ZCS를 이용한 고주파 공진형 상용주파수 전원에 관한 연구

論 文
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A Study on Commercial Frequency Source with High Frequency Resonant Type using ZCS

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Abstract-This paper describes a new dc-ac inverter system which for achieving sinusoidal ac waveform makes use of parallel loaded high frequency resonant inverter consisting of full bridge. Each one of the pair of switches in the inverter is driven to synchronous output frequency and the other is driven to PWM signal with resonant frequency proportional to magnitude of sine wave. A forced discontinuous conduction mode is used to realize the quasi-sinusoidal pulse in each switching period. Therefore the inverter generates sinusoidal modulated output voltage including carrier frequency that is resonant frequency. Carrier frequency components of modulated output voltage is filtered by low pass filter. Since current through switches is always zero at its turn-on in the proposed inverter, low stress and low switching loss is achieved. Operating characteristics of the proposed system is analyzed in per unit system using computer simulation. The output voltage of it includes low harmonics and it is almost close to sine wave. Also, the theoretical analysis is proved through the experimental test.

Keyword : Sinusoidal ac waveform, Discontinuous conduction mode, Quasi-sinusoidal pulse, Carrier frequency

1. Introduction

In recent years, the UPS(Uninterrupted Power Supply) is used as the power supply of electronic devices, communication equipment and medical treatment electronic systems. It is necessary for above apparatus(UPS) to have an inverter to convert DC to AC[1][2].

There are a lot of papers introducing how to control the inverter. The control methods they have been using are phase control and PWM methods till now[3][4]. The inverter controlled by these methods has some problems such as switching stress and loss, higher order harmonic components. Thus, we usually use ZVS and ZCS techniques to remove them [5][6][7].

Also, there is an inverter using high frequency resonance that includes VVVF technique. However, the inverter is too complicated because there are two resonant reactors and condensers, respectively[8].

In addition, the load assignment between arms in the inverter bridge circuit is unbalance. This paper proposes a simple resonant inverter that is similar to conventional full bridge Inverter. Thus, it is possible to reduce the number of switching devices, resonant inductors and condensers in the inverter circuit.

The proposed inverter operates as discontinuous conduction mode and makes the output voltage formed sinusoidal quasi-resonance pulse row by using a high resonant controlled PWM inverter.

The output voltage becomes a sinusoidal wave by LPF(Low Pass Filter). In addition, the switching stresses and losses are reduced because the switches are turned on only when the current through the inverter circuit is made zero. The theoretical analysis is proved through the experimental test.

2. Operating principles of the proposed inverter

Fig. 1 shows a high frequency resonant inverter connected in parallel with the load. This is the proposed inverter circuit which is similar to the conventional full bridge inverter. In Fig.1, the switches S1, S2, S3, S4 are self turn-off devices for PWM control. L and C are the resonant reactor and the resonant condenser, respectively. Lf and Cf are the filter reactor and the filter condenser, and R is the load resistor. Diodes D1, D2 are for composing high frequency resonant loop, and Diode D is

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for blocking a resonant current conducting backward. Fig. 2 shows on/off timing diagram of the switches S1, S2, S3, S4 in order to make the output voltage close to the sinusoidal wave.

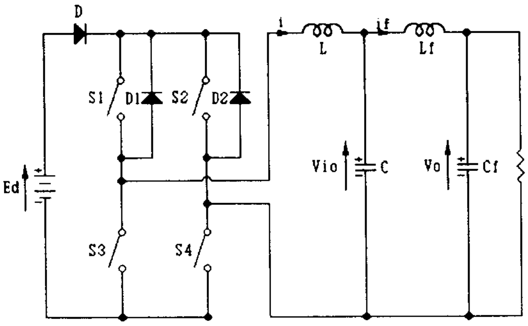


Fig. 1 Proposed inverter using ZCS

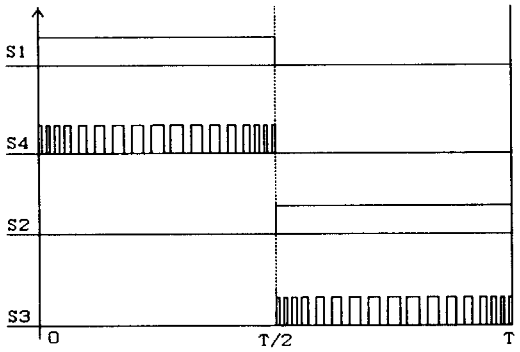


Fig. 2 on · off timing diagram

The proposed circuit shown in Fig. 1 consists of two switch family proportional to (S1, S4) and (S2, S3) as a full bridge type. The S1 and S2 are operated in turn to get the sinusoidal commercial frequency voltage and synchronize with the frequency of the source. The S3 and S4 are operated by the PWM signal in the frequency of the source.

Thus, if the operating frequency of the switches S1, S2 is changed, the frequency of the output voltage will be changed. Also, if the duty ratio of the switches S3, S4 is altered, the amplitude of the output voltage will be altered. Therefore, the proposed circuit can be operated as VVVF(Variable Voltage Variable Frequency).

In the case of not using the filters Lf and Cf, the operating mode, as like shown in Table 1, decided by L and C of the proposed circuit during one period can be divided into three modes. According to the states of the S1 and S2, each mode can be divided into two again.

These two modes have the same operation except that the voltage and current is in the opposite direction.

1 and 0, in table 1, represent the state of turn-on and off of the switching devices, respectively. Fig. 3 shows the equivalent circuit of the each mode.

To simplify the analysis of the circuit, the state equation is represented in per unit, as shown in Table 2. After this, all values are represented by the per-unit method.

Table 1 Operating Mode

Operating Mode	Mode 1		Mode 2		Mode 3	
	a	a′	b	b′	c	c′
S1	1	0	1	0	0	0
S2	0	1	0	1	0	0
S3	0	1	0	0	0	0
S4	1	0	0	0	0	0
D	1	1	0	0	0	0
D1	0	0	0	1	0	0
D2	0	0	1	0	0	0

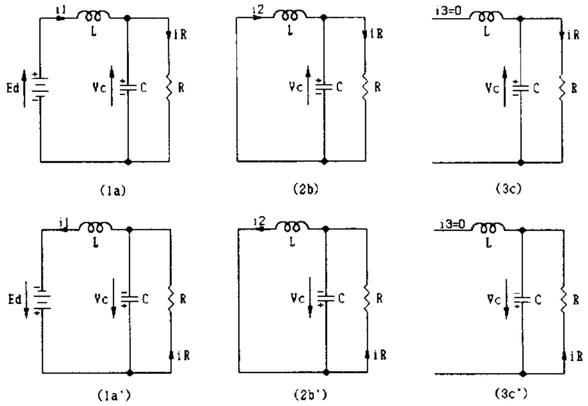


Fig. 3 Equivalent circuit of each mode

Table 2 Reference value of per unit system

Parameter	Reference value
Voltage	$V_b = E_d$
Current	$I_b = V_b/Z_b$
Reactor	$L_b = 44(\mu H)$
condenser	$C_b = 1.24(\mu F)$
Impedance	$Z_b = \sqrt{L_b/C_b}$
Time	$t_b = 2\pi\sqrt{L_b C_b}$
Frequency	$f_b = 1/t_b$
Angle Frequency	$\omega_b = 1/\sqrt{L_b C_b}$
Power	$P_b = V_b I_b$

The operation of the circuit at k times switching is as follows:

(1) Mode 1(a, a′)⇒$(t_o \sim t_1)$

In this mode, the condenser C is charged by the current i_1 flowing out from the source. The solutions of the current i_1 and the voltage v_{c1} can be obtained by the state equation as follows:

$$\begin{cases} i_1(t_k)=\dfrac{1}{R}+e^{-\alpha t_k}(k_{11}\sin\beta t_k+k_{12}\cos\beta t_k) \\ v_{c1}(t_k)=1+e^{-\alpha t_k}(k_{13}\sin\beta t_k+k_{14}\cos\beta t_k) \end{cases} \tag{1}$$

Where $\alpha=\dfrac{1}{2RC}$, $\beta=\sqrt{\omega^2-\alpha^2}$

$$k_{11}=\{\alpha k_{12}+(1-v_{01})/L\}/\beta,\quad k_{12}=i_{01}-\frac{1}{R}$$

$$k_{13}=\frac{i_{01}}{C}-\frac{\alpha(1-v_{01})}{\beta},\quad k_{14}=v_{01}-1,\quad t_k=t-t_0$$

The v_{01} equals $v_{c3}=(T_r-t_2)$ because the initial value of the v_{01} equals the final value of that. In the case of the mode 1(a), sn, sign of Ed, is 1. Under the mode 1(a′), sn is -1.

(2) Mode 2(b, b′)⇒(t_1~t_2)

In this mode, the source is disconnected, and the energy of the inductor is discharged.

The solutions of the current i_2 and the voltage v_{c2} can be obtained by the state equation as follows:

$$\begin{cases} i_2(t_k)=e^{-\alpha t_k}(k_{21}\sin\beta t_k+k_{22}\cos\beta t_k) \\ v_{c2}(t_k)=e^{-\alpha t_k}(k_{23}\sin\beta t_k+k_{24}\cos\beta t_k) \end{cases} \tag{2}$$

Where, $k_{21}=\{\alpha i_{02}-\dfrac{v_{02}}{L}\}/\beta$, $k_{22}=i_{02}$

$$k_{23}=\{\frac{i_{02}}{C}-\alpha v_{02}\}/\beta,\quad k_{24}=v_{02},\quad t_k=t-t_1$$

The i_{02} and v_{02} are the initial values of the mode 2. Thus, $i_{02}=i_1(t_1-t_0)$, $v_{02}=v_{c1}(t_1-t_0)$.

(3) Mode 3(c, c′)⇒(t_2~T_r)

In this mode, the current i_3 flowing through the inductor L is decreased to zero, the condenser C starts to discharge through the load. The solutions of the current i_3 and the voltage v_{c3} can be found by the state equation as follows:

$$\begin{cases} i_3(t_k)=0 \\ v_{c3}(t_k)=v_{03}e^{-2\alpha t_k} \end{cases} \tag{3}$$

Where, $t_k=t-t_2$, $v_{03}=v_{c2}(t_2-t_1)$.

The waveforms of the voltage and the current of the mode 3 during the k times a resonant period Tr are shown in Fig. 4.

When the switch S1(S2) is turned on at t = t_0, the current flowing through the resonant condenser and the load from the source starts to conduct and the energy of the inductor is stored.

The proposed system is using the high frequency resonant inverter operating as discontinuous conduction mode. The duty ratio of S3 and S4 has to be less than 1/2 in order to keep a time interval that is 50(%) of a switching period.

Thus, under the condition of the duty ratio, the switch is turned off at t = t_1, the stored energy of the inductor begins to discharge and the resonant current reduces and becomes zero at t = t_2.

The current of the inductor doesn't conduct until next signal and the charged energy of the condenser starts to discharge through the load.

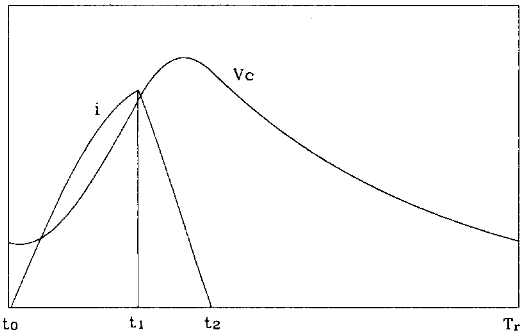


Fig. 4 Basic waveforms

3. Operating characteristics of the proposed inverter

To make the output voltage formed sinusoidal quasi-resonance pulse row like that shown in Fig. 4, the turn-on switching signal width t_{on} should be in proportion to the magnitude of the sinusoidal wave like shown by Eq (4).

$$t_{on}=\frac{1}{2}m\sin(9.28\times10^{-5}\pi ft) \tag{4}$$

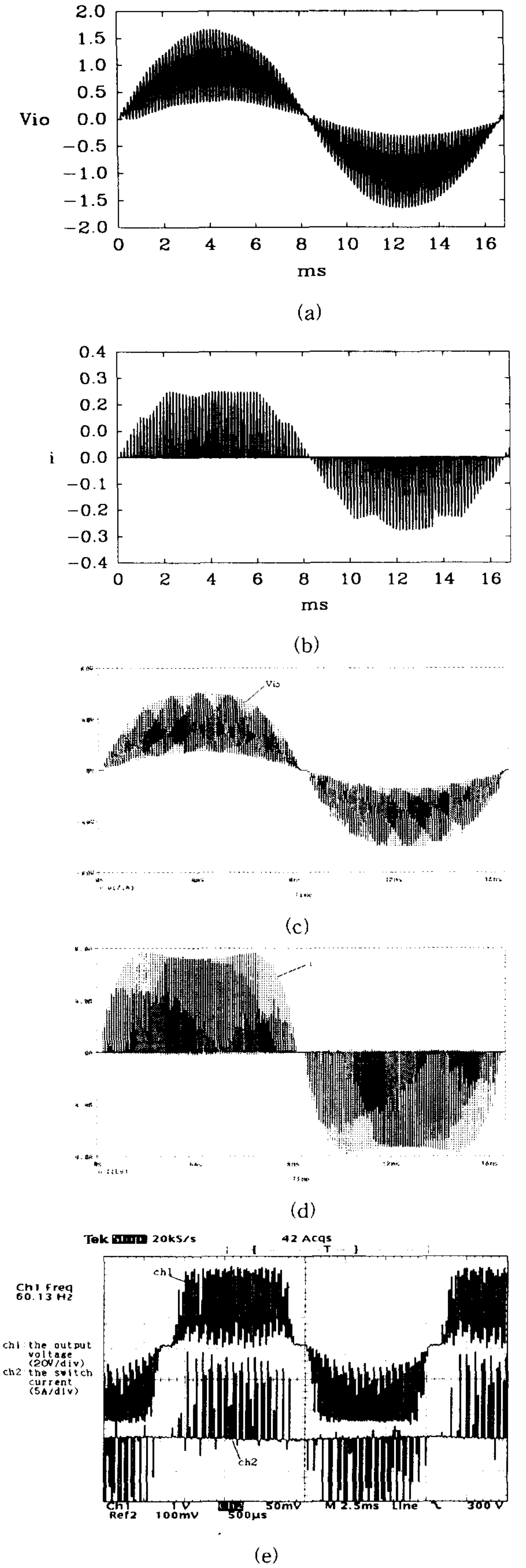
Where m is the amplitude modulation ratio.

According to the Eq (4), in the absence of the filters, the theoretical, the Pspice and the experimental waveforms of the output voltage and switch current are shown in Fig. 5(a), (b), (c), (d), (e) respectively.

Table 3. shows the circuit parameters of Fig. 5(c), (d), (e)

Table 3 The circuit parameters

Circuit Parameter	Rating	Circuit Parameter	Rating
Input Voltage	50(V)	Resonant reactor	44(μH)
Resonant condenser	1.24(μF)	Filter reactor	3(mH)
Filter condenser	20(μF)	Load Resistor	20(Ω)



(a)

(b)

(c)

(d)

(e)

Fig. 5 the output voltage(v_{io}) and switch current(i) of the proposed inverter(without filters)

The sinusoidal wave of the commercial frequency is modulated according to the carrier frequency i.e, the resonant frequency, the waveform of the output voltage being shown in Fig. 5.

Thus, as shown from Fig.1, since the carrier frequency components are included in the waveform of the output voltage, it can be reduced by adding the LPF composed of L_f and C_f, yielding finally, the sinusoidal wave of the output voltage.

The state and the output equation including the filter condenser resistor r_c, and the inductor resistor r_l, is given by

$$\begin{cases} \dfrac{dx_1}{dt} = A_1 x_1 + B_1 u_1 \\ y_1 = C_1 x_1 \end{cases} \qquad (5)$$

Where, $x_1 = [\,i_f \;\; v_{cf}\,]'$, $y_1 = v_0$, $u_1 = v_{io}$

$$A_1 = \begin{bmatrix} -\dfrac{r_1 r_c + R(r_l + r_c)}{L_f(R+r_c)} & -\dfrac{R}{L_f(R+r_c)} \\ \dfrac{R}{C_f(R+r_c)} & -\dfrac{1}{C_f(R+r_c)} \end{bmatrix}$$

$$B_1 = \left[\frac{1}{L_f} \;\; 0\right]', \quad C_1 = \left[\frac{r_c R}{R+r_c} \;\; \frac{R}{R+r_c}\right]$$

The transfer function of the output voltage v_o with respect to the filter input voltage v_{io} is given by

$$G(s) = \frac{v_o(s)}{v_{io}(s)} = \frac{a_o + a_1 s}{b_o + b_1 s + b_2 s^2} \qquad (6)$$

Where, $a_o = R$, $a_l = r_c R C_f$, $b_o = r_l + R$
$b_l = L_f + r_l R C_f + r_c r_l C_f + r_c R C_f$
$b_2 = L_f C_f (R + r_c)$

The inductor and the condenser using low pass filter circuit should be chosen by the current flowing through the condenser when no load current is less than 5(%) of the rated value. The cut-off frequency is decided by the range of controlled frequency. In this paper, in the case of R = 1.5, the cut-off frequency f_c is 0.04. Thus, the filters, L_f and C_f are 70 and 16, respectively. The frequency response of the transfer function of the equation (6) is shown in Fig. 6.

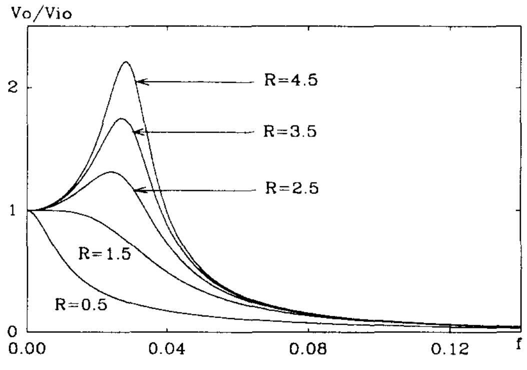
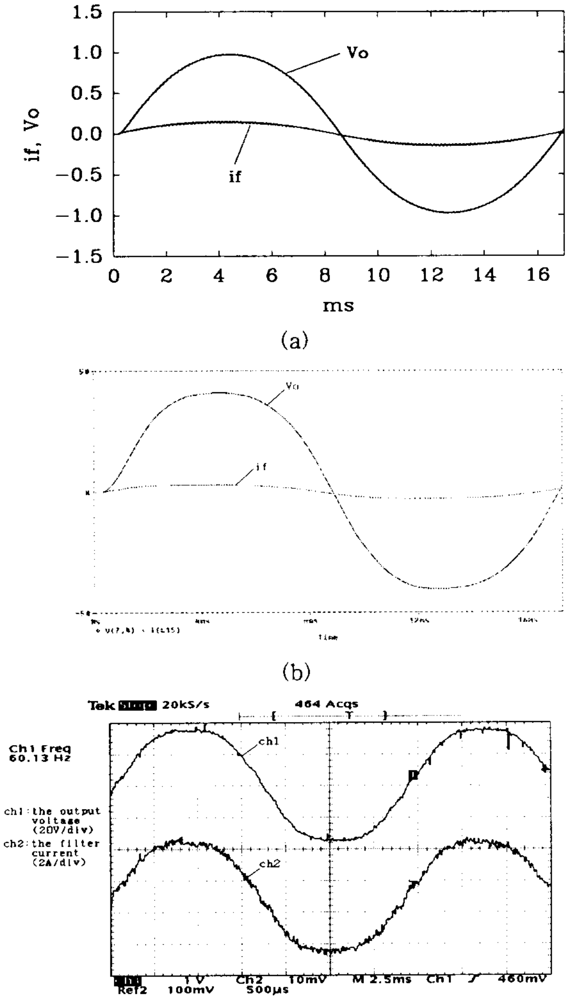


Fig. 6 Frequency characteristics of the transfer function.



(a)

(b)

(c)

Fig. 7 The output voltage(v_o) and filter inductor current(i_f) of the proposed inverter(with filters)

The state equations and the output equation including the filter are given by

$$\left\{\begin{aligned} \frac{dx}{dt} &= Ax+Bu \\ y &= Cx \end{aligned}\right. \tag{7}$$

Where, $x = [\, i_l \quad i_f \quad v_c \quad v_{cf}]^t$, $u = 1$

$$A = \begin{bmatrix} 0 & 0 & -\frac{m_1}{L} & 0 \\ 0 & 0 & \frac{1}{L_f} & -\frac{1}{L_f} \\ \frac{m_2}{C} & -\frac{1}{C} & 0 & 0 \\ 0 & \frac{1}{C_f} & 0 & -\frac{1}{RC_f} \end{bmatrix}$$

, $y = v_o$

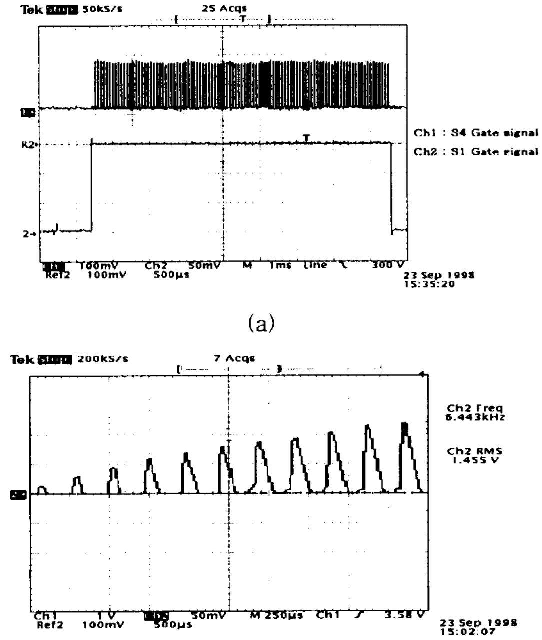
$$B = \begin{bmatrix} \frac{m_1 sn}{L} & 0 & 0 & 0 \end{bmatrix}^t, \quad C = [0 \quad 0 \quad 0 \quad 1]$$

Where ml and m2 are 1 in the mode 1 and 2, and 0 in the mode 3. sn is 1 in the mode a, b, c, and -1 in the mode a′, b′, c′.

The solution of the equation (7) can be obtained by the numerical analysis method.

Fig. 7(a), (b), (c) shows the output voltage and filter inductor current waveforms that are obtained by the simulation using the Runge-Kutta method, the Pspice and experiment, respectively.

The circuit parameters have already been represented in Fig. 5(c), (d), (e)



(a)

(b)

Fig. 8 Gate signals and switch current.

Fig. 8(a), (b) shows the gate signals of the switch S4 , S1 and the switch current(i).

As shown from Fig. 8(a), (b), we can know that the ZCS operation is realized by the gate signals of the switch S4, S1.

In order to analyze the harmonic components, the harmonic factor is essential and defined by

$$H.F = \frac{harmonic\ \ component}{fundamental\ \ component} = \frac{\sum_{h=2}^{\infty} V_h}{V_1}$$
$$= \sqrt{\frac{V_r^2 - V_d^2}{V_1} - 1} \tag{8}$$

Fig. 9 shows the rms values(v_r) and the harmonic factors with respect to the amplitude modulation ratios. In Fig. 9, the amplitude ratio is in proportion to the rms value and harmonic factor. Thus, when the amplitude modulation ratio is increased, the harmonic factor is also increased. Accordingly, the maximum output power in the permitted harmonic factor can be decided by Fig. 9.

According to the change of the load resistance, the rms voltage of the output and the harmonic factor are altered and shown in Fig. 10.

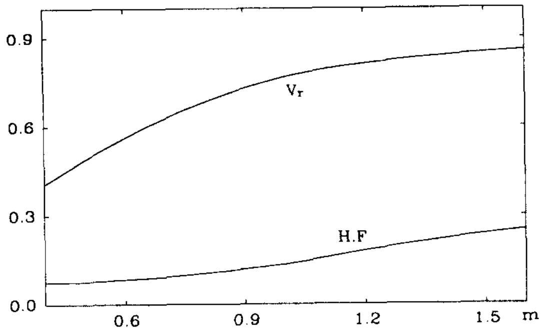


Fig. 9 The rms voltage and harmonic factor with amplitude modulation ratio m.

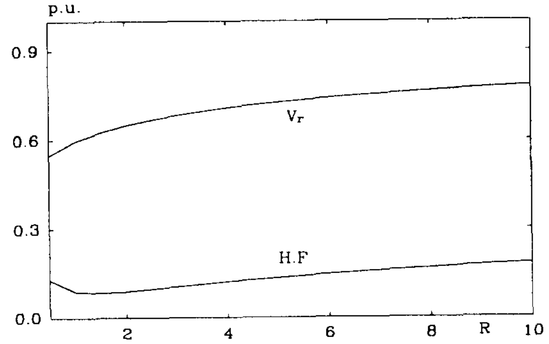


Fig. 10 The rms value and harmonic factor of the output voltage with load resistance.

The load resistance which is proportional to the output voltage is shown in Fig. 10. In order to keep the output voltage constant we need a proper control method.

4. Control of the proposed inverter system

In Fig. 9, since the output voltage is changed with the amplitude modulation ratio m, m is decided to keep the output voltage constant using the PI controller to reduce the error between the output voltage and the reference one. The m is as follows:

$$m = k_p(v_{ref} - v_o) + k_i \int (v_{ref} - v_o)dt \tag{9}$$

Where $v_{ref} = v_{rm}\sin(9.28 \times 10^{-5}\pi f t)$

The total control system to control the inverter system proposed in this paper is shown in Fig. 11.

The signal of the square wave of the output frequency can be got by a signal generator of the square wave or the commercial frequency changed to the square wave using the zero crossing detector. This square wave is converted to the pulse signal several times multiplied by the resonant frequency.

The m is decided to keep the output voltage constant using the PI controller to reduce the error between the detected output voltage and the reference one, and the signal of PWM is synchronized with the pulse signal of the frequency multiplier and generated by Eq. (4).

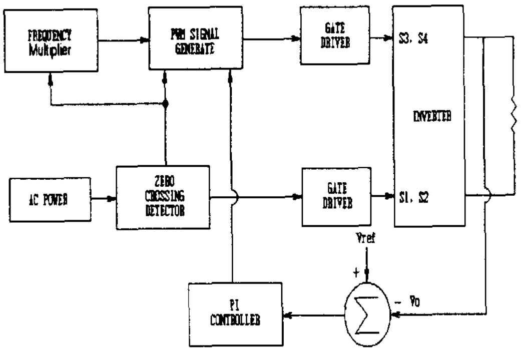


Fig. 11 Control system diagram with ZCS type a high frequency resonant inverter.

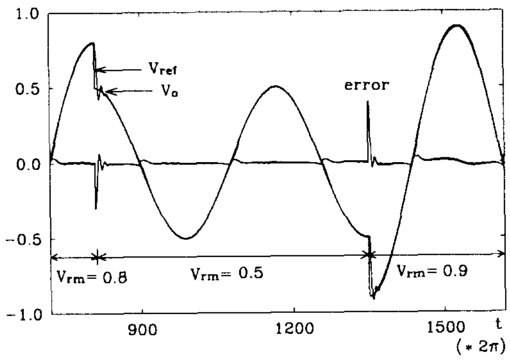


Fig. 12 Dynamic characteristics of the output voltage with varying reference voltage.

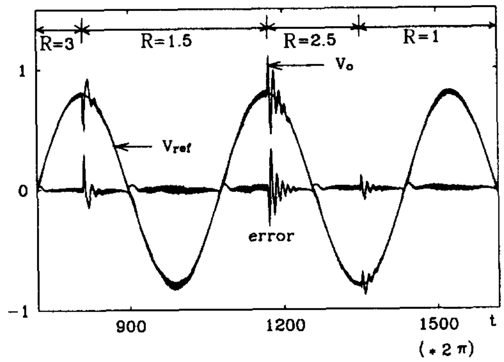


Fig. 13 Dynamic characteristic of the output voltage with varying load.

The dynamic characteristics of the output voltage controlled by this system are shown in the Fig. 12 and 13. Even though the reference voltage and the load are changed, the error is very small, and the output voltage follows the reference voltage well. Thus we can conclude that the proposed inverter has a good response.

5. Conclusion

This paper proposed the resonant inverter system that could control the amplitude and the frequency of the output voltage using PWM and lessen the problems such as switching stress, loss, and harmonics, of conventional high frequency inverters.

As the theoretical waveforms compare to the experimental waveforms, we could certify the operating characteristics of the proposed system.

The results is as follows ;

1) The proposed system is simple because the numbers of resonant condensers, reactors and switching devices are reduced.
2) The current stress and loss of the switching device is decreased because the proposed system is composed of a circuit using the high frequency resonant characteristics.
3) In the proposed system we can get the sinusoidal wave of the output because it includes only a few harmonics.
4) The response characteristics of the system are fast because a PI controller controls the error between the output voltage and the reference voltage.

In the future, we expect that this inverter can be used as the UPS of communication equipment and rectifier, induction appliance etc.

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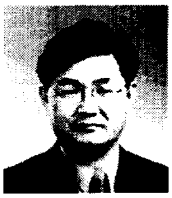

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