Power Electronics

ODL WEEK 2 LECTURE

- Contents for this lecture from the course plan for ODL

ODL Week 2
PWM Inverters (contd.) and Resonant Pulse Inverters:

Design of single phase PWM inverters
Intro. to Resonant converters
Series resonant inverter with unidirectional switching


بسم الله الرحمن الرحيم

8-2-2 SQUARE-WAVE SWITCHING SCHEME

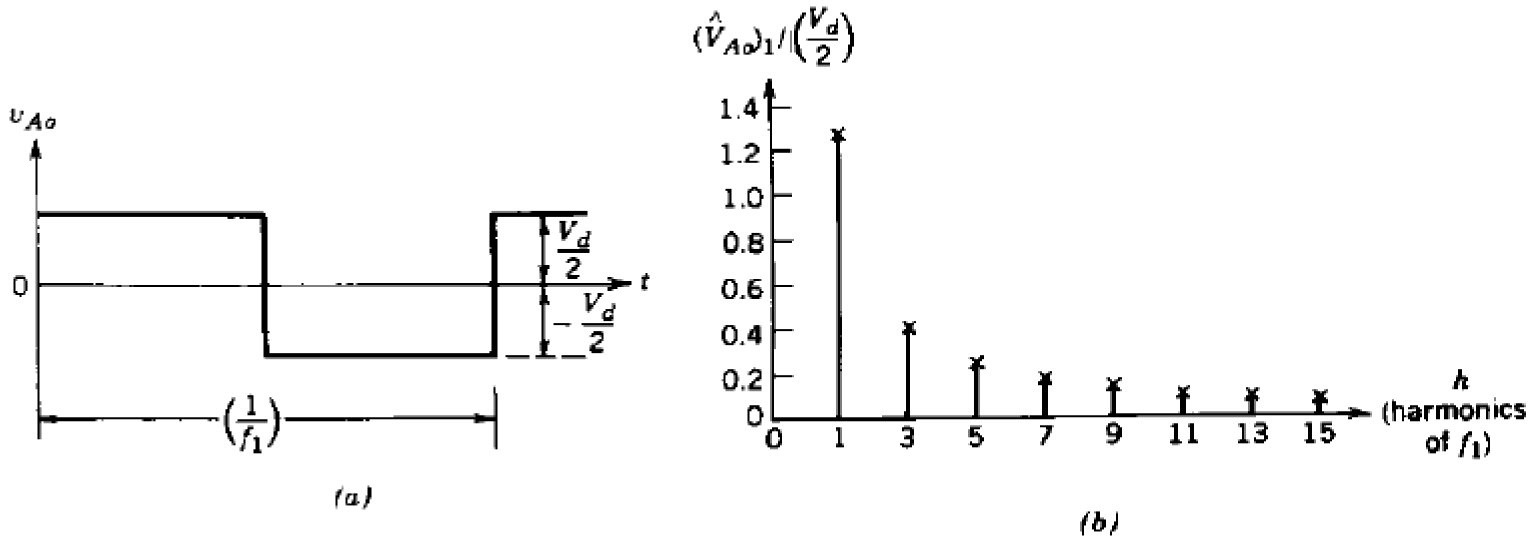


Figure 8-9 Square-wave switching.

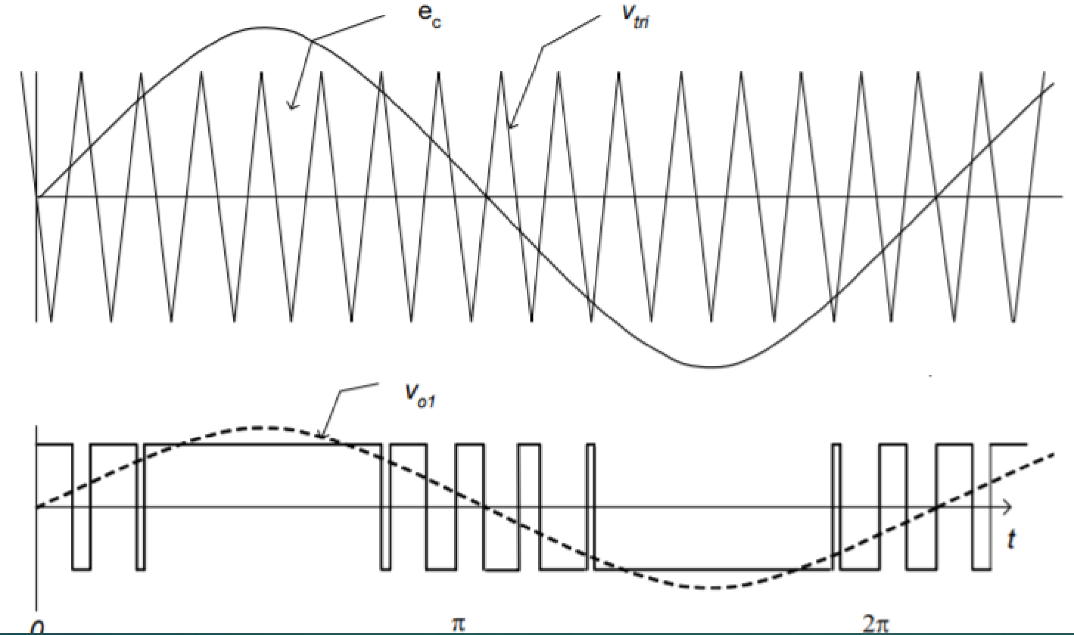
In the square-wave switching scheme, each switch of the inverter leg of Fig. 8-4 is on for one half-cycle (180°) of the desired output frequency. This results in an output voltage waveform as shown in Fig. 8-9*a*. From Fourier analysis, the peak values of the fundamental-frequency and harmonic components in the inverter output waveform can be obtained for a given input V_d as

$$(\hat{V}_{Ao})_1 = \frac{4}{\pi}\frac{V_d}{2} = 1.273\left(\frac{V_d}{2}\right)$$

$$(\hat{V}_{Ao})_h = \frac{(\hat{V}_{Ao})_1}{h}$$

the square-wave switching is also a special case of the sinusoidal PWM switching when m_a becomes so large that the control voltage waveform intersects with the triangular waveform in Fig. 8-5a only at the zero crossing of $v_{control}$. Therefore, the output voltage is independent of m_a in the square-wave region, as shown in Fig. 8-8.

One of the advantages of the square-wave operation is that each inverter switch changes its state only twice per cycle, which is important at very high power levels where the solid-state switches generally have slower turn-on and turn-off speeds. One of the serious disadvantages of square-wave switching is that the inverter is not capable of regulating the output voltage magnitude. Therefore, the dc input voltage V_d to the inverter must be adjusted in order to control the magnitude of the inverter output voltage.



8-3 SINGLE-PHASE INVERTERS

8-3-1 HALF-BRIDGE INVERTERS (SINGLE PHASE)

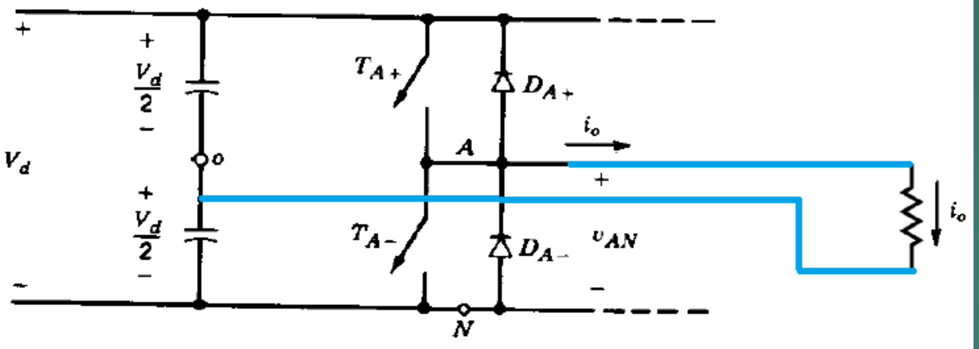


Figure 8-4 One-leg switch-mode inverter.

Sufficiently large capacitances should be used such that it is reasonable to assume that the potential at point o remains essentially constant with respect to the negative dc bus N. Therefore, this circuit configuration is identical to the basic one-leg

In a half-bridge inverter, the peak voltage and current ratings of the switches are as follows:

$$V_T = V_d \tag{8-15}$$

and

$$I_T = i_{o,\text{peak}} \tag{8-16}$$

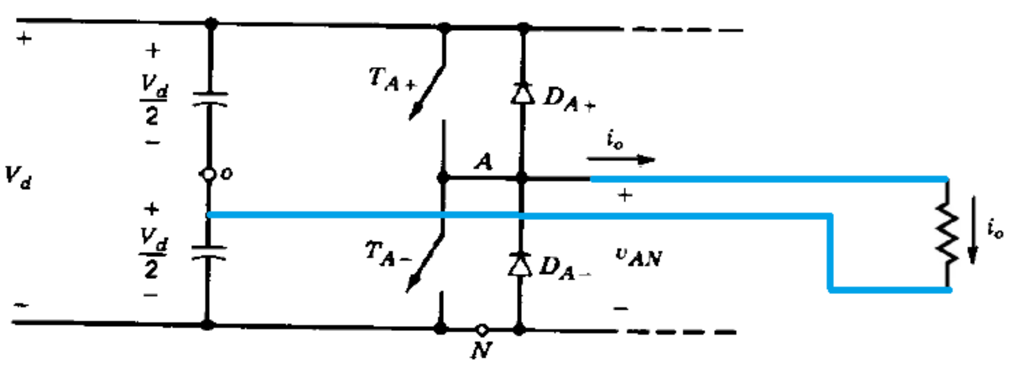


Figure 8-4 One-leg switch-mode inverter.

In a half-bridge inverter, the peak voltage and current ratings of the switches are as follows:

$$V_T = V_d \quad (8\text{-}15)$$

and

$$I_T = i_{o,\text{peak}} \quad (8\text{-}16)$$

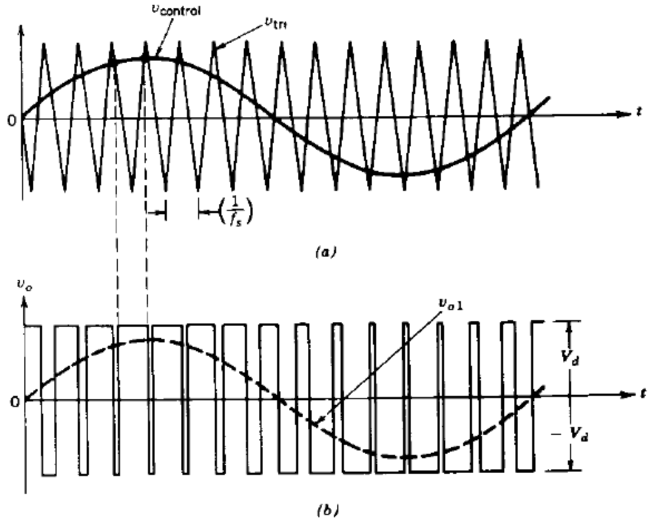
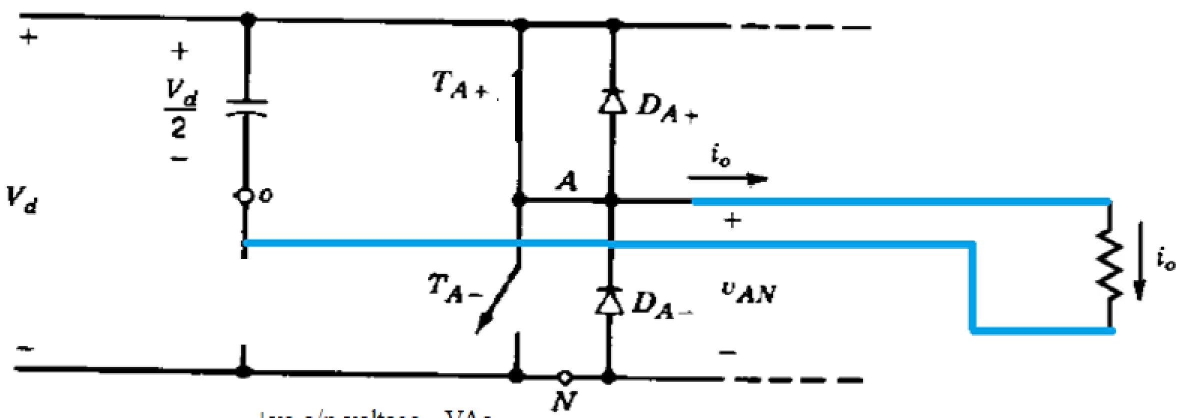
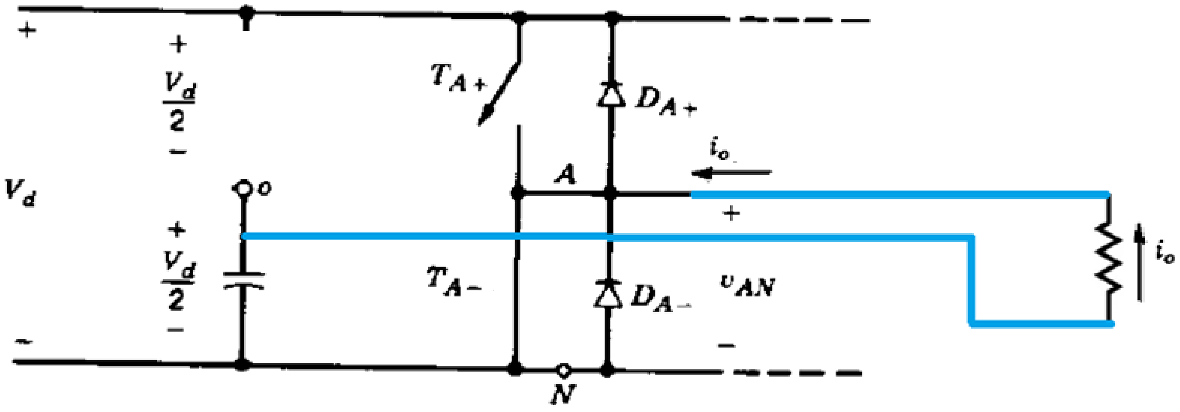


Figure 8-12 PWM with bipolar voltage switching.



+ve o/p voltage - VAo



-ve o/p voltage VAo

8-3-2 FULL-BRIDGE INVERTERS (SINGLE PHASE)

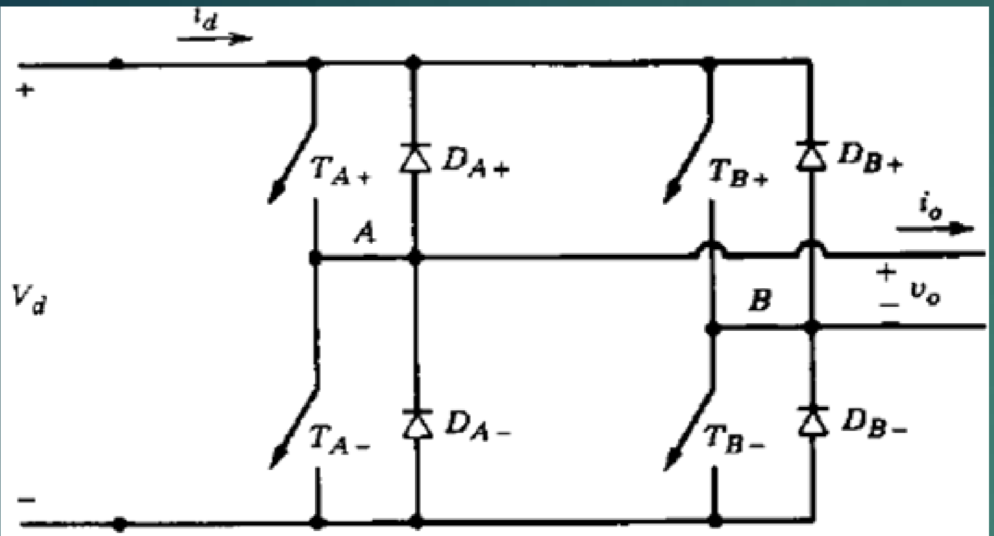


Figure 8-11 modified
Single-phase full-bridge inverter.

With the same dc input voltage, the maximum output voltage of the full-bridge inverter is twice that of the half-bridge inverter. This implies that for the same power, the output current and the switch currents are one-half of those for a half-bridge inverter. At high power levels, this is a distinct advantage, since it requires less paralleling of devices.

Here, the diagonally opposite switches (T_{A+}, T_{B-}) and (T_{A-}, T_{B+}) from the two legs in Fig. 8-11 are switched as switch pairs 1 and 2, respectively.

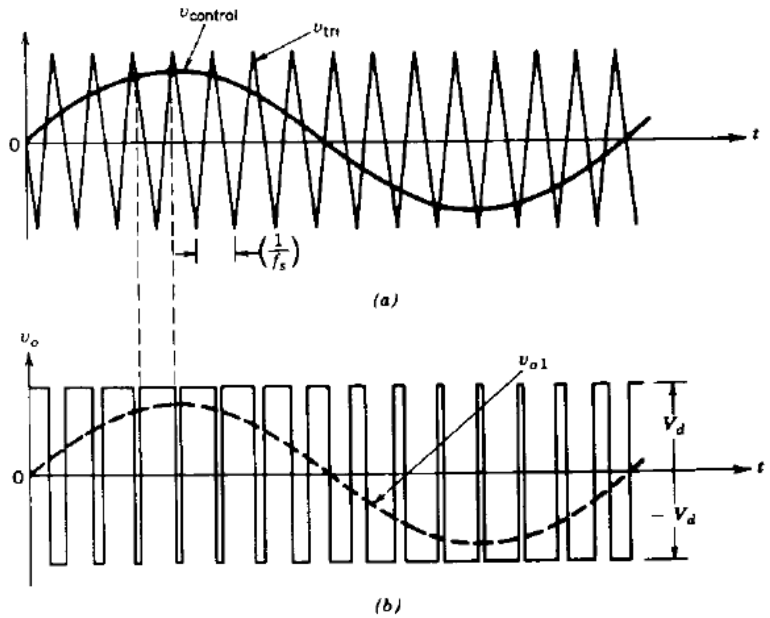


Figure 8-12 PWM with bipolar voltage switching.

Working of diodes

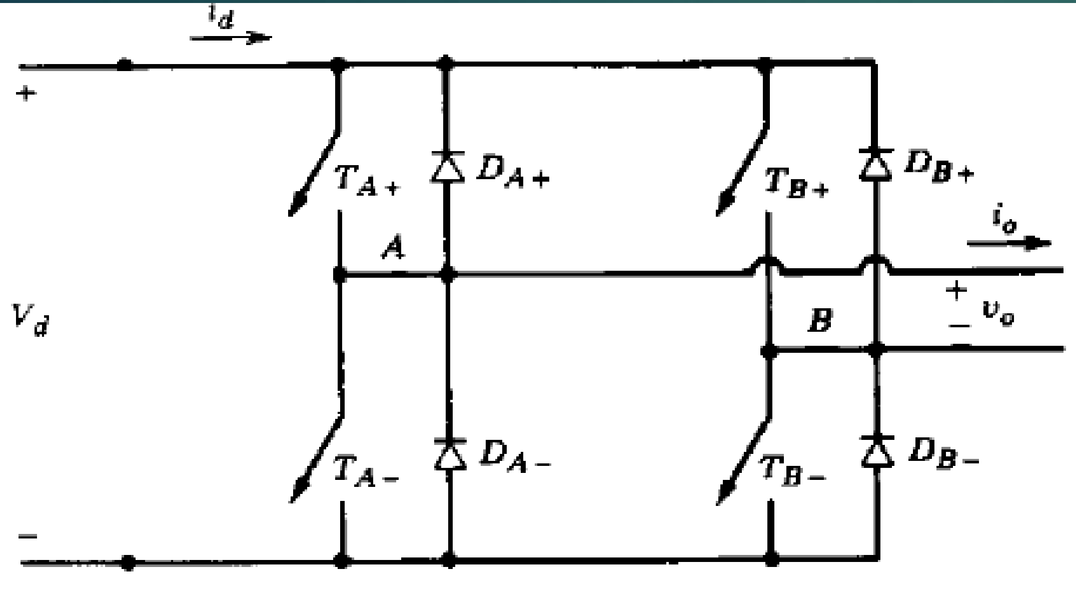
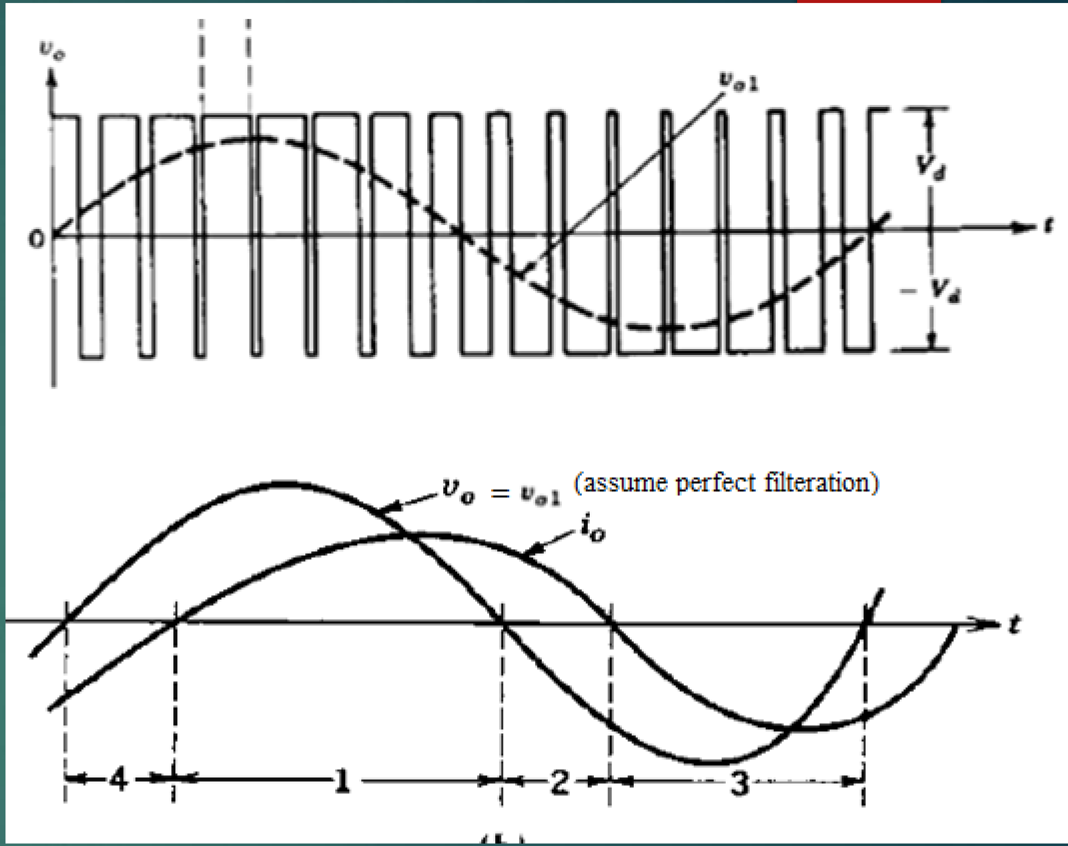
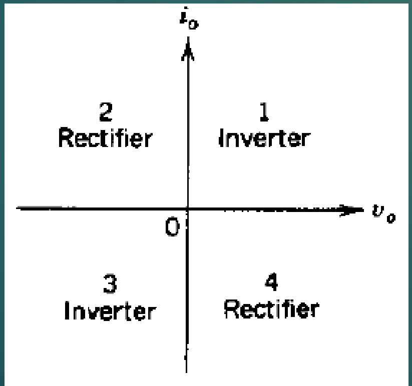


Figure 8-11 Single-phase full-bridge inverter.





(c)

■ ***Example 8-2*** In the full-bridge converter circuit of Fig. 8-11, $V_d = 300$ V, $m_a = 0.8$, $m_f = 39$, and the fundamental frequency is 47 Hz. Calculate the rms values of the fundamental-frequency voltage and some of the dominant harmonics in the output voltage v_o if a PWM bipolar voltage-switching scheme is used.

Solution From Eq. 8-18, the harmonics in v_o can be obtained by multiplying the harmonics in Table 8-1 and Example 8-1 by a factor of 2. Therefore from Eq. 8-11, the rms voltage at any harmonic h is given as

$$(V_o)_h = \frac{1}{\sqrt{2}} \cdot 2 \cdot \frac{V_d}{2} \frac{(\hat{V}_{Ao})_h}{V_d/2} = \frac{V_d}{\sqrt{2}} \frac{(\hat{V}_{Ao})_h}{V_d/2}$$

$$= 212.13 \frac{(\hat{V}_{Ao})_h}{V_d/2} \tag{8-21}$$

Therefore, the rms voltages are as follows:

Fundamental: $V_{o1} = 212.13 \times 0.8 = 169.7\ V$ at 47 Hz
$(V_o)_{37} = 212.13 \times 0.22 = 46.67\ V$ at 1739 Hz
$(V_o)_{39} = 212.13 \times 0.818 = 173.52\ V$ at 1833 Hz
$(V_o)_{41} = 212.13 \times 0.22 = 46.67\ V$ at 1927 Hz
$(V_o)_{77} = 212.13 \times 0.314 = 66.60\ V$ at 3619 Hz
$(V_o)_{79} = 212.13 \times 0.314 = 66.60\ V$ at 3713 Hz
etc. ■

$$(V_{Ao})_h = \frac{1}{\sqrt{2}} \frac{V_d}{2} \frac{(\hat{V}_{Ao})_h}{V_d/2}$$

$$= 106.07 \frac{(\hat{V}_{Ao})_h}{V_d/2} \tag{8-11}$$

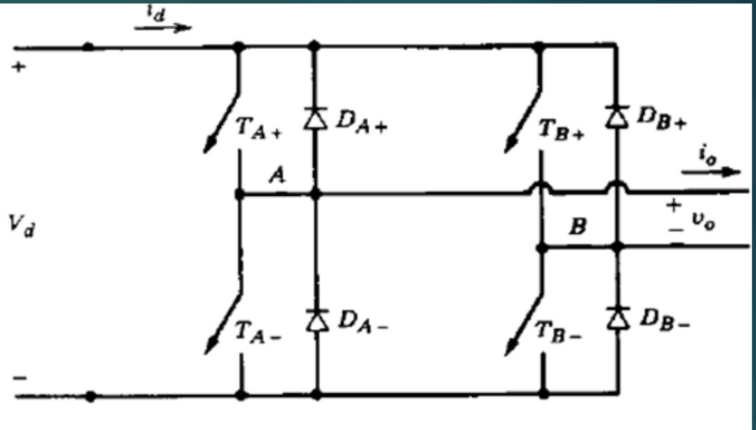


Figure 8-11 modified
Single-phase full-bridge inverter.

Table 8-1 Generalized Harmonics of v_{Ao} for a Large m_f.

h \ m_a	0.2	0.4	0.6	0.8	1.0
1 Fundamental	0.2	0.4	0.6	0.8	1.0
m_f	1.242	1.15	1.006	0.818	0.601
$m_f \pm 2$	0.016	0.061	0.131	0.220	0.318
$m_f \pm 4$					0.018
$2m_f \pm 1$	0.190	0.326	0.370	0.314	0.181
$2m_f \pm 3$		0.024	0.071	0.139	0.212
$2m_f \pm 5$				0.013	0.033
$3m_f$	0.335	0.123	0.083	0.171	0.113
$3m_f \pm 2$	0.044	0.139	0.203	0.176	0.062
$3m_f \pm 4$		0.012	0.047	0.104	0.157
$3m_f \pm 6$				0.016	0.044
$4m_f \pm 1$	0.163	0.157	0.008	0.105	0.068
$4m_f \pm 3$	0.012	0.070	0.132	0.115	0.009
$4m_f \pm 5$			0.034	0.084	0.119
$4m_f \pm 7$				0.017	0.050

Note: $(\hat{V}_{Ao})_h/\frac{1}{2}V_d$ [$= (\hat{V}_{AN})_h/\frac{1}{2}V_d$] is tabulated as a function of m_a.

Slide from ODL Week 1 Lecture – for the half bridge inverter

Table 8-1 Generalized Harmonics of v_{Ao} for a Large m_f.

h \ m_a	0.2	0.4	0.6	0.8	1.0
1 Fundamental	0.2	0.4	0.6	0.8	1.0
m_f	1.242	1.15	1.006	0.818	0.601
$m_f \pm 2$	0.016	0.061	0.131	0.220	0.318
$m_f \pm 4$					0.018
$2m_f \pm 1$	0.190	0.326	0.370	0.314	0.181
$2m_f \pm 3$		0.024	0.071	0.139	0.212
$2m_f \pm 5$				0.013	0.033
$3m_f$	0.335	0.123	0.083	0.171	0.113
$3m_f \pm 2$	0.044	0.139	0.203	0.176	0.062
$3m_f \pm 4$		0.012	0.047	0.104	0.157
$3m_f \pm 6$				0.016	0.044
$4m_f \pm 1$	0.163	0.157	0.008	0.105	0.068
$4m_f \pm 3$	0.012	0.070	0.132	0.115	0.009
$4m_f \pm 5$			0.034	0.084	0.119
$4m_f \pm 7$				0.017	0.050

Note: $(\hat{V}_{Ao})_h / \frac{1}{2}V_d\ [= (\hat{V}_{AN})_h / \frac{1}{2}V_d]$ is tabulated as a function of m_a.

■ ***Example 8-1*** In the circuit of Fig. 8-4, $V_d = 300$ V, $m_a = 0.8$, $m_f = 39$, and the fundamental frequency is 47 Hz. Calculate the rms values of the fundamental-frequency voltage and some of the dominant harmonics in v_{Ao} using Table 8-1.

Solution From Table 8-1, the rms voltage at any value of h is given as

$$(V_{Ao})_h = \frac{1}{\sqrt{2}} \frac{V_d}{2} \frac{(\hat{V}_{Ao})_h}{V_d/2}$$
$$= 106.07 \frac{(\hat{V}_{Ao})_h}{V_d/2} \tag{8-11}$$

Therefore, from Table 8-1 the rms voltages are as follows:

Fundamental: $(V_{Ao})_1 = 106.07 \times 0.8 = 84.86\ V$ at 47 Hz
$(V_{Ao})_{37} = 106.07 \times 0.22 = 23.33\ V$ at 1739 Hz
$(V_{Ao})_{39} = 106.07 \times 0.818 = 86.76\ V$ at 1833 Hz
$(V_{Ao})_{41} = 106.07 \times 0.22 = 23.33\ V$ at 1927 Hz
$(V_{Ao})_{77} = 106.07 \times 0.314 = 33.31\ V$ at 3619 Hz
$(V_{Ao})_{79} = 106.07 \times 0.314 = 33.31\ V$ at 3713 Hz
etc. ■

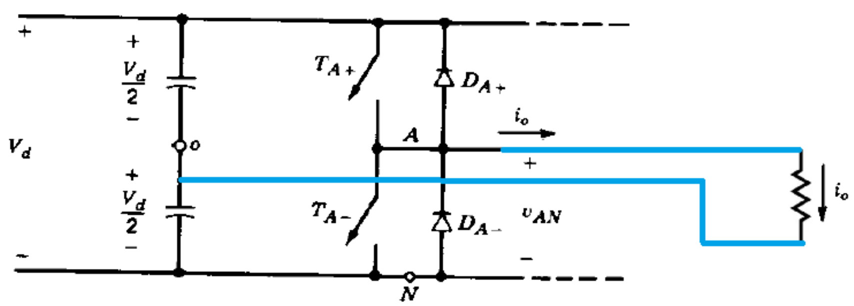


Figure 8-4 One-leg switch-mode inverter.

8-3-2-2 PWM with Unipolar Voltage Switching

In PWM with unipolar voltage switching, the switches in the two legs of the full-bridge inverter of Fig. 8-11 are not switched simultaneously, as in the previous PWM scheme. Here, the legs A and B of the full-bridge inverter are controlled separately by comparing v_{tri} with $v_{control}$ and $-v_{control}$, respectively. As shown in Fig. 8-15*a*, the comparison of $v_{control}$ with the triangular waveform results in the following logic signals to control the switches in leg A:

$$\begin{aligned} v_{control} > v_{tri}&: \quad T_{A+} \text{ on} \quad \text{and} \quad v_{AN} = V_d \\ v_{control} < v_{tri}&: \quad T_{A-} \text{ on} \quad \text{and} \quad v_{AN} = 0 \end{aligned} \tag{8-29}$$

The output voltage of inverter leg A with respect to the negative dc bus N is shown in Fig. 8-15*b*. For controlling the leg B switches, $-v_{control}$ is compared with the same triangular waveform, which yields the following:

$$\begin{aligned} (-v_{control}) > v_{tri}&: \quad T_{B+} \text{ on} \quad \text{and} \quad v_{BN} = V_d \\ (-v_{control}) < v_{tri}&: \quad T_{B-} \text{ on} \quad \text{and} \quad v_{BN} = 0 \end{aligned} \tag{8-30}$$

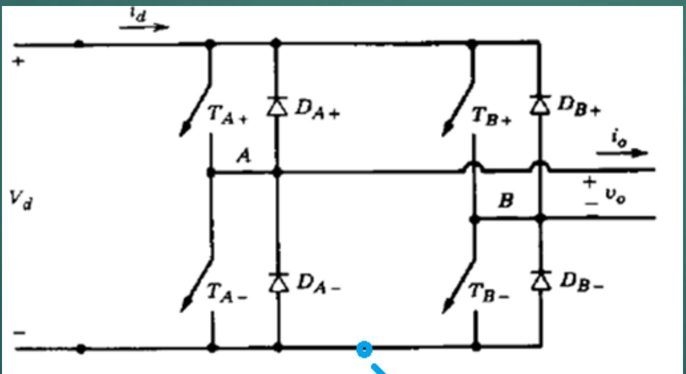


Figure 8-11 modified
Single-phase full-bridge inverter.

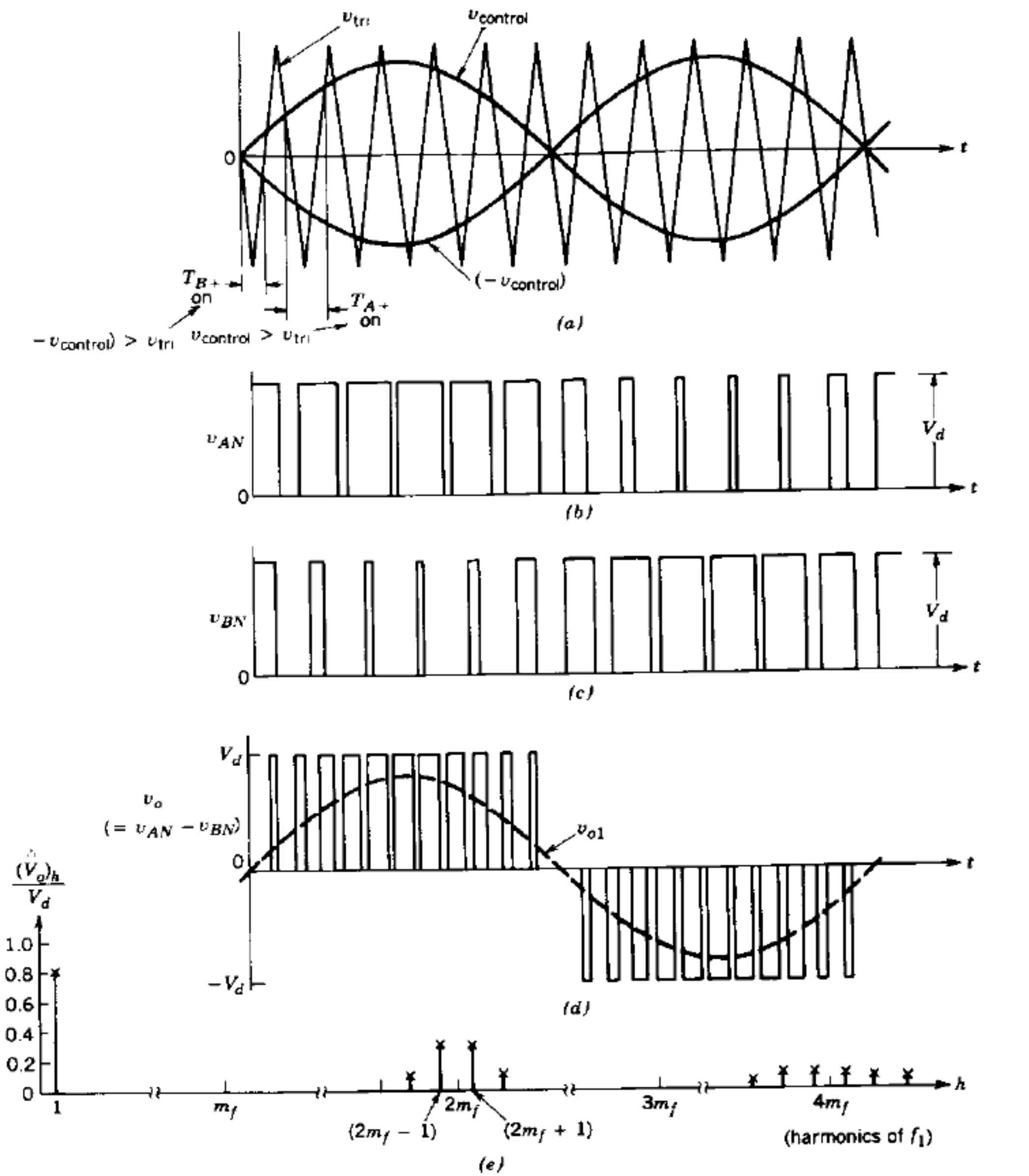


Figure 8-15 PWM with unipolar voltage switching (single phase).

This scheme has the advantage of "effectively" doubling the switching frequency as far as the output harmonics are concerned, compared to the bipolar voltage-switching scheme. Also, the voltage jumps in the output voltage at each switching are reduced to V_d, as compared to $2V_d$ in the previous scheme.

The advantage of "effectively" doubling the switching frequency appears in the harmonic spectrum of the output voltage waveform, where the lowest harmonics (in the idealized circuit) appear as sidebands of twice the switching frequency. It is easy to understand this if we choose the frequency modulation ratio m_f to be even (m_f should be odd for PWM with bipolar voltage switching) in a single-phase inverter. The voltage

■ ***Example 8-3*** In Example 8-2, suppose that a PWM with unipolar voltage-switching scheme is used, with $m_f = 38$. Calculate the rms values of the fundamental-frequency voltage and some of the dominant harmonics in the output voltage.

Solution Based on the discussion of unipolar voltage switching, the harmonic order h can be written as

$$h = j(2m_f) \pm k \tag{8-34}$$

where the harmonics exist as sidebands around $2m_f$ and the multiples of $2m_f$. Since h is odd, k in Eq. 8-34 attains only odd values. From Example 8-2

$$(V_o)_h = 212.13\frac{(V_{Ao})_h}{V_d/2} \tag{8-35}$$

Using Eq. 8-35 and Table 8-1, we find that the rms voltages are as follows:

At fundamental or 47 Hz: $V_{o1} = 0.8 \times 212.13 = 169.7\ V$

At $h = 2m_f - 1 = 75$ *or* 3525 *Hz*: $(V_o)_{75} = 0.314 \times 212.13 = 66.60\ V$

At $h = 2m_f + 1 = 77$ *or* 3619 *Hz*: $(V_o)_{77} = 0.314 \times 212.13 = 66.60\ V$

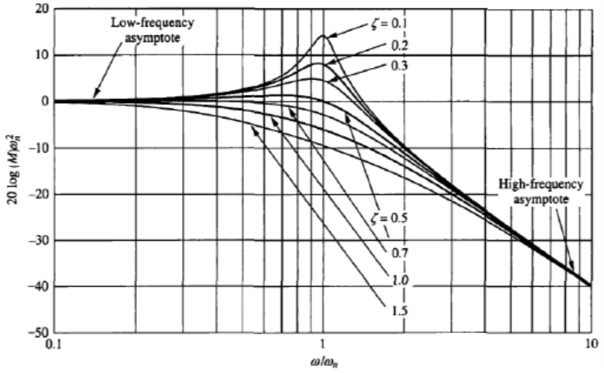
etc.

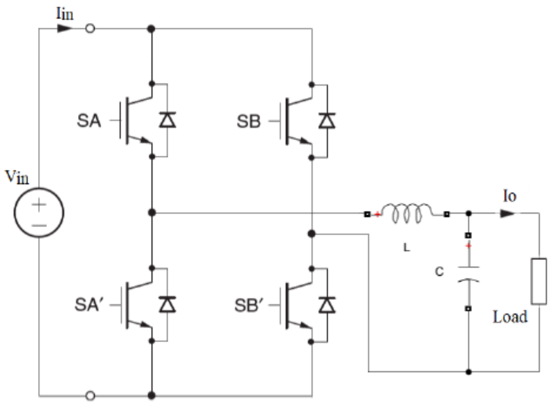
Comparison of the unipolar voltage switching with the bipolar voltage switching of Example 8-2 shows that, in both cases, the fundamental-frequency voltages are the same for equal m_a. However, with unipolar voltage switching, the dominant harmonic voltages centered around m_f disappear, thus resulting in a significantly lower harmonic content. ■

Design

Design a single phase full bridge inverter as shown the figure. The input voltage is 24 Volts. The output voltage is required to be 10V at 50Hz and the load is 100W at 0.8 p.f. Bipolar PWM will be employed in the inverter with a switching frequency of 1050 Hz. Ignore the losses/voltage drops in switches. Calculate the

a. the required voltage and current ratings for the IGBTs and the diodes (assume pure sine wave output)
b. Analyze the circuit to calculate the rms value of the 21st and 23rd harmonics before the filter.
c. Subsequently, design the LC filter to reduce these harmonics to less than or equal to 4% of the fundamental output voltage. You may use $\zeta = 0.1$ line for ease.
d. Re-analyze your design and calculate the output voltage by assuming 0.7V forward voltage drop of diode and 1 V for the IGBT.





Generalized Harmonics of v_o for a Large m_f.

h \ m_a	0.2	0.4	0.6	0.8	1.0
1 Fundamental	0.2	0.4	0.6	0.8	1.0
m_f	1.242	1.15	1.006	0.818	0.601
$m_f \pm 2$	0.016	0.061	0.131	0.220	0.318
$m_f \pm 4$					0.018

Note: $(\hat{V}_{Ao})_h / \frac{1}{2}V_d$ [$= (\hat{V}_{AN})_h / \frac{1}{2}V_d$] is tabulated as a function of m_a

Note: The table is for half-bridge inverter, multiply values by 2 for full bridge inverter.

Solution/Answers - Final Exam Power Electronics. 30/12/2019.

<u>Q # 1</u>

(a) $\hat{V}_{o1} = m_a V_d$ so $m_a = \hat{V}_{o1}/V_d$

$= \frac{10 \times \sqrt{2}}{24} \simeq 0.6$

$P_o = V_{o1} I_{o1\text{-rms}} \cos\theta$

so

$I_{o1\text{-rms}} = \frac{P_o}{V_{o1}\cos\theta} = \frac{100}{(10)(0.8)} = 12.5\text{ A}$

$I_{o1\text{-pk}} = 12.5 \times \sqrt{2} = 17.67\text{ A}$

<u>Ratings</u>

$I_{IGBT\text{-pk}} = I_{diode\text{-pk}} = 17.67\text{ A}$

$V_{IGBT\text{-pk}} = V_{diode\text{-pk}} = 24\text{ V}$

(b) $\frac{\hat{V}_{o\text{-}21}}{\frac{1}{2}V_d} = 1.006$ for $m_a = 0.6$ & half-bridge inv.

for full bridge

$\frac{\hat{V}_{o\text{-}21}}{\frac{1}{2}V_d} = 2.012$

so $\hat{V}_{o\text{-}21} = 24.14\text{ V}$

$V_{o\text{-}21\text{-rms}} = 17.07\text{V}$

<u>Similarly</u>

$\frac{\hat{V}_{o\text{-}23}}{\frac{1}{2}V_d} = (0.131)2$

& $V_{o\text{-}23\text{-rms}} = (0.131)(2)\left(\frac{1}{2}V_d\right)\left(\frac{1}{\sqrt{2}}\right) = 2.22\text{V}$

(c) Max. harmonic voltage = 4% of fundamental

$= (4/100)(10\text{V})$

$= 0.4\text{V rms}$

$V_{o21\text{-rms}}$ has to be reduced from 17.07V to 0.4V

i.e. attenuation of $\frac{0.4}{17.07} = 0.0234$

$= -32.6\text{ dB}$

from Bode plot, this occurs around $\frac{\omega}{\omega_n}\left(= \frac{f}{f_n}\right) = 6$

i.e. f_n should be $\leq \frac{1}{6}$ of f_{21}

i.e. $\leq \frac{1050}{6} = 175\text{ Hz}$

choosing $f_n = 175\text{ Hz}$

if $L = 1\text{mH}$ then from $f_n = \frac{1}{2\pi\sqrt{LC}}$

$C = \frac{(2\pi f_n)^2}{L} \approx 0.828\text{ mF}$

Subsequently for 23rd harmonic, attenuation will be more — & since its value is already small (2.2V), it will definitely be less than 0.4V after attenuation

(d) With 1V drop in IGBT, PWM voltage moves b/w +22 —— −22V

so,

$\hat{V}_{o1} = m_a V_d$

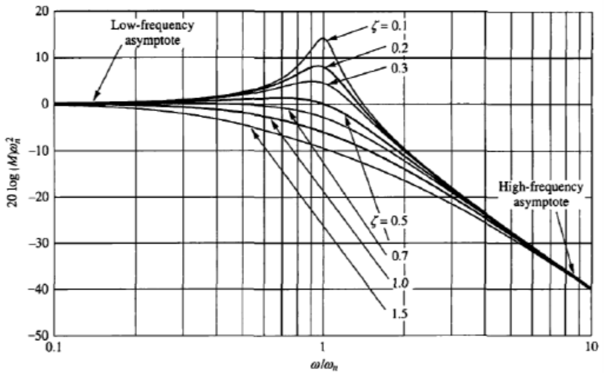
$= (0.6)(22) = 13.2\text{V}$

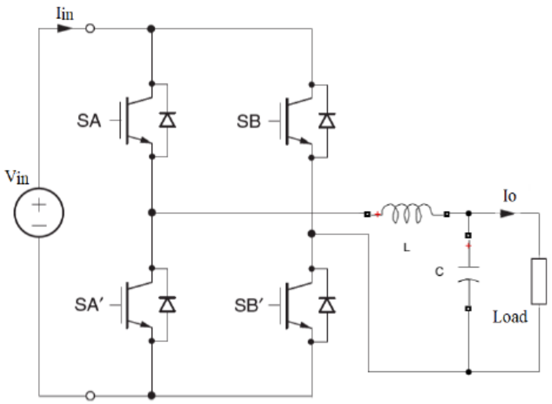
$V_{o1\text{-rms}} = 9.33\text{V}$

Design

Design a single phase full bridge inverter as shown the figure. The input voltage is 24 Volts. The output voltage is required to be 10V at 50Hz and the load is 100W at 0.8 p.f. Bipolar PWM will be employed in the inverter with a switching frequency of 1050 Hz. Ignore the losses/voltage drops in switches. Calculate the

a. the required voltage and current ratings for the IGBTs and the diodes (assume pure sine wave output)
b. Analyze the circuit to calculate the rms value of the 21st and 23rd harmonics before the filter.
c. Subsequently, design the LC filter to reduce these harmonics to less than or equal to 4% of the fundamental output voltage. You may use $\zeta = 0.1$ line for ease.
d. Re-analyze your design and calculate the output voltage by assuming 0.7V forward voltage drop of diode and 1 V for the IGBT.





Generalized Harmonics of v_o for a Large m_f.

h \ m_a	0.2	0.4	0.6	0.8	1.0
1 Fundamental	0.2	0.4	0.6	0.8	1.0
m_f	1.242	1.15	1.006	0.818	0.601
$m_f \pm 2$	0.016	0.061	0.131	0.220	0.318
$m_f \pm 4$					0.018

Note: $(\hat{V}_{Ao})_h / \frac{1}{2}V_d$ [$= (\hat{V}_{AN})_h / \frac{1}{2}V_d$] is tabulated as a function of m_a

Note: The table is for half-bridge inverter, multiply values by 2 for full bridge inverter.

Solution/Answers - Final Exam Power Electronics. 30/12/2019.

Q # 1

(a) $\hat{V}_{o1} = m_a V_d$ so $m_a = \hat{V}_{o1}/V_d$

$= \frac{10 \times \sqrt{2}}{24} \simeq 0.6$

$P_o = V_{o1} I_{o1\text{-rms}} \cos\theta$

so

$I_{o1\text{-rms}} = \frac{P_o}{V_{o1}\cos\theta} = \frac{100}{(10)(0.8)} = 12.5\,A$

$I_{o1\text{-pk}} = 12.5 \times \sqrt{2} = 17.67\,A$

Ratings

$I_{IGBT\text{-pk}} = I_{diode\text{-pk}} = 17.67\,A$

$V_{IGBT\text{-pk}} = V_{diode\text{-pk}} = 24\,V$

(b) $\frac{\hat{V}_{o\text{-}21}}{\frac{1}{2}V_d} = 1.006$ for $m_a = 0.6$ & half-bridge inv.

for full bridge

$\frac{\hat{V}_{o\text{-}21}}{\frac{1}{2}V_d} = 2.012$

so $\hat{V}_{o\text{-}21} = 24.14\,V$

$V_{o\text{-}21\text{-rms}} = 17.07\,V$

Similarly

$\frac{\hat{V}_{o\text{-}23}}{\frac{1}{2}V_d} = (0.131)2$

& $V_{o\text{-}23\text{-rms}} = (0.131)(2)\left(\frac{1}{2}V_d\right)\left(\frac{1}{\sqrt{2}}\right) = 2.22\,V$

(c) Max. harmonic voltage = 4% of fundamental

$= (4/100)(10\,V)$

$= 0.4\,V$ rms

$V_{o21\text{-rms}}$ has to be reduced from 17.07 V to 0.4 V

i.e. attenuation of $\frac{0.4}{17.07} = 0.0234$

$= -32.6$ dB

from Bode plot, this occurs around $\frac{\omega}{\omega_n} \left(= \frac{f}{f_n}\right) = 6$

i.e. f_n should be $\leqslant \frac{1}{6}$ of f_{21}

i.e. $\leqslant \frac{1050}{6} = 175\,Hz$

choosing $f_n = 175\,Hz$

if $L = 1\,mH$ then from $f_n = \frac{1}{2\pi\sqrt{LC}}$

$C = \frac{(2\pi f_n)^2}{L} \simeq 0.828\,mF$

Subsequently for 23rd harmonic, attenuation will be more — & since its value is already small (2.2V), it will definitely be less than 0.4V after attenuation

(d) With 1V drop in IGBT, PWM voltage moves b/w +22 —— -22V

so,

$\hat{V}_{o1} = m_a V_d$

$= (0.6)(22) = 13.2\,V$

$V_{o1\text{-rms}} = 9.33\,V$

CLO3 Improvement Quiz 30 Jan. 2020 **EE410 Power Electronics**

Time: 15 minutes, Marks=10 Session 2016 (7th Semester) Reg. No Soluti

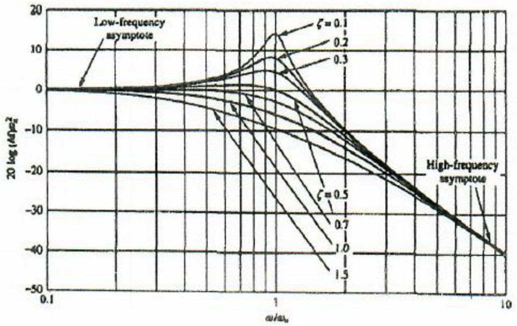
Design a single phase full bridge inverter as shown the figure. The input voltage is 20 Volts. The output voltage is required to be 10V at 50Hz and the load is 100W at 0.8 p.f. Bipolar PWM will be employed in the inverter with a switching frequency of 950 Hz. Ignore the losses/voltage drops in switches. Calculate the

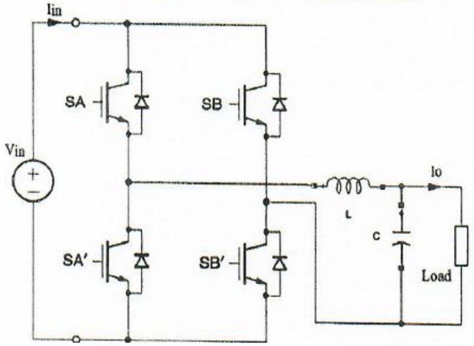
a. the required peak voltage and peak current ratings for the IGBTs and the diodes (assume pure sine wave output)

b. Analyze the circuit to calculate the rms value of the 19th[st] and 21[st] harmonics before the filter.

c. Subsequently, design the LC filter to reduce these harmonics to less than or equal to 3% of the fundamental output voltage. You may use ζ= 0.1 line for ease.

d. Mention the differences in the design of the LC filter if unipolar PWM is used.





Generalized Harmonics of v_o for a Large m_f.

h \ m_a	0.2	0.4	0.6	0.8	1.0
1 Fundamental	0.2	0.4	0.6	0.8	1.0
m_f	1.242	1.15	1.006	0.818	0.601
$m_f \pm 2$	0.016	0.061	0.131	0.220	0.318
$m_f \pm 4$					0.018

Note: $(\hat{V}_{Ao})_h / \frac{1}{2}V_d$ [$= (\hat{V}_{AN})_h / \frac{1}{2}V_d$] is tabulated as a function of m_a

Note: The table is for half-bridge inverter, multiply values by 2 for full bridge inverter.

(a)

$P_o = V_{o1} I_{o-1rms} \cos\theta$ so

$I_{o1rms} = P_o / (V_{o1} \cdot \cos\theta) = (100)/(10 \cdot 0.8)$

$= 12.5$ A

$I_{o1-pk} = 12.5 \times \sqrt{2} = 17.67$ A

Ratings

$I_{IGBT-pk} = I_{diode-pk} = 17.67$ A

$V_{IGBT-pk} = V_{diode-pk} = 20$ V

(2)

(b) $\hat{V}_{o1} = m_a V_d$ so $m_a = \hat{V}_{o1}/V_d$

$= \frac{10 \times \sqrt{2}}{20} \simeq 0.7$

$\frac{\hat{V}_{o-19}}{\frac{1}{2} V_d} = \left(\frac{1.006 + 0.818}{2}\right) * 2$

interpolation

for full bridge.

$\hat{V}_{o-19} = \frac{20}{2}(0.912)\, 2 = 18.24$ V

$V_{o-19-rms} = 12.9$ V

(2)

similarly,

$\hat{V}_{o-23} = \frac{20}{2}\left(\frac{0.131 + 0.22}{2}\right) * 2$

$V_{o-23-rms} = 2.48$ V

(c) Max. harmonics = 3% of 10 = 0.3 V rms

V_{o-19} rms has to be reduced to 0.3V

i.e attenuation $\geqslant 0.3/18.24 = 0.0164$

$= -35.67$

from Bode. plot, this is around $\frac{\omega}{\omega_n} = 8$

so f_n should be $\leqslant \frac{1}{8}$ of f_h -

i.e $\leqslant 118.75$ Hz

choose $f_n = 110$ Hz

then $110 = \frac{1}{2\pi\sqrt{LC}}$ if $L = 1$mH then

$C \simeq 2$ mF

21st harmonic will be suppressed more & since value is small, it will definittely be less than 3%.

(4)

(d) With unipolar PWM, the harmonics donot occur at m_f or $m_f \pm 2$, rather at $2m_f$ & around it - so these will be calculated as in part (c) & then LC design will follow as in part (d)

(2)

CHAPTER 7

Resonant Pulse Inverters

7.1 INTRODUCTION

The switching devices in converters with a pulse-width-modulation (PWM) control can be gated to synthesize the desired shape of the output voltage or current. However, the devices are turned on and off at the load current with a high *di*/*dt* value. The switches are subjected to a high-voltage stress, and the switching power loss of a device increases linearly with the switching frequency. The turn-on and turn-off loss could be a significant portion of the total power loss. The electromagnetic interference is also produced due to high *di*/*dt* and *dv*/*dt* in the converter waveforms.

The disadvantages of PWM control can be eliminated or minimized if the switching devices are turned "on" and "off" when the voltage across a device or its current becomes zero [1]. The voltage and current are forced to pass through zero crossing by creating an *LC*-resonant circuit, thereby called a *resonant pulse converter*. The resonant converters can be broadly classified into eight types:

Series resonant inverters
Parallel resonant inverters
Class E resonant converter
Class E resonant rectifier
Zero-voltage-switching (ZVS) resonant converters
Zero-current-switching (ZCS) resonant converters
Two-quadrant ZVS resonant converters
Resonant dc-link inverters

7.2 SERIES RESONANT INVERTERS

The series resonant inverters are based on resonant current oscillation. The resonating components and switching device are placed in series with the load to form an underdamped circuit. The current through the switching devices falls to zero due to the natural characteristics of the circuit. If the switching element is a thyristor, it is said to be self-commutated.

This type of inverter produces an approximately sinusoidal waveform at a high-output frequency, ranging from 200 to 100 kHz, and is commonly used in relatively fixed output applications (e.g., induction heating, sonar transmitter, fluorescent lighting, or ultrasonic generators). Due to the high-switching frequency, the size of resonating components is small.

There are various configurations of series resonant inverters, depending on the connections of the switching devices and load. The series inverters may be classified into two categories:

1. Series resonant inverters with unidirectional switches
2. Series resonant inverters with bidirectional switches

There are three types of series resonant inverters with unidirectional switches—basic, half-bridge, and full-bridge.

An application of Resonant converters (and Power Electronics)

Working Principle of an Electronic Ballast

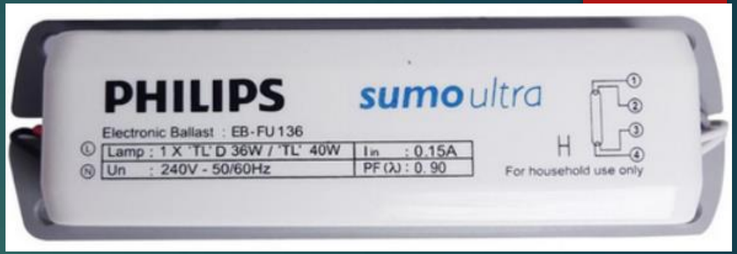
Electronic ballast takes supply at 50 – 60 Hz. It first converts AC voltage into DC voltage. After that, filtration of this DC voltage is done by using capacitor configuration. Now filtered DC voltage is fed to the high-frequency oscillation stage where oscillation is typically square wave and frequency range are from 20 kHz to 80 kHz.

$$L\frac{dI}{dt}$$

$$L\frac{dI}{dt}$$

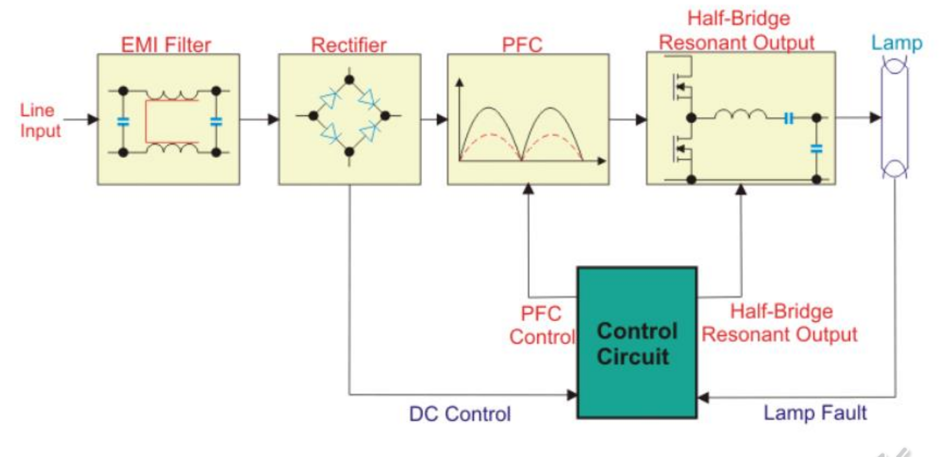
Hence output current is with very high frequency. A small amount of inductance is provided to be associated with a high rate of change of current on high frequency to generate high valued . Generally, more than 400 V is required to strike the gas discharge process in fluorescent tube light. When the switch is ON, the initial voltage across the lamp becomes 1000 V around due to high valued, hence gas discharge takes place instantaneously.

Once the discharge process is started, the voltage across the lamp is decreased below 230V up to 125V and then this electronic ballast allows limited current to flow through this lamp. This control of voltage and current is done by the control unit of the electronic ballast. In running condition of fluorescent lamp, electronic ballast acts as a dimmer to limit current and voltage.



Philips White 40 W Electronic Ballast Choke, Electrical Choke

Basic Circuitry of an Electronic Ballast



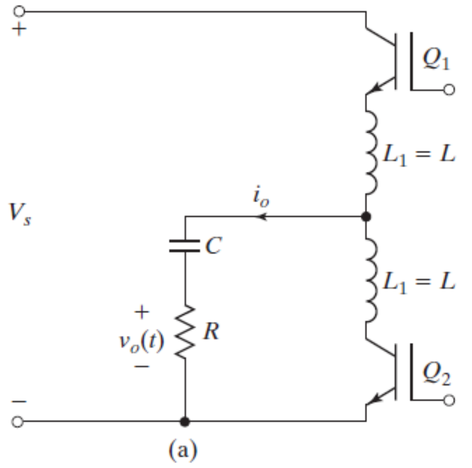
7.2.1 Series Resonant Inverters with Unidirectional Switches

Figure 7.1a shows the circuit diagram of a simple series inverter using two unidirectional transistor switches. When transistor Q_1 is turned on, a resonant pulse of current flows through the load and the current falls to zero at $t = t_{1m}$ and Q_1 is self-commutated.

Turned on transistor Q_2 causes a reverse resonant current through the load and Q_2 is also self-commutated. The circuit operation can be divided into three modes and the equivalent circuits are shown in Figure 7.1b.

The power flow from the dc supply is discontinuous. The dc supply has a high peak current and would contain harmonics.

The load current $i_1(t)$ must be zero and Q_1 must be turned off before Q_2 is turned on. Otherwise, a short-circuit condition results through the transistors and dc supply. Therefore, the available off-time t_{2m} ($= t_{off}$), known as the *dead zone*, must be greater than the turn-off time of transistors, t_{off}.



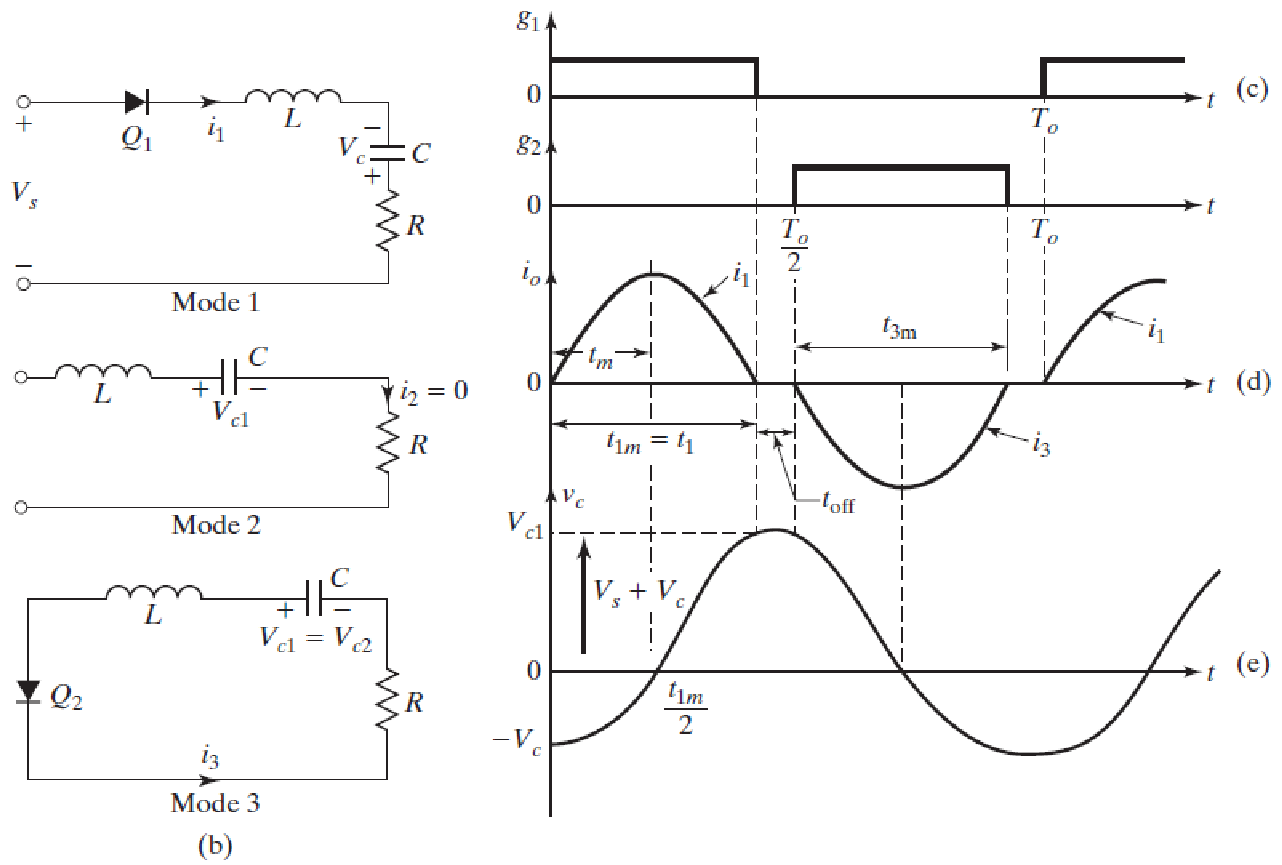


FIGURE 7.1

Basic series resonant inverter. (a) Circuit, (b) Equivalent circuits, (c) Gating signals, (d) Output current, and (e) Capacitor voltage.

7.2.1 Series Resonant Inverters with Unidirectional Switches

Figure 7.1a shows the circuit diagram of a simple series inverter using two unidirectional transistor switches. When transistor Q_1 is turned on, a resonant pulse of current flows through the load and the current falls to zero at $t = t_{1m}$ and Q_1 is self-commutated.

Turned on transistor Q_2 causes a reverse resonant current through the load and Q_2 is also self-commutated. The circuit operation can be divided into three modes and the equivalent circuits are shown in Figure 7.1b.

The power flow from the dc supply is discontinuous. The dc supply has a high peak current and would contain harmonics.

The load current $i_1(t)$ must be zero and Q_1 must be turned off before Q_2 is turned on. Otherwise, a short-circuit condition results through the transistors and dc supply. Therefore, the available off-time $t_{2m} (= t_{off})$, known as the *dead zone*, must be greater than the turn-off time of transistors, t_{off}.

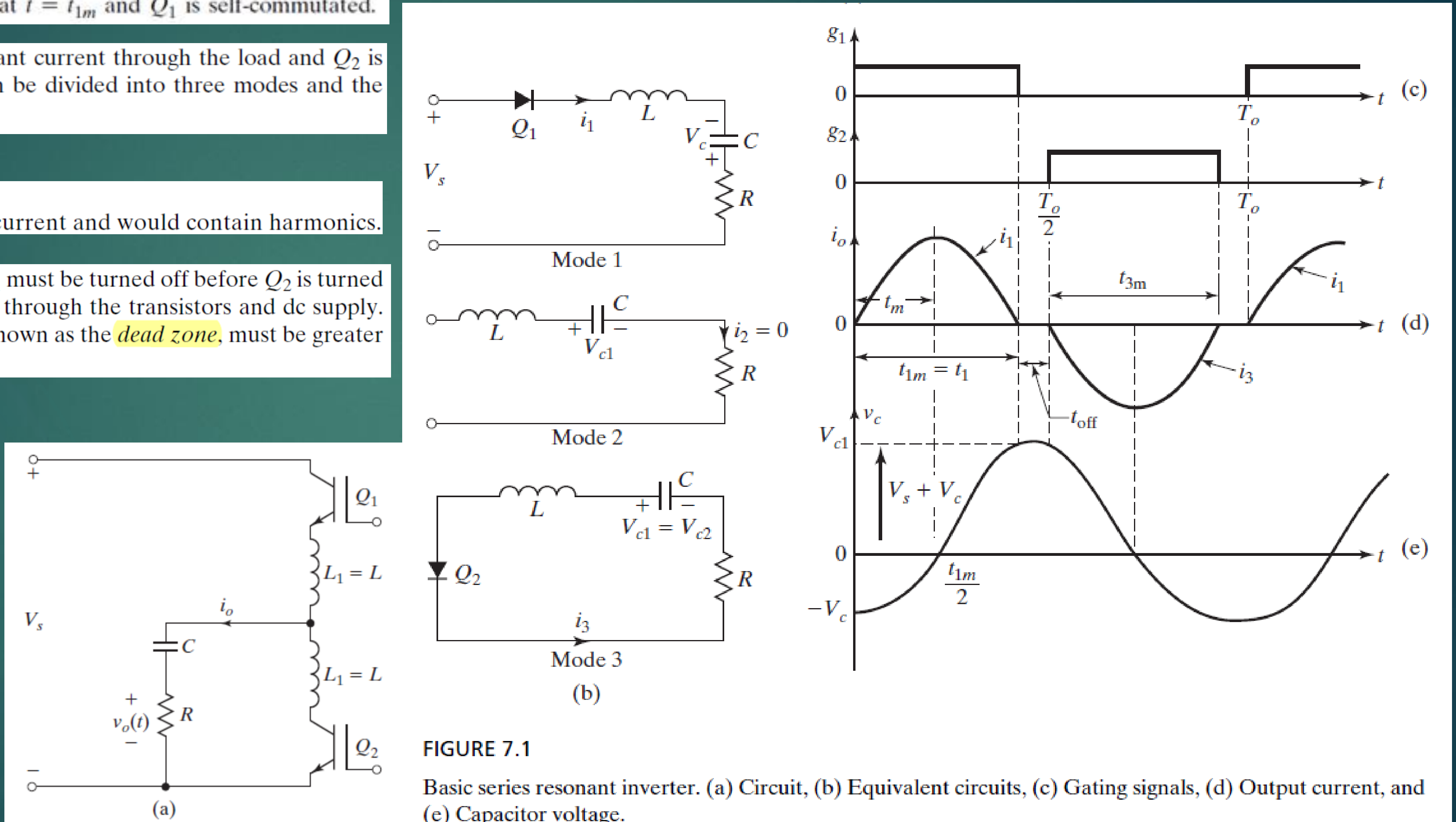


FIGURE 7.1

Basic series resonant inverter. (a) Circuit, (b) Equivalent circuits, (c) Gating signals, (d) Output current, and (e) Capacitor voltage.

Full Bridge Series Resonant Inverter

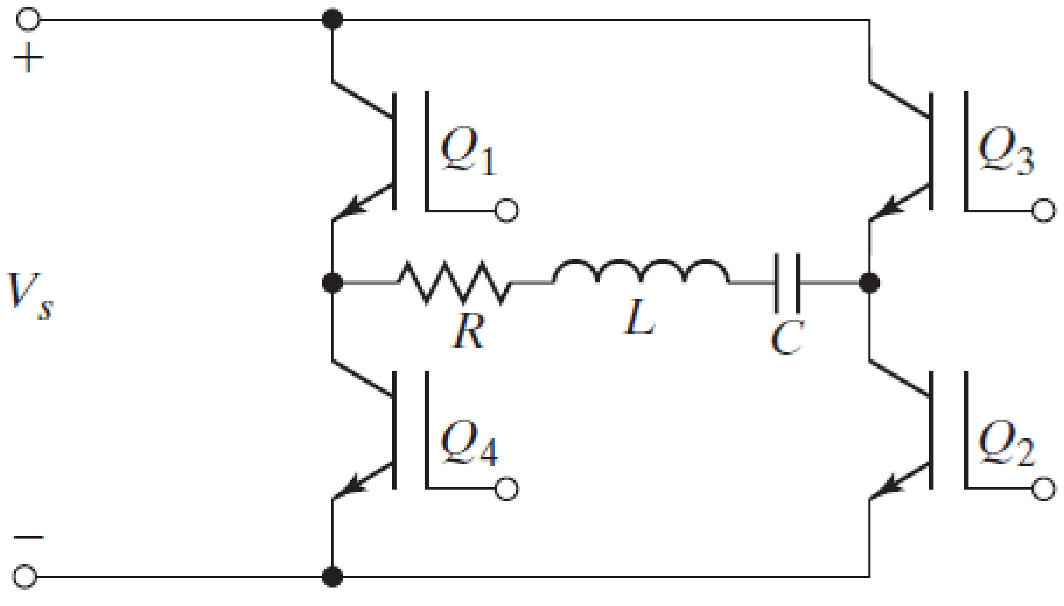


FIGURE 7.4

Full-bridge series resonant inverter.

9.4 THE SERIES RESONANT INVERTER

The series resonant inverter (dc-to-ac converter) of Fig. 9-3*a* is one application of resonant converters. In a series resonant inverter, an inductor and a capacitor are placed in series with a load resistor. The switches produce a square wave voltage, and the inductor-capacitor combination is selected such that the resonant frequency is the same as the switching frequency.

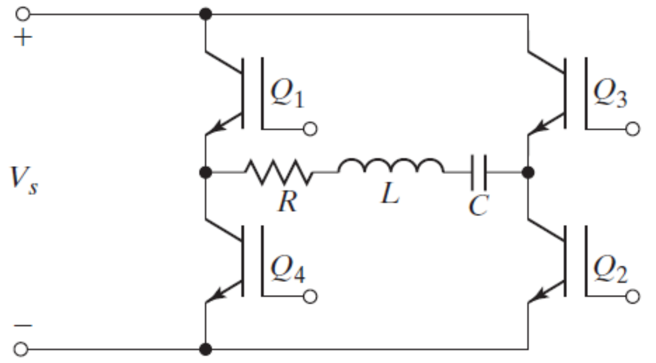


FIGURE 7.4

Full-bridge series resonant inverter.

$$V_{out} = \frac{R}{Z_{total}} * V_{in} = \frac{R}{\sqrt{R^2 + (X_L - X_C)^2}} * V_{in}$$

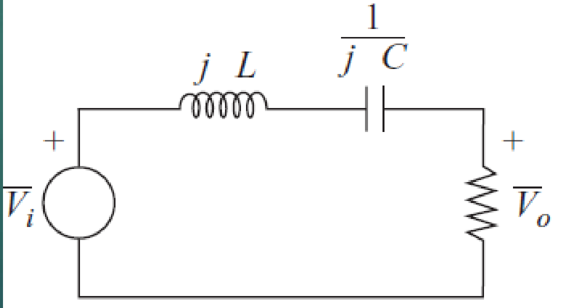
The input and output voltage amplitudes are related by

$$\frac{V_o}{V_i} = \frac{R}{\sqrt{R^2 + (\omega L - (1/\omega C))^2}} = \frac{1}{\sqrt{1 + ((\omega L/R) - (1/\omega RC))^2}} \quad (9\text{-}45)$$

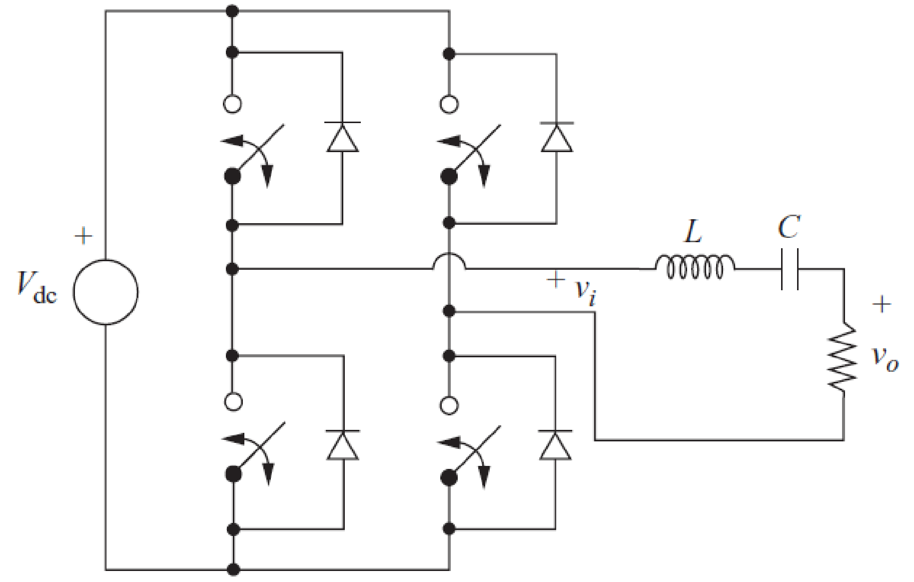
Resonance is at the frequency

$$\omega_0 = \frac{1}{\sqrt{LC}} \quad (9\text{-}46)$$

$$f_0 = \frac{1}{2\pi\sqrt{LC}} \quad (9\text{-}47)$$

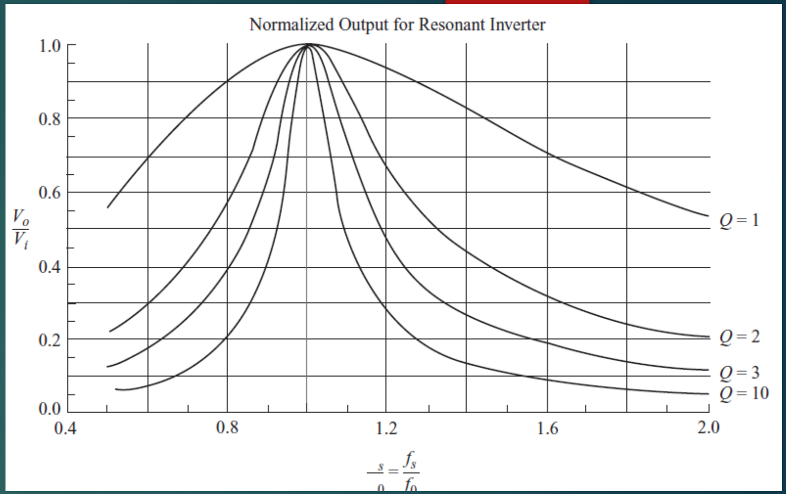


(*b*)



(*a*)

At resonance, the impedances of the inductance and capacitance cancel, and the load appears as a resistance. If the bridge output is a square wave at frequency f_0, the LC combination acts as a filter, passing the fundamental frequency and attenuating the harmonics. If the third and higher harmonics of the square wave bridge output are effectively removed, the voltage across the load resistor is essentially a sinusoid at the square wave's fundamental frequency.



The amplitude of the fundamental frequency of a square wave voltage of $\pm V_{dc}$ is

$$V_1 = \frac{4V_{dc}}{\pi} \tag{9-48}$$

$$\frac{V_o}{V_i} = \frac{R}{\sqrt{R^2 + (\omega L - (1/\omega C))^2}} = \frac{1}{\sqrt{1 + ((\omega L/R) - (1/\omega RC))^2}} \tag{9-45}$$

quality factor Q.

$$Q = \frac{\omega_0 L}{R} = \frac{1}{\omega_0 RC} \tag{9-49}$$

Equation (9-45) can be expressed in terms of ω_0 and Q:

$$\frac{V_o}{V_i} = \frac{1}{\sqrt{1 + Q^2((\omega/\omega_0) - (\omega_0/\omega))^2}} \tag{9-50}$$

The total harmonic distortion (THD, as defined in Chap. 2) of the voltage across the load resistor is reduced by increasing the Q of the filter. Increasing inductance and reducing capacitance increase Q.

$$\text{THD} = \sqrt{\frac{\sum_{n \neq 1} I_{n,\,\text{rms}}^2}{I_{1,\,\text{rms}}^2}} = \frac{\sqrt{\sum_{n \neq 1} I_{n,\,\text{rms}}^2}}{I_{1,\,\text{rms}}} = \sqrt{\frac{I_{\text{rms}}^2 - I_{1,\,\text{rms}}^2}{I_{1,\,\text{rms}}^2}}$$

EXAMPLE

A Resonant Inverter

A 10-Ω resistive load requires a 1000-Hz, 50-V rms sinusoidal voltage. The THD of the load voltage must be no more than 5 percent. An adjustable dc source is available. (*a*) Design an inverter for this application.

■ Solution

The full-bridge converter of Fig. 9-3*a* with 1000-Hz square wave switching and series resonant *LC* filter is selected for this design. The amplitude of a 50-V rms sinusoidal voltage is $\sqrt{2}(50) = 70.7$ V. The required dc input voltage is determined from Eq. (9-48).

$$70.7 = \frac{4V_{dc}}{\pi}$$

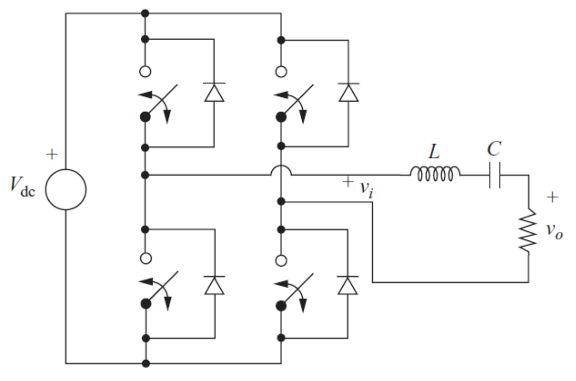
$$V_{dc} = 55.5 \text{ V}$$

The resonant frequency of the filter must be 1000 Hz, establishing the *LC* product. The *Q* of the filter and the THD limit are used to determine the values of *L* and *C*. The third harmonic of the square wave is the largest and will be the least attenuated by the filter. Estimating the THD from the third harmonic,

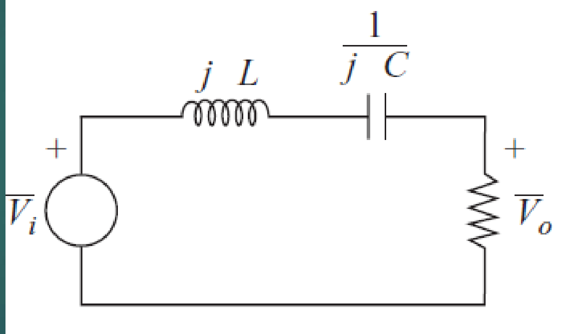
$$\text{THD} = \frac{\sqrt{\sum_{n \neq 1} V_n^2}}{V_1} \approx \frac{V_3}{V_1} \qquad (9\text{-}51)$$

where V_1 and V_3 are the amplitudes of the fundamental and third harmonic frequencies, respectively, across the load. Using the foregoing approximation, the amplitude of the third harmonic of the load voltage must be at most

$$V_3 < (\text{THD})(V_1) = (0.05)(70.7) = 3.54 \text{ V}$$



(*a*)



(*b*)

For the square wave, $V_3 = V_1/3 = 70.7/3$. Using Eq. (9-50), Q is determined from the magnitude of the third harmonic output with the third harmonic input, 70.7/3, at $\omega = 3\omega_0$.

$$\frac{V_{o,3}}{V_{i,3}} = \frac{3.54}{70.7/3} = \sqrt{\frac{1}{1 + Q^2((3\omega_0/\omega_0) - (\omega_0/3\omega_0))^2}}$$

Solving the preceding equation for Q results in $Q = 2.47$. Using Eq. (9-49),

$$L = \frac{QR}{\omega_0} = \frac{(2.47)(10)}{2\pi(1000)} = 3.93 \text{ mH}$$

$$C = \frac{1}{Q\omega_0 R} = \frac{1}{(2.47)(2\pi)(1000)(10)} = 6.44 \ \mu\text{F}$$

Power delivered to the load resistor at the fundamental frequency is $V_{\text{rms}}^2/R = 50^2/10 = 250$ W. Power delivered to the load at the third harmonic is $(2.5^2)/10 = 0.63$ W, showing that power at the harmonic frequencies is negligible.