Investigation into the Effect of Switching Frequency on the Performance of Current Controlled Inverter Employing Space Vector Modulation Technique

By

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Abstract

In this paper the effect of switching frequency on the performance of the space vector modulation (SVM) current controlled three-phase PWM inverter loaded by interior permanent magnet synchronous motor (IPMSM), is investigated. The effects of the switching frequency on the stator space reference voltage vector $V_s(t)$ and then on the inverter operation modes are investigated. The idea of event driven indicative flag that on-line indicates the inverter mode of operation is presented. Computer simulation to compare the SVM performance at different switching frequencies is developed. Effects of changing the inverter dc link voltage, reference speed and mechanical load torque on the inverter modes of operation have been presented.

1- Introduction

Recent developments in high switching frequencies power devices have enabled us to develop ultrasonic carrier PWM control techniques. These high carrier frequencies have enabled us to improve the overall performance, especially for applications that require both quick response and accurate control. Mainly, current controller forces the load current to follow the command current in the same power apparatus \cite{1}. These devices may be AC motors or uninterruptible power supplies (UPS's). Current controller techniques have become an intensive subject of research due to the offered advantages for improving the dynamic performance of high-performance AC drives, applying vector control \cite{2,4}.

In conventional techniques of PWM such as regular sampling and subharmonics techniques the desired PWM switching points are determined according to the intersection points between the modulating wave and a triangular carrier wave at a certain switching frequency. The direct meaning of such arrangements is that the pulse position within the switching period is not controllable \cite{2,3}. The null time is the period at which the upper three switches of inverter or the lower three switches are active and as consequence the inverter output voltage is zero. These null times are not controllable in conventional techniques It has been proved that the null times or non-conducting times ratio at pulse beginning and pulse end have a significant effect on the current harmonics \cite{9}.

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Another disadvantage of the conventional techniques is the computational effort required for each one of the three phases, where the switching periods during one carrier period are usually calculated independently for each phase. This adds more burdens to any microprocessor-based system. These disadvantages have led to the development of SVM technique.

Space vector modulation (SVM) current controller is becoming more popular for providing PWM control for high-performance inverter-fed AC drives. SVM is gaining substantial attention as a promising field of research. Many researches have been executed to establish a relationship between regular-sampled PWM and space vector modulation (SVM) techniques and the effect of the proportion of null states (inverter zero output voltage) [3, 7, 8]. Other researches have focused on developing a theoretical criterion for the calculating the lower order harmonics of the current wave [5, 7, 8]. Others have paid special attention to the stator space voltage vector ($V_s$) and PWM voltage wave and on developing voltage compensation techniques for $V_s$ when exceeding the inverter voltage limits [6, 9].

This paper is focusing on the demonstration of three-phase PWM inverter different modes of operation according to the approach developed in [6, 9]. It is also focusing on the demonstration of the effect of the switching frequency ($f_s$) on the inverter modes of operation and on the stator space voltage vector ($V_s$). Also developing statistical relationship between $f_s$ and the inverter modes of operation has been investigated.

Interior permanent magnet synchronous motor (IPMSM) has many advantages, such as higher efficiency, owing to the absence of rotor copper losses, lower no-load current and much less sensitive to the motor parameter variations. Thus the IPMSM drive plays vitally important role in the motion-control applications in the range of low to medium power. To effectively control IPMSM, SVM technique is applied as a current controller on the inverter.

2- Concept of voltage space vector

For a system of balanced three-phase voltage $V_a(t)$, $V_b(t)$ and $V_c(t)$ of cycle period $2\pi$. The space voltage vector $V_s(t)$ can be expressed as

$$V_s(t) = (V_{\alpha}(t) + aV_{\beta}(t) + a^2V_{\gamma}(t))$$  

Where:  

$$a = e^{j2\pi t}$$

This definition gives a practical and friendly tool to handle three-phase quantities due to the reduction of notation complexity. This concept is the basis of SVM technique [6].

3- Inverter Mathematical Model

Figure (1) represents the switching logic of the inverter. Every two IGBTs on one arm are represented as one switch. The six IGBTs of inverter circuit are represented as three switches $N_\alpha$, $N_\beta$, and $N_\gamma$.  

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If one of the two switches on the same arm is one, the upper IGBT is conducting. While if it is zero, the lower one is conducting. The relationship between the three switches and the inverter output voltages are described according to the following equations [1].

\[
V_{a\,\text{PWM}} = \frac{V_{dc}}{3.0} (2N_a - N_b - N_c) \quad (2)
\]

\[
V_{b\,\text{PWM}} = \frac{V_{dc}}{3.0} (2N_b - N_a - N_c) \quad (3)
\]

\[
V_{c\,\text{PWM}} = \frac{V_{dc}}{3.0} (2N_c - N_b - N_a) \quad (4)
\]

Where:
- \(V_{a\,\text{PWM}}, V_{b\,\text{PWM}}\) and \(V_{c\,\text{PWM}}\) PWM actual output voltage of phase \(a, b\) and \(c\) respectively.

![Fig. (1) Equivalent circuit of voltage source inverter.](image1)

![Fig. (2) Schematic diagram of inverter](image2)
The conduction modes of the inverter are composed of combinations of the inverter switches $N_a$, $N_b$, and $N_c$. The detailed organization of the inverter switches is illustrated in Fig. (2). The eight switching modes of the inverter switches are demonstrated in table (1).

Recalling the concept of the voltage space vector that has demonstrated in section (2) and equations (1-4). The inverter load voltages are expressed according to equation (5).

$$
V_i = \begin{cases} 
0 & \text{k = 0,7} \\
\frac{2}{3} V_{dc} e^{i(k-1)\pi/3} & \text{k = 1,2,3,......6} 
\end{cases}
$$

(5)

Where:

- $V_i$: Inverter voltage vectors
- $k$: inverter voltage vector number that has been demonstrated in the first column in table (1).

The inverter voltage vectors are illustrated in Fig. (3).

<table>
<thead>
<tr>
<th>$N_a$</th>
<th>$N_b$</th>
<th>$N_c$</th>
<th>$V_{as}$</th>
<th>$V_{bs}$</th>
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</tr>
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</table>

Table 1 Inverter switching logic table and inverter phase and line voltage

Fig. (3) Space vector $V_i (t)$ and inverter voltage
4- Space vector modulation (SVM), concepts and theory

Space vector modulation is based on the concept of approximating the rotating reference voltage vector $V_s(t)$ to those that can be realized by three-phase PWM inverter [2,3]. This process is illustrated in Fig. (3).

As illustrated in Fig. (3) the stator space voltage vector $V_s(t)$ is superimposed on the stationary locus of the six output voltages of the inverter ($V_1$, $V_2$, $V_3$, ..., $V_6$), which is spatially oriented by $\pi/3$ intervals plus two null voltages $V_0$ and $V_7$. The voltage vector $V_s(t)$ is decomposed into its equivalent components of the eight-inverter voltage vectors according to the sector where the voltage vector is located, the conducting sequence for sector 1 will be as followed:

$$V_0 \rightarrow V_1 \rightarrow V_3 \rightarrow V_7 \rightarrow V_2 \rightarrow V_1 \rightarrow V_0$$

This conducting sequence is analyzed according to the switching modes that have been mentioned before in table (1). As illustrated in Fig. (3) the first step in the conducting sequence is $V_0$, the inverter switches which will conduct during this step is (4 6 2). The second is $V_1$, the conducting switches are (1 6 2). The third step is $V_2$, the conducting switches are (1 3 2). The fourth step is $V_3$, the conducting switches are (1 3 5). This process is repeated vice versa during the second half of the switching period ($T_s$). $T_s$ is divided into two equal halves, known and a sampling period. Just like regular sampling scheme, every switching period contains two equal sampling times, one for the leading edge and the other is for trailing edge. But in the contrary to regular sampling scheme, the widths of the two samples are equal and the pulse is centered inside the switching period. Also in SVM pulse position inside the switching period is completely controllable through changing between the null times at the beginning and the end of the switching period according to the criteria required. But in regular sampling this merit is absolutely out of the scope. The detailed process of PWM in SVM for sector 1 is illustrated in Fig. (4) [2-4]. This conducting sequence can be applied for other sectors.

![Fig. (4) Resultant three phase PWM waveform of one switching period](image-url)
From Fig. (4), it is obvious that:
\[ T_J/2 = t_0 + t_1 + t_2 + t_7 \]  

Where:
- \( T_J/2 \) is known as sampling period.
- \( t_1 + t_2 \) is known as modulation period.
- \( t_0 \) and \( t_7 \) are known as null periods or non-conducting periods.

5- SVM analysis and implementation with IPMSM

The practical implementation and analysis of the SVM current controlled inverter is achieved according to the block diagram shown in Fig. (5).

Fig. (5) Block diagram of SVM current scheme

Where:
- \( i_{qr}^*, i_d^* \): Synchronously rotating quadrature axis and direct axis reference currents respectively.
- \( i_q, i_d \): Synchronously rotating quadrature axis and direct axis actual currents respectively.
- \( V_{qr}^{*}, V_{dc}^{*} \): Stator synchronously rotating reference voltages before decoupling circuit.
- \( V_{qr}, V_{dc} \): Stator synchronously rotating reference voltages after decoupling circuit.
- \( \alpha \): Inclination angle.
- \( i \): Sector number.

The details of each block will be explained in the next subsections of this paper.
5.1 PI current controllers

Both reference and actual currents of d-axis and q-axis are compared. The error signals are manipulated through two PI controllers. The outputs are the stator voltage referred to rotor reference frame of both d-axis and q-axis.

Where: -

- $K_p$ and $K_i$: the PI current controller parameters.

\[
\begin{align*}
V_{qd} &= (i_{qr} - i_{qr})K_p + \int K_i(i_{qr} - i_{qr})dt \quad (7) \\
V_{qd} &= (i_{dr} - i_{dr})K_p + \int K_i(i_{dr} - i_{dr})dt \quad (8)
\end{align*}
\]

5.2 Stator Reference Voltage calculation

The d-axis and q-axis currents cannot be controlled independently only by $V_{qd}$ and $V_{qd}$ because of the cross-coupling effects such as $\omega L_{q} i_{qr}$ and $\omega L_{d} i_{dr}$. These factors are dominant in electrical machines with large inductance like IPMSM. Therefore these factors affect both current and torque responses. The back-EMF feedforward compensation scheme is applied to reduce the effect of disturbances and to achieve d-q axis decoupling control. The details of this function are illustrated in Fig. (6).

The back-EMF of q-axis: -

\[
E_{qr} = \omega_r (L_{dr} i_{dr} + \psi_f) \quad (9)
\]

The back-EMF of d-axis: -

\[
E_{dr} = -\omega_r (L_{qr} i_{qr}) \quad (10)
\]

Where: -

- $L_{dr}, L_{qr}$: direct and quadrature axis inductance, respectively.
- $\psi_f$: constant flux linkage of permanent magnets.
- $\omega_r$: the electrical angular velocity of IPMSM.

The stator space voltage vector $V_s$ magnitude is calculated as follows: -

\[
V_s = \sqrt{V_{qr}^2 + V_{dr}^2} \quad (11)
\]
5.3 Gating pulses periods calculations

These calculations are mainly based on the vector diagram demonstrated in Fig. (7). According to the concept of SVM the voltage vector is decomposed into its equivalents of eight inverter voltage vectors. This process mainly depends on inclination angle ($\alpha$) and sector number ($i$) as explained below. Calculations of $\alpha$ and $i$ are detailed in [9].

![Decomposition of stator space voltage vector ($V_s$) into inverter voltage vectors.](image)

$V_s$ is decomposed into its equivalent inverter voltage vectors $V_1$ and $V_2$.

\[
U_{s2} = b \cos(30^\circ) = 2b / \sqrt{3} = 2V_s \sin(\alpha) / \sqrt{3} \quad (12)
\]

\[
U_{s1} = V_s \cos(\alpha) - U_{s2} \cos(60^\circ) = V_s \cos(\alpha) - V_s \sin(\alpha) / \sqrt{3} \quad (13)
\]

Where:

$U_{s2}$ and $U_{s1}$: Projection of voltage vector $V_s$ into inverter vectors $V_2$ and $V_1$, respectively.

Periods, $t_1$ and $t_2$, are directly proportional to $U_{s1}$ and $U_{s2}$, respectively. These time intervals, $t_1$ and $t_2$, are calculated as follows [9]:

\[
t_1 = 1.5U_{s1}T_s / V_{dc} \quad (14)
\]

\[
t_2 = 1.5U_{s2}T_s / V_{dc} \quad (15)
\]

By substituting equations (12-13) into equations (14-15) we get

\[
t_2 = \sqrt{3}V_s \sin(\alpha)T_s / V_{dc} \quad (16)
\]

\[
t_1 = \frac{3}{2}V_sT_s \left( \cos(\alpha) - \frac{1}{\sqrt{3}} \sin(\alpha) \right) / V_{dc} \quad (17)
\]

From equation (6) and considering $t_0 = t_1$; then:

\[
t_0 = (T_s/2 - t_1 - t_2) / 2 \quad (18)
\]
6- Inverter operation modes and voltage limitations

In SVM one of the most undesired behavior when the space vector \( V_s(t) \) exceeds \((1/\sqrt{3})V_{dc}\). In this case the modulation periods must be rescaled to avoid negative null voltage times [9]. This rescaling process is done through the following equations:

\[
\begin{align*}
t_1'' &= \frac{t_1'}{t_1' + t_2'} \cdot (T_s / 2) \\
t_2'' &= \frac{t_2'}{t_1' + t_2'} \cdot (T_s / 2) \\
t_0 &= t_0' = 0.0
\end{align*}
\]

Where:
- \( t_1' \) and \( t_2' \): The modulation times for abnormal operation mode before rescaling.
- \( t_1'' \) and \( t_2'' \): The rescaled modulation times for abnormal operation mode.

The inverter operation modes are illustrated in Fig. (8). Detailed description for every mode is demonstrated in the incoming subsections.

![Fig. (8) SVM current controlled inverter operation modes.](image)

6.1 Linear mode

To insure a sinusoidal inverter output voltage waveform, the magnitude of stator space voltage vector \( V_s(t) \) must lie within the circle with radius \((1/\sqrt{3})V_{dc}\). In this case the inverter mode of operation is called linear.

6.2 Inverter operation in overmodulation range

In SVM technique, inverter voltage vector \( V_s(t) \) sometimes exceeds the value \((1/\sqrt{3})V_{dc}\) and it is still less than \((2/3)\ V_{dc}\), in this case the over modulation range is reached [6,9]. Reaching this mode of operation, the inverter loses its linear characteristics and a voltage compensation technique is required. The proposed scheme in overmodulation range will be described with reference to Fig. (8). As voltage vector increases above \((1/\sqrt{3})V_{dc}\) and still less than \((2/3)\ V_{dc}\), then the reference voltage vector exceeds inverter capability and therefore a modified voltage trajectory has to be set [6].

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As illustrated in Fig. (8). Reaching overmodulation range, every inverter sector is divided into three regions. First region is the middle region. It is completely out of inverter capability. The other two regions are the lower and upper sides. They are within inverter capability. These three regions are defined according to the inclination angle. Therefore there are two values of $\alpha$ because the overmodulation circle intersects with hexagon boundaries in two points. The compensation scheme that will be applied is based on changing the inclination angle to the inverter capability regions with keeping the voltage vector amplitude constant.

$$\alpha_g = \cos^{-1}((1/\sqrt{3})V_{dc}/V_s)$$  \hspace{1cm} (20)

Where $\alpha_g$ is called compensation angle.

$$\alpha_{lower} = \frac{\pi}{6} - \alpha_g$$ \hspace{1cm} \hspace{1cm} (21)

$$\alpha_{upper} = \frac{\pi}{6} + \alpha_g$$

Where $\alpha_{lower}$ and $\alpha_{upper}$ are the boundaries of inverter capabilities.

Inclination angle ($\alpha$) is modified according to the following conditions:

$$\begin{align*}
\alpha &= 0 \leq \alpha \leq \alpha_{lower} \\
\alpha &= \left(\frac{\pi}{3} - \alpha_g\right) \quad \alpha_{lower} \leq \alpha < \left(\frac{\pi}{6}\right) \\
\alpha &= \left(\frac{\pi}{3} + \alpha_g\right) \quad \left(\frac{\pi}{6}\right) \leq \alpha \leq \alpha_{upper} \\
\alpha &= \alpha_{upper} \leq \alpha \leq \left(\frac{\pi}{6}\right)
\end{align*}$$ \hspace{1cm} (22)

6.3 Inverter operation in six-step mode

Under completely saturated conditions $V_s \geq (2/3) V_{dc}$, the active modulation time is kept constant at $T_s$ and the inverter switches from circular space vector trajectory to hexagonal trajectory. This mode of operation is illustrated in Fig. (8).

7- Simulation results

The idea of event driven indicative flag is presented to demonstrate inverter operation mode. This flag takes three values, zero indicates linear mode, one indicates overmodulation mode and two indicates six-step mode. At rated values of motor speed, load torque and dc link voltage, the performance of SVM current controller operating at different switching frequencies is investigated. Figure (9) shows the performance at low switching frequency that equals 5 kHz. The deterioration of the stator current waveform is obvious in Fig. (9-a). Figure (9-d) indicates that inverter operates most of the time at overmodulation and six-step modes. Figure (9-b) shows that the motor stator phase voltage waveform is highly distorted. When increasing the switching frequency to 10 kHz the performance of the SVM current controller is better than that at 5 kHz, as shown in Fig. (10). Figure (10-d) indicates that the inverter operation at overmodulation and six-step modes are
less than that shown in Fig. (9-d). Figure (11) indicates the performance at switching frequency equals 15 kHz. This figure shows that this performance is the best because the inverter operates all the time at the linear mode, as shown in Fig. (11-d).

In Fig. (12) statistical relationship is illustrated between $f_s$ and the ratio of inverter operation modes. To avoid operation in relatively high ratio of six-step mode, there is a critical switching frequency ($f_c$) that inverter must not operate below. This frequency ($f_c$) is defined as the minimum frequency that inverter begins to operate totally at linear mode with no operation in six-step mode. Figure (12) indicates that this frequency is almost equal to 15.0 kHz at rated values of motor speed, load torque and dc link voltage.

Effects of the per unit values of dc link voltage ($V_{dc}$) on the SVM are illustrated in Fig. (13) and Fig. (14). This relationship has been derived at rated motor speed and motor load. Figure (13) demonstrates the effects of $V_{dc}$ on the SVM performance at critical switching frequency ($f_s = 15.0$ kHz). It is observed that at rated $V_{dc}$, the ratio of inverter operation at linear mode is 100%. While reducing $V_{dc}$, the inverter operates most of the time at six-step mode. Figure (14) demonstrates the effects of $V_{dc}$ on the critical switching frequency ($f_c$). It is observed that reducing $V_{dc}$, the critical switching frequency ($f_c$) must be increased to fulfill the required inverter operation at linear mode and then avoid distortion in stator current waveform.

The effect of motor speed on the inverter critical switching frequency ($f_c$), at rated values of both load and dc link ($V_{dc}$) is illustrated in Fig. (15). It indicates that the reference speed variations has no effect on $f_c$. The critical switching frequency ($f_c$) is constant at 15.0 kHz regardless any motor speed variations.

The effect of motor mechanical load on the inverter critical switching frequency ($f_c$), at rated values of speed and dc link ($V_{dc}$) has been demonstrated in Fig. (16). It is obvious, from this figure, that at loads less than the rated motor load, changing motor load has no effect on the value of $f_c$. Also this figure shows that when overloading the inverter, $f_c$ must be increased.
Fig. (9) SVM with switching frequency $f_s = 5.0$ kHz.
(a) Phase (a) current waveform.
(b) Phase (a) voltage waveform.
(c) Stator voltage vector $V_s$ waveform.
(d) Indicative flag waveform.

Fig. (10) SVM with switching frequency $f_s = 10.0$ kHz.
(a) Phase (a) current waveform.
(b) Phase (a) voltage waveform.
(c) Stator voltage vector $V_s$ waveform.
(d) Indicative flag waveform.
Fig. (11) SVM with switching frequency $f_s = 15$ kHz.

(a) Phase (a) current waveform.
(b) Phase (a) voltage waveform.
(c) Stator voltage vector $V_s$ waveform.
(d) Indicative flag waveform.

Fig. (12): Ratio of inverter operating modes at different switching frequencies.

Fig. (13): Inverter operation modes at different $V_{dc}$ with $f_s = 15$ kHz.
Fig. (14): The effect of \( V_{dc} \) on critical switching frequency \( (f_c) \).

Fig. (15): The effect of motor speed on critical switching frequency \( (f_c) \).

Fig. (16): The effect of motor load on critical switching frequency \( (f_c) \).

8- Conclusions

Complete inverter mathematical model and the basic concepts of SVM technique have been demonstrated in this paper. Inverter operation modes have been verified according to the amplitude and position of voltage vector \((V_s(i))\) with respect to the inverter hexagon.
It has been proved that with increasing the switching frequency, the inverter tends to operate in its linear mode. On decreasing the switching frequency, the inverter tends to operate in six-step mode, where the inverter is not capable of fulfilling the reference voltage vector $V_s(t)$. The direct result of such operation in six-step mode is the deterioration of the current wave. Statistical relationship has been developed between switching frequency and the ratio of inverter operation modes. The effects of changing dc link voltage, motor speed and mechanical load torque on the inverter modes of operation have been investigated.

9. Appendix

Motor parameters

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Base speed $\omega_b = 377$ rad/sec.

Current controller parameters

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10. References

دراسة تأثير تردد التوصيل والفصل على أداء العاكس ذو حكم التيار ذو موائمة فراغية إتجاهية

د/ فهمي الأولي
د/ مصطفى الشيشني
د/ علاء شبير عبد الغفار
قسم الهندسة الكهربائية-كلية الهندسة-جامعة المنوفية

حكم التيار هو الجزء الأساسي من منظومة التسبيح الكهربائي ذات التحكم الأجاهي. حيث يؤثر بشكل كبير على الأداء العام لمنظومة التحكم في المحرك بشكل كبير. الوظيفة الأساسية لحكم التيار هو التحكم في تيار الحمل للمحرك بحيث يتبع تيار الأمر المتولى من حكم السرعة. في هذا البحث تم تقديم نموذج رياضي خاص بالعاكس المغمد لمحرك ترانزيستور ذو موائمة فراغية. فالمطابقة من صمامات العاكس مثالية (لا يوجد زمن تأخير في الفتح والقفل)

هناك العديد من حاكم التيار التقليدية. مثل حاكم التيار ذو التبديل المباشط (Hysteresis) وحاكم التبديل المباشط ذو الوجه المثلثية (Triangular) وكمثال للبحث في الفترة الأخيرة هو حاكم التيار ذو الموائمة الفراغية الأجاهية (SVM). هذا النوع من حاكمات التيار يعتمد بشكل أساسي على تقريب الموجة الفراغية الخاص بجهد العرض الثابت حسب النموذج الرياضي لنموذج حاكم العاكس (Inverter) يتم بعد ذلك حساب أزمة التبديل الخاصة بأزمة التوصيل لكل وحدات الترانزيستور لكل وجه من الثلاثة أوجه الخاص بالعاكس حسب الجهود المفرقة لنموذج الجهود.

في هذا البحث تم دراسة تأثير تردد القبلة والفصل على كل من موائمة التيار لكل وجه من الأوجه ودراسة تأثير تردد على حالة التشغيل الخاصة بالعاكس. حيث تم تحديدها على مقاس (f) على كل من موائمة التيار لكل وجه. حيث تم تحديدها على مقاس (t) لحامدة التبديل الداخلي بالعاكس. في البداية فإن التمدد (Six-Step) واضح خاصية (Overmodulation) فوق الموئامات فوق التمدد. (Vdc). هنالك ثلاث حالات تشغيل للعاكس (أ) حالة فوق الموئامات (ب) حالة فوق التمدد (c) حالة الدخل (d) فسرت حالة السابقة عندما يتم استخدام الوقائع في حالة الخفيفة لكن نظرًا لعملية التمدد لحمادة تيار الحالة (e) فستكون شائعة فإن التمدد (Vdc) يتم على حالات التشغيل الخاصة بالعاكس. يتم أيضًا دراسة تأثير تردد القبلة المستمر (Vdc) على حالة التشغيل للعاكس. وتم عمل علاقات إحصائية تعبير عن هذه العلاقات.