of the load

$\omega_f =$ Corner frequency of the filter

Nonlinearity caused by:
- dead zone
- signal-dependent delay
- signal-dependent rise and fall times

Can be reduced through application of negative feedback



Small-signal bandwidth: firt-order LP product: $B_f = \frac{2V_S}{2\pi V_t R_i C_i}$

Large-signal stability:
Rate of change at the output of the integrator should not exceed that of the triangular signal: $f_{sw} > \pi B_f$

Further improvements (both methods require frequency compensation):
- Include output filter in the feedback loop
- Include one more integrator in the feedback loop

Influence of timing noise in the MOS driver of the noise performance

Noisy PWM resulting from timing noise in the MOS driver.

Spectral density: $S_\tau\left[\frac{s^2}{\mathrm{Hz}}\right]$

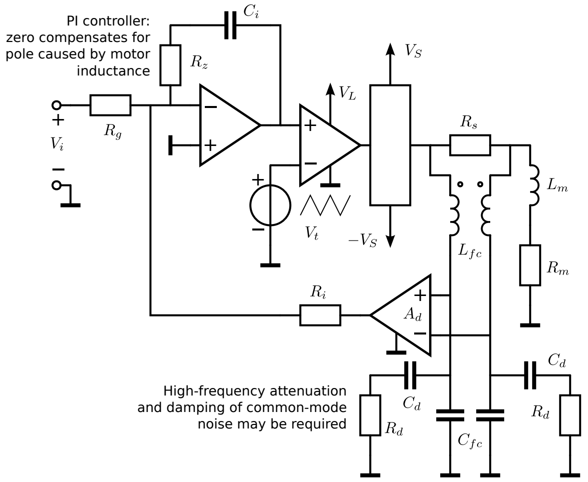
Equivalent voltage noise density spectrum at the input of the comparator:

$$S_v = S_\tau\frac{V_S^2}{T^2}\left[\frac{V^2}{\mathrm{Hz}}\right]$$

In isolated MOS / IGBT drivers: spurious frequency components may be present.

Current driver for inductive loads

Without output filter



Compensation of the effect of a small cable capacitance requires two extra (complex) poles in the loop.

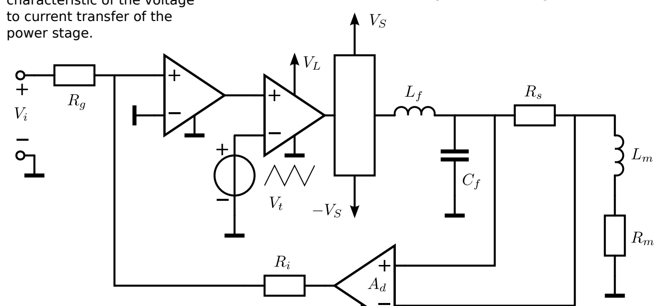
With output filter

An output filter can be applied to:
- Relax the CMRR requirements and/or CM filter requirements for the current sense circuit
- Reduce the influence of high-frequency through the cable capacitance on the transfer
- Reduce emission levels

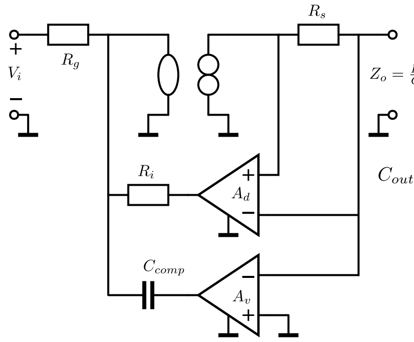
The output filter strongly complicates the design of the dynamic performance and the frequency stability of the current driver. An over sampling digital controller with integrated pulse width modulator may be a flexible and attractive alternative for an analog controller.

Analog I controller with compensation of the frequency characteristic of the voltage to current transfer of the power stage.

The effect of the cable capacitance can be reduced using remote sensing.



The effect of the cable capacitance and the motor capacitance can be reduced through application of capacitive voltage feedback at the load, thereby creating a negative output capacitance that compensates for the load capacitance.



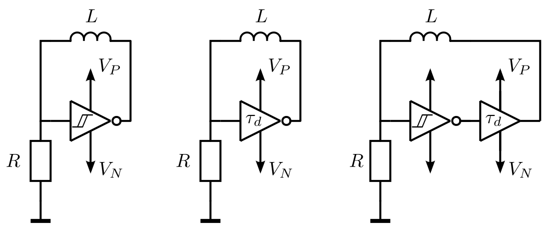
$$Z_o = \frac{D}{C} = -\frac{A_dR_s}{sR_iA_vC_{comp}}$$

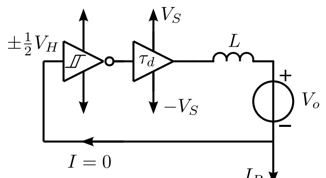
$$C_{out} = -C_{comp}\frac{A_vR_i}{A_dR_s}$$

Fixed ripple current PWM

Hysteresis-based self-oscillating class D voltage driver

First-order LR oscillators





Basic operation

Schmitt-trigger switching levels: $\pm\frac{1}{2}V_H$

Peak-to-peak value of the ripple current:

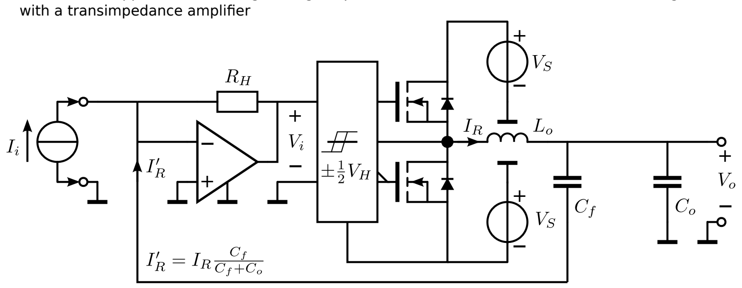
$$I_R = \frac{V_H}{R} + \frac{2V_S}{L}\tau_d$$

Oscillation frequency:

$$f = \frac{1}{T} = \frac{1-\frac{V_o^2}{V_S^2}}{\frac{2LV_H}{RV_S}+4\tau_d}$$

Alternative configuration

Measure the ripple current through a large capacitor attenuate it and convert it into a voltage with a transimpedance amplifier



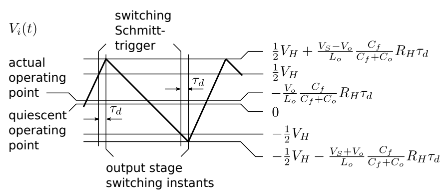
$$I'_R = I_R\frac{C_f}{C_f+C_o}$$

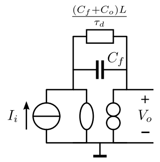
If the delay in the oscillator equals zero, the mean value of the voltage at the input of the comparator does not depend on the signal and the circuit behaves as an ideal transimpedance integrator: $\frac{V_o}{I_i} = -\frac{1}{sC_f}$

The output filter does not introduce (observable) complex poles in this transfer.

The DC transfer can be fixed accurately:
- Resistor in parallel with C_f
- Extra feedback loop with non-inverting integrator (with a zero) for lower distortion

A finite delay limits the gain at low frequencies:

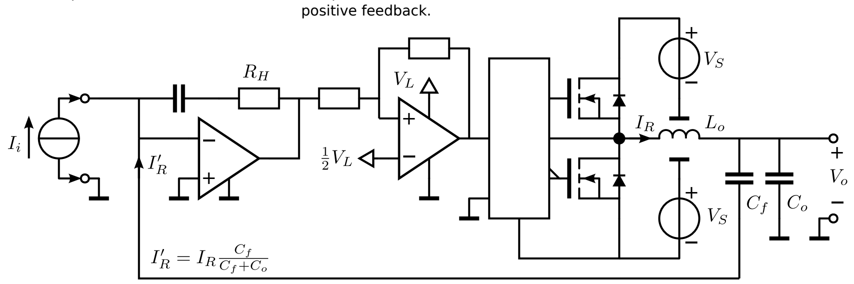




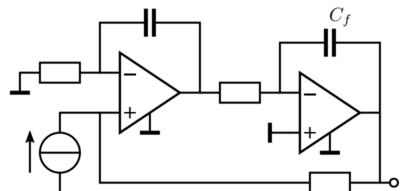
The reduction of the low-frequency gain can be compensated for by inserting a capacitor in series with

For a low timing noise, construct the Schmitt-trigger with a low noise comparator with positive feedback.

$$\frac{V_o}{I_i} = -\frac{(C_f+C_o)L}{C_f\tau_d}\frac{1}{1+s\frac{(C_f+C_o)L}{\tau_d}}$$



$$I'_R = I_R\frac{C_f}{C_f+C_o}$$



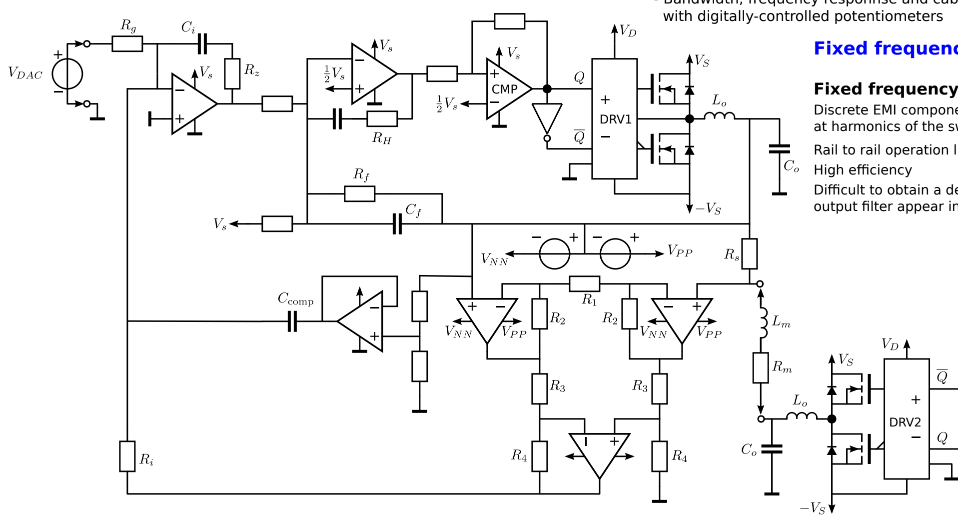
First stage: extra loop integrator
Second stage: Self oscillating class D transimpedance integrator (from figure above)

Hysteresis-based self-oscillating class D current driver

Simplified diagram

Putting it all together
Further detailing required
Improvement: differential voltage sensing

- PI controller with zero on motor time constant for setting the bandwidth
- Fixed DC gain in power stage for adjusting the frequency response
- Low-noise current-sense amplifier as with over-all feedback class AB driver
- Cable capacitance compensation
- Full-bridge amplifier by driving the MOS drivers of bothe halves with opposite phase
- Bandwidth, frequency responsne and cable capacitance compensation can be made adjustable with digitally-controlled potentiometers



Fixed frequency versus fixed ripple current driver

Fixed frequency

Discrete EMI components with relatively high amplitude at harmonics of the switching frequency

Rail to rail operation limited by minimum pulse width

High efficiency

Difficult to obtain a desired frequency response because the poles of the output filter appear in the loop gain of the current control loop

Fixed ripple current

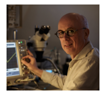
Wide spectrum EMI with lower amplitude at unknown frequencies

Ripple voltage increases with voltage excursion

Switching frequency drops below Nyquist at large voltage excursions

Higher efficiency because switching losses drop at large voltage excursions

Relatively easy to obtain a desired frequency response because the poles of the output filter do not appear in the loop gain of the current control loop.



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