

Single-Phase to Three-Phase Power Conversion Which Utilizes Parametric Oscillation

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1. Introduction

Transformation from any polyphase to the other different polyphase is easily accomplished by means of the transformer connections. In the statical transformation, however, from the single-phase to the polyphase, the suitable method has not yet been established, in particular some papers are published for single-phase to three-phase conversion concerning with ac electrification of railway.¹⁾⁶⁾

The three-phase motors offer distinct advantages over dc and single-phase ac motors. The dc motors have commutators which get dirty and short, brushes which wear and sometimes make poor contact, and mechanical contacts which constantly make and break during the process of speed regulation. All of these mechanical features are quite undesirable in view of their reliability. Single-phase ac motors, in addition to coming up to rated speed rather slowly, require starting circuitry.⁷⁾

Parametric oscillation is not an unfamiliar phenomenon, a playground swing and Melde's experiment are the examples of parametric oscillations in mechanical systems. In an electrical system, inductance and capacitance are the parameters which determine the resonant frequency. Parametric oscillation therefore can be produced in a resonant circuit periodically varying one of the reactive elements composing the resonant circuit.⁸⁾

A three-phase converter, described in this paper, is essentially circuit with a reactive element varying periodically at frequency $2f$ which generates a parametric oscillation at the subharmonic frequency f . In practice, the periodical variation is accomplished by applying an exciting current of frequency $2f$ to a balanced pair of nonlinear reactors.

A parametric oscillation could be accomplished by the cut-type core

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made of the oriented silicon steel sheets instead of the ferrite core.⁴⁾⁵⁾ If the primary circuit of the two oscillating circuits with the same magnetic properties is excited with the two phase currents which differ by π radians each other, the oscillating voltages at their secondaries have a phase difference of $\pi/2$ radians. The theoretical analysis and its experimental results of the single-phase to three-phase converter which utilizes parametric oscillation are described in this paper.

2. Theoretical Analysis

Basic principle. The parametric oscillating circuit is essentially a resonant circuit in which either the inductance or the capacitance is made to vary periodically. The parametric oscillating circuit diagram is shown in Fig. 1. The parametric oscillating circuit in Fig. 1 consists of coils wound around two magnetic toroidal cores M_a and M_b , a capacitor C . Each of the cores has two windings and these are connected together in a balanced configuration, one winding forming a resonant circuit with the capacitor C and being tuned to frequency f . An exciting current, which is a superposition of a dc bias current I_0 and a commercial frequency current of frequency $2f$, I_{2f} , is applied to the other winding, causing periodic variation in the inductance $L(t)$ of the resonant circuit at frequency $2f$ (see Fig. 3). The oscillating circuit using the cut-type magnetic core with three-legs made of the oriented silicon steel sheets is shown similarly in Fig. 2.

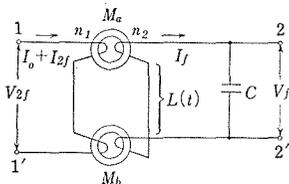


Fig. 1. Circuit diagram of parametric oscillation.

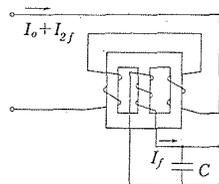


Fig. 2. Circuit diagram of parametric oscillation used cut-type with three legs.

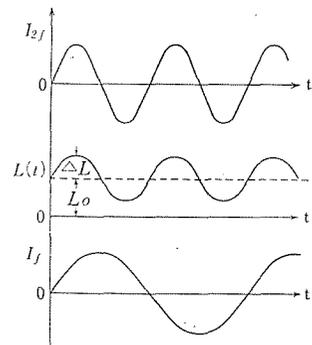


Fig. 3. Wave shapes of exciting current, I_{2f} , inductance, $L(t)$, and oscillating current, I_f .

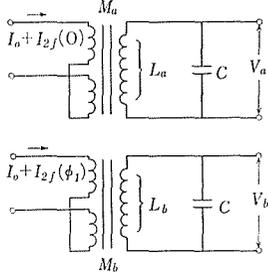


Fig. 4. Circuit diagram of two-phase transformable converter which utilizes parametric oscillation.

When the exciting circuits, which have two magnetic cores of three legs with the same magnetic properties and geometrical configuration, are excited by the exciting currents $I_{2f}(0)$ and $I_{2f}(\phi_1)$ which differ by ϕ_1 radians with respect to each other (Fig. 4), the inductances $L_a(t)$ and $L_b(t)$ of the resonant circuits are varied as

$$L_a(t) = L_0 (1 + 2\Gamma \sin 2\omega t), \quad (1)$$

$$L_b(t) = L_0 \{1 + 2\Gamma \sin (2\omega t + \phi_1)\}, \quad (2)$$

where $\omega = 2\pi f$, Γ is modulus of parametric excitation. Let us assume the presence of a sinusoidal ac current, I_f , in the resonant circuit at frequency f , which can be broken down into two components as follows:

$$I_f = I_s \sin \omega t + I_c \cos \omega t. \quad (3)$$

Then, assume that the rate of the variation of amplitudes of the sine and cosine components, I_s and I_c , are small compared with ω , the induced voltages, V_a , and, V_b , in the magnetic cores M_a and M_b will be given by respectively

$$\begin{aligned} V_a &= \frac{d}{dt} (L_a I_f) \\ &= \omega L_0 (I_s \cos \omega t - I_c \sin \omega t) \\ &\quad + 3\Gamma \omega L_0 (I_s \sin \omega t + I_c \cos \omega t) \\ &\quad - \Gamma \omega L_0 (I_s \sin \omega t - I_c \cos \omega t), \end{aligned} \quad (4)$$

$$\begin{aligned} V_b &= \frac{d}{dt} (L_b I_f) \\ &= \omega L_0 (I_s \cos \omega t - I_c \sin \omega t) \\ &\quad + 3\Gamma \omega L_0 \{I_s \sin (3\omega t + \phi_1) + I_c \cos (3\omega t + \phi_1)\} \\ &\quad - \Gamma \omega L_0 \{I_s \sin (\omega t + \phi_1) - I_c \cos (\omega t + \phi_1)\}. \end{aligned} \quad (5)$$

In (4) and (5), the first terms show the voltages due to a constant inductance L_0 , and the second terms or the third harmonic terms may be neglected in our approximation, since they are off resonance. The third terms, which are essential for the generation of the second subharmonic, show that the variable

parts of the inductances behave like a negative resistance for the sine component I_s , but behave like a positive resistance for the cosine component I_c .

Therefore, provided that the circuit (see Fig. 4) is nearly tuned to frequency f , the sine component I_s of any small oscillation will build up exponentially, while its cosine component will damp out rapidly. If the circuit were exactly linear, the amplitude would continue to grow indefinitely. Actually, the nonlinear B - H curve of the cores causes detuning of the resonance circuit and hysteresis loss also increases with increasing amplitude, so that a stationary state will rapidly be established, as in vacuum-tube oscillators.

The phase difference between the induced voltages, V_a and V_b , from (5) is given by⁵⁾

$$\text{third term} = -\Gamma \omega L_0 \sqrt{(I_s \sin \phi_1 - I_c \cos \phi_1)^2 + (I_c \sin \phi_1)^2} \sin(\omega t + \theta), \quad (6)$$

where

$$\theta = \tan^{-1} \frac{\sin \phi_1 - \cos \phi_1}{\sin \phi_1}, \quad \pi/2 < \phi \leq \pi. \quad (7)$$

When the phase difference ϕ_1 between these exciting currents, $I_{2f}(0)$ and $I_{2f}(\phi_1)$, is equal to π radians, $\phi_1 = \pi$, we can obtain that the phase difference, θ , between the two oscillating voltages is equal to $\pi/2$ radians at their secondaries. Therefore the three-phase power conversion from the single-phase power could be obtained with the two single-phase converters (parametric oscillating circuits) by connecting them open Δ , which is also called T . Both these connections are unsymmetrical and give slightly unbalanced voltages under load. The amount of this unbalancing is small and it's negligible under ordinary conditions, especially with T connections.

Oscillating output power in single-phase converter. A single-phase converter is a resonant circuit with a nonlinear inductance $L(t)$ varying periodically at frequency $2f$ which generates a parametric oscillation at the subharmonic frequency f .

If a dc bias field, H_0 , and a ac field of frequency $2f$, H_{2f} , are impressed on the two cores (see Fig. 1), then B - H loop becomes asymmetrical minor loop with respect to the origin,¹⁰⁾ Fig. 5.

The induced magnetic flux Φ by current i on the core with winding number of turns, n , will be given by

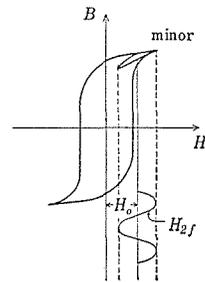


Fig. 5. B - H loop under excitation. H_0 is dc bias field, H_{2f} is ac field at frequency $2f$.

$$\Phi = g(H) = g(ni) \quad (8)$$

Total flux Φ_a and Φ_b in the each cores, M_a and M_b , are given by

$$\left. \begin{aligned} \Phi_a &= g(n_1 I_0 + n_1 i_{2f} + n_2 i_f) \\ \Phi_b &= g(n_1 I_0 + n_1 i_{2f} - n_2 i_f) \end{aligned} \right\} \quad (9)$$

Let us assume that $n_1 i_{2f} \pm n_2 i_f$ is smaller than $n_1 I_0$, flux Φ_a and Φ_b are derived as follows:

$$\begin{aligned} \Phi_a &= g(n_1 I_0) + g'(n_1 I_0)(n_1 i_{2f} + n_2 i_f) \\ &\quad + \frac{1}{2!} g''(n_1 I_0)(n_1 i_{2f} + n_2 i_f)^2 + \frac{1}{3!} g'''(n_1 I_0)(n_1 i_{2f} + n_2 i_f)^3 + \dots \end{aligned} \quad (10)$$

$$\begin{aligned} \Phi_b &= g(n_1 I_0) + g'(n_1 I_0)(n_1 i_{2f} - n_2 i_f) \\ &\quad + \frac{1}{2!} g''(n_1 I_0)(n_1 i_{2f} - n_2 i_f)^2 + \frac{1}{3!} g'''(n_1 I_0)(n_1 i_{2f} - n_2 i_f)^3 + \dots \end{aligned} \quad (11)$$

Because of $n_1 I_0 > n_1 i_{2f} \pm n_2 i_f$, until the fourth term of the expansion only needs to be considered. The secondary induced voltages, V_a and V_b , are expressed following:

$$\left. \begin{aligned} v_a &= -n_2 \frac{d\Phi_a}{dt} \\ v_b &= -n_2 \frac{d\Phi_b}{dt} \end{aligned} \right\} \quad (12)$$

Oscillating voltage of frequency f , v_f , in the terminals 2-2' (Fig. 1), can be expressed as

$$\begin{aligned} v_f &= v_a - v_b = -n_2 \frac{d}{dt}(\Phi_a - \Phi_b) \\ &= -2n_2^2 \frac{d}{dt} \left\{ g'(n_1 I_0) i_f + g''(n_1 I_0) n_1 i_{2f} i_f \right. \\ &\quad \left. + \frac{1}{2} g'''(n_1 I_0) (n_1^2 i_{2f}^2 + \frac{1}{3} n_2^2 i_f^3) \right\}. \end{aligned} \quad (13)$$

Putting

$$i_{2f} = I_{2f} \cos 2\omega t = \frac{1}{2} I_{2f} (\varepsilon^{j2\omega t} + \varepsilon^{-j2\omega t}), \quad (14)$$

$$i_f = I_f \cos(\omega t + \phi_2) = \frac{1}{2} I_f \{ \varepsilon^{j(\omega t + \phi_2)} + \varepsilon^{j(\omega t + \phi_2)} \}, \quad (15)$$

and then substitution of (14) and (15) into (13) yields

$$v_f = -j\omega I_f \varepsilon^{j(\omega t + \phi_2)} \left\{ n_2^2 g' (n_1 I_0) + \frac{1}{2} n_1 n_2^2 I_f g'' (n_1 I_0) \varepsilon^{-j2l_2} + \frac{1}{4} n_1^2 n_2^2 I_f^2 g''' (n_1 I_0) \right\}. \quad (16)$$

(16) represents the oscillating voltage in resonant circuit with capacitor C and neglects the harmonic terms, which are off resonance.

For the relation, in general, between the magnetic field, H , and the magnetic density, B , on the core, we obtain

$$H = U \sinh uB. \quad (17)$$

Equation (17) will be rewritten as:

$$\phi = \frac{s}{u} \sinh^{-1} \frac{ni}{Ul}, \quad (18)$$

where n is the winding number of turns on the core, u and U are constants for the core that depends upon the magnetic saturation characteristics and the material dimension, l is the length of the magnetic path.

Therefore, if $(n_1 I_0 / Ul)^2 \gg 1$

$$\left. \begin{aligned} g' (n_1 I_0) &= \frac{s}{un_1 I_0} \\ g'' (n_1 I_0) &= -g' (n_1 I_0) \frac{1}{n_1 I_0} \\ g''' (n_1 I_0) &= g' (n_1 I_0) \frac{2}{(n_1 I_0)^2} \end{aligned} \right\}. \quad (19)$$

Substituting (19) for (16), the oscillating voltage, v_f , is given by

$$v_f = -j\omega I_f \varepsilon^{j(\omega t + \phi_2)} \left[n_2^2 g' (n_1 I_0) - \frac{1}{2} \left(\frac{I_2 f}{I_0} \right) n_2^2 g' (n_1 I_0) \varepsilon^{-j2l_2} + \frac{1}{4 I_0^2} \left(\frac{n_2}{n_1} \right)^2 n_2^2 g' (n_1 I_0) I_f^2 + \frac{1}{2} \left(\frac{I_2 f}{I_0} \right)^2 n_2^2 g' (n_1 I_0) \right] \quad (20)$$

The terms in the parentheses, in (20), are inductance of the exciting circuit, (20) will be rewritten as

$$v_f = -j\omega I_f \varepsilon^{j(\omega t + \phi_2)} \left[L_0 - \frac{1}{2} \Delta L \varepsilon^{-j2l_2} + \frac{1}{2} \beta L_0 I_f^2 \right], \quad (21)$$

where

$$L_0 = n_2^2 g' (n_1 I_0) + \left(\frac{I_2 f}{I_0} \right)^2 n_2^2 g' (n_1 I_0), \quad (22)$$

$$\Delta L = \left(\frac{I_2 f}{I_0} \right) n_2^2 g' (n_1 I_0), \quad (23)$$

$$L'(t) = \frac{1}{2 I_0^2} \left(\frac{n_2}{n_1} \right)^2 n_2^2 g' (n_1 I_0) I_f^2 = \beta L_0 I_f^2, \quad (24)$$

$$\beta = \frac{1}{2 I_0^2} \left(\frac{n_1}{n_2} \right)^2. \quad (25)$$

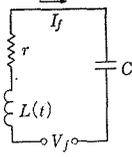


Fig. 6. Secondary tuning circuit diagram.

An oscillating circuit consisting tuning capacitor C , inductance $L(t)$ and effective resistance r connected in series is shown in Fig. 6.

Let $I_f e^{j(\omega t + \phi_2)}$ be the oscillating current of frequency f , the loop emf, v_f , is given by

$$v_f = r [1 + jQ(2\delta + \beta I_f^2 - \Gamma \varepsilon^{-j2\phi_2})] I_f e^{j(\omega t + \phi_2)} \quad (26)$$

where

$$Q = \frac{\omega L_0}{r},$$

$$\Gamma = \frac{\Delta L}{2L_0},$$

$$\delta = \frac{1}{2\omega L_0} \left(\omega L_0 - \frac{1}{\omega C} \right).$$

The absolute value of the oscillating voltage of frequency f , V_f , becomes

$$V_f = r I_f \sqrt{1 + Q^2 (2\delta + \beta I_f^2 - \Gamma)^2}. \quad (27)$$

Losses in a single-phase converter. The losses in a converter are the iron loss, the primary and secondary copper losses. The iron loss is nearly constant and nearly independent of the load. The primary and secondary copper losses vary as the squares of the primary and secondary currents. The iron loss is caused by the variation of the flux in the iron core and depends upon the frequency, the maximum value of the flux density in the core, the quality of the iron, the wave form of the time variation of the flux in the core, the thickness of the laminations and the volume or weight of the core. The iron loss can be separated into the loss due to eddy currents

and the loss due to hysteresis. These components follow different laws.

Hysteresis loss. — Let B_m be the maximum flux density of a converter and let $2f$ be the frequency, the hysteresis loss, P_h , is

$$P_h = 2f\sigma_h B_m^2 w \cdot 10^{-2}, \quad B_m > 1 \text{ wb/m}^2, \quad (28)$$

where w is the weight of a converter core and σ_h is a constant for used material.

Eddy-current loss. — Let k_f be the wave form factor and let Δ be the thickness of the laminations, eddy-current loss, P_e , is

$$P_e = \sigma_e (2f \Delta k_f B_m)^2 w \cdot 10^{-2}, \quad (29)$$

where σ_e is a constant determined by used material.

The expression for the iron loss, P_{h+e} , can be written

$$P_{h+e} = 2B_m^2 (\sigma_h f + 2\sigma_e \Delta^2 k_f^2 f^2 \cdot 10^{-2}) w \cdot 10^{-2}. \quad (29)$$

Copper loss. — Let r_1 be the primary winding resistance and let r_2 and r'_2 are the secondary winding resistance in the tuning and the load circuits respectively, in Fig. 7. The copper loss, P_c , in a single-phase converter is

$$P_c = r_1 I_0^2 + k_{m1} r_1 I_2 f^2 + k_{m2} (r'_2 I_f'^2 + r_2 I_f^2), \quad (30)$$

where k_m is the ratio of the ac resistance for the dc resistance, in general, $k_m = 1.1 \sim 1.25$.

Conversion efficiency of a single-phase converter. The conversion efficiency of a single-phase converter, η , is given by

$$\eta = \frac{V_f' I_f' \cos \phi_2}{V_f' I_f' \cos \phi_2 + P_{h+e} + r_1 I_0^2 + k_{m1} I_2 f^2 r_1 + k_{m2} (r'_2 I_f'^2 + r_2 I_f^2)}. \quad (31)$$

If the iron loss, P_{h+e} , and the secondary oscillating voltage, V_f , are assumed constant, the maximum conversion efficiency occurs for that value of the secondary oscillating current, I_f , which make the differential coefficient of the efficiency with respect to the secondary oscillating current, I_f' , zero. The maximum conversion efficiency is to be obtained under the following condition:

$$P_{h+e} + r_1 I_0^2 + k_{m1} r_1 I_2 f^2 + k_{m2} r_2 I_f^2 = k_{m2} r'_2 I_f'^2. \quad (32)$$

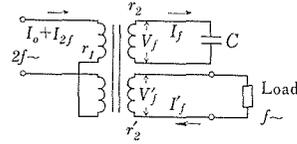


Fig. 7. Circuit diagram of single-phase (2f-f) converter.

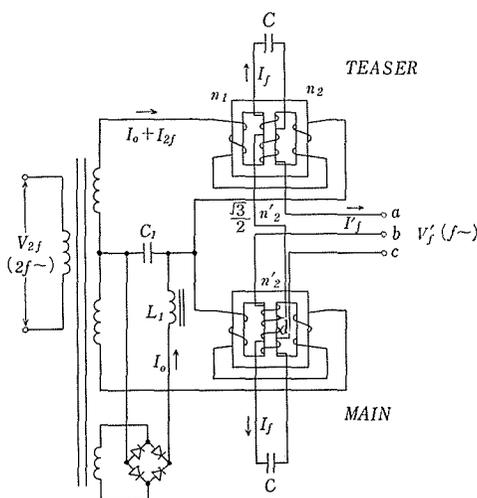


Fig. 8. Complete schematic of three-phase converter.

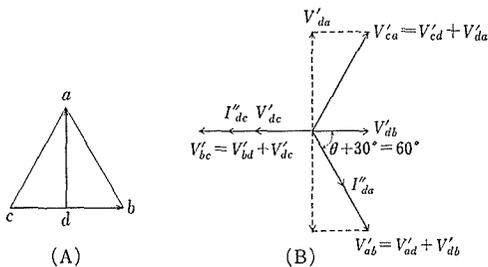


Fig. 9. Vector diagram for the T connection.

c. If the secondaries are similarly connected, they supply three-phase power at a voltage which, except for the impedance drops, is equal to the impressed voltage divided by the ratio of transformation.

If the impressed primary voltages are balanced, the secondary voltages V_{ab} , V_{bc} , and V_{ca} are each equal to $2V_{cd}$. The voltages V_{da} , V_{dc} and also V_{db} , V_{db} are in quadrature.

$$\dot{V}_{ca} = \dot{V}_{da} + \dot{V}_{cd} \tag{33}$$

The angle acd is 60 degrees and

$$\frac{V_{da}}{V_{ca}} = \sin 60^\circ = 0.866. \tag{34}$$

The teaser converter, therefore, should be wound for a voltage which is

Three-phase conversion. If the primary circuit of the two oscillating circuits (converters) with the same current ratings but with different voltage ratings is excited with respect to the two phases which differ by π radians with each other, their secondary oscillating voltages has a phase difference of $\pi/2$ radians. The complete schematic of the three-phase converter is shown in Fig. 8.

The secondary winding of the converter, which is called a teaser, is connected to the middle of the other, which is called a main, Fig. 9 (A) serves either for the diagram of connections or for the vector diagram of voltages. The teaser converter is indicated by ad and the main converter by cb . The three-phase voltages are impressed across the terminals a , b ,

86.6 percent of the voltage of the main converter.

A vector diagram for the T connection is shown in Fig. 9 (B). The load is assumed to be balanced with respect to the secondary voltage. The angle of lag θ is assumed to be 30 degrees with respect to the secondary voltage. All vectors are referred to the secondary. The voltage V'_{da} is in phase with the Y voltage of the system. The converter da carries line current. Therefore the power factor $\cos\phi_2$ for the teaser converter is the same as the power factor of the three-phase load. The capacity of the T system for the three-phase conversion is somewhat less than the sum of the capacities of the two converters.

Consider a load at power factor $\cos\phi_2$ and assume that the teaser converter da is wound the correct voltage. Let the line current and line voltage of the three-phase system be I_f and V_f , the output power of the teaser converter, P_t , is given by

$$P_t = V_f' I_f' \cos\phi_2. \quad (35)$$

Because there is the transformable operation between the resonant circuit and the secondary load circuit, the three-phase output power, P_{out} , becomes

$$P_{out} = \sqrt{3} \alpha r_2 I_f I_f' \sqrt{1 + Q^2 (2\delta + \beta I_f^2 - I')^2}, \quad (36)$$

where α is the turn ratio between the resonant circuit and the oscillating circuit.

3. Experimental Results

The cores used in the three-phase converter are of cut-type with silicon steel sheets. These magnetization characteristics have flux density $B = 1.74$ wb/m² at magnetic strength $H = 2 \cdot 10^{-2}$ ampere-turns/m. The chemical components and geometrical dimension of the core are shown in Table-1 and Fig. 10. The winding number of turns and its resis-

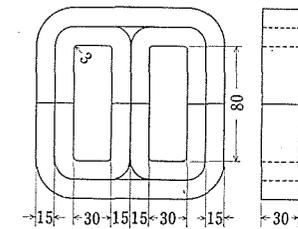


Fig. 10. Dimension of core.

Table-1. Chemical components of core

Si	2.5~3.5%
C, O, N, Al, Mn	extremely small
Fe	all of the rest

tances are shown in Table-2.

Oscillation diagram. Fig. 11 shows typical examples of the oscillation diagram (main converter) of an actual single-phase converter. In the figure, the abscissa represents the exciting current of frequency $2f$, I_{2f} , and the ordinate represents the

Table-2. Winding number of turns and its resistance on the converter

	Teaser	Main
n_1	300 turns	300 turns
n_2	500 turns	500 turns
n_2'	433 turns	500 turns
r_1	0.9 ohms	0.9 ohms
r_2	4.7 ohms	4.7 ohms
r_2'	4.6 ohms	5.7 ohms

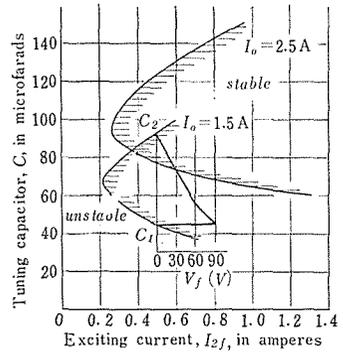


Fig. 11. Oscillation diagram.

tuning capacitor C . The single-phase converter certainly begins to oscillate by the all exciting current of frequency $2f$ in these stable regions.

From the oscillation diagram in Fig. 11, we can obtain the quantity Γ_c as follows:

$$\Gamma_c = \frac{C_2 - C_1}{C_1 + C_2}, \tag{37}$$

where C_1 and C_2 are the upper and the lower limit tuning capacitor in the binary oscillation region respectively, which are naturally changed by the dc bias I_0 and the exciting current I_{2f} . In practice, if Q of the tuning circuit is large, and make an exception of the influence of febrility, Γ_c corresponding to modulus of parametric excitation, Γ' .

Γ_c characteristics dependence the exciting current I_{2f} , dc bias I_0 as parameters, is shown in Fig. 12. When the dc bias I_0 is equal to 1.5 amperes, $\Gamma_c = 0.35$ for $I_{2f} = 0.5$ amperes.

Single-phase conversion charateristics. Fig. 13 shows the resistive load characteristics of the oscillating load voltage, V_f , and the oscillating current of frequency f (30 c/s), I_f , the exciting current of frequency $2f$ (60 c/s), I_{2f} , and the resonant current, I_f , in the main converter under the experimental conditions of $V_{2f} = 80$ volts, $I_0 = 1.5$ amperes and $C = 60$ microfarads.

The figure shows that the oscillating load voltage, V_{2f} , has the constant-voltage characteristics in the high resistive load

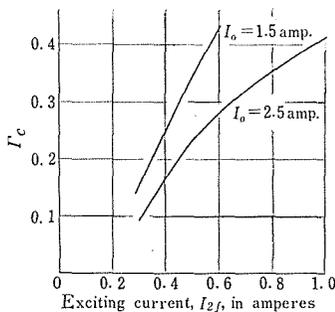


Fig. 12. Γ_c Characteristics

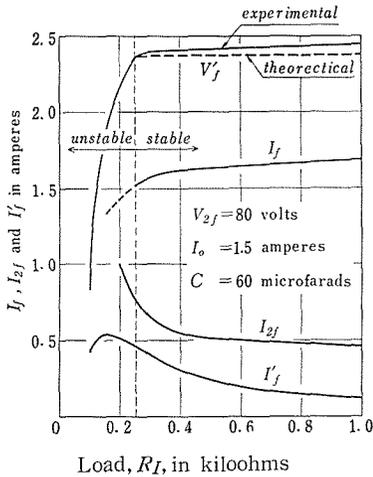


Fig. 13. Load characteristics of a single-phase converter.

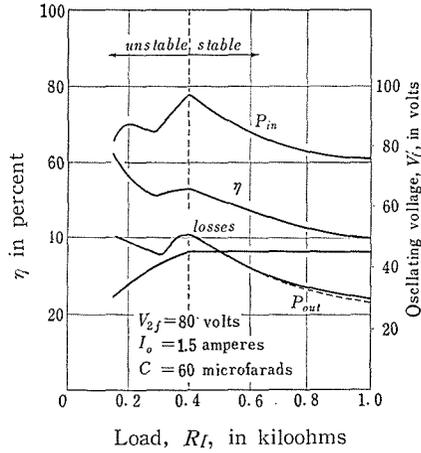


Fig. 14. P_{in} , P_{out} and efficiency, η .

region above 250 ohms. As the resistive load of the single-phase converter increases, the oscillating load current of frequency f , I_f , decreases exponentially. All these characteristics in Fig. 13 gives rise to the unstable phenomenon in the lower resistive load region less than 250 ohms.

The exciting input power, P_{in} , and the oscillating output power, P_{out} , of the single-phase converter (main) are shown in Fig. 14. The comparison of the experimental and the theoretical value, from (27), is also shown in the same figure. The derived single-phase conversion efficiency, η , from the figure is about 50 percent in the higher load region. The descent of the conversion efficiency is caused by the losses in a converter, which consist of

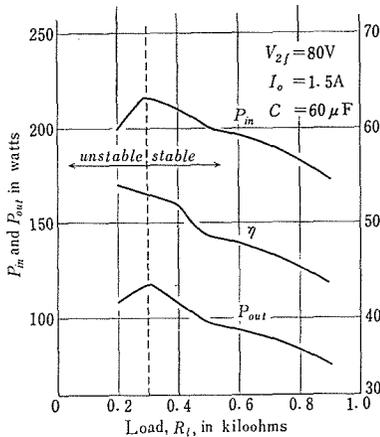


Fig. 15. Loss distribution in a single-phase converter.

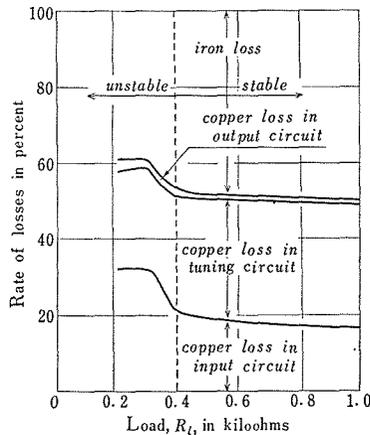


Fig. 16. P_{in} , P_{out} and efficiency in a three-phase converter.

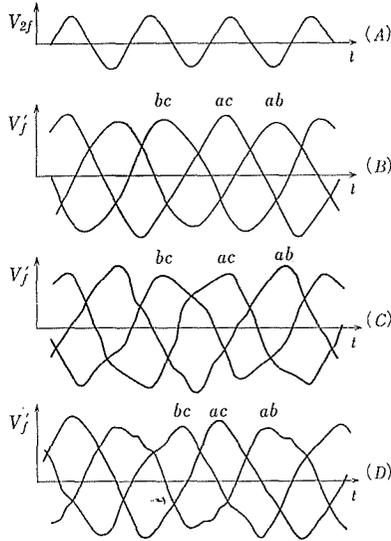


Fig. 17. Oscillograms of exciting voltage, V_{2f} , -(A) and oscillating load voltages, V_f , of the no load-(B), resistive load-(C) and inductive load-(D).

also occupies about half of all losses, the three-phase conversion efficiency, η , is nearly 50 percent. The performance of the three-phase converter is stabilized in the higher resistive load more than 300 ohms, because of detuning effect by the lower resistive load, the oscillation becomes unstable at less than the load of 300 ohms.

Fig. 17 shows the oscillograms of the exciting voltage of frequency $2f$, V_{2f} (A), and the oscillating load voltage of frequency f , V_f , of the no load (B), the resistive load (C), and the inductive load (D). The resistive load is connected with Δ connection, the value of resistance is equal to 500 ohms respectively, and the inductive load is the three-phase induction motor, which has the synchronous speed of 400 revolutions per minute.

4. Conclusion

This paper presents the basic principle and the experimental results of the single-phase to three-phase power conversion which utilizes parametric excitation. Though the obtained conversion efficiency of the three-phase converter is yet nearly 50 percent, it is possible to obtain the more higher conversion efficiency by means of the economy-sized.

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the primary copper loss, $2r_1 (I_0^2 + k_{m1} I_{2f}^2)$, the iron loss, P_{h+e} , and the secondary copper loss, $k_{m2} (I_f'^2 r_2 + I_f'^2 r_2)$.

Fig. 15 shows the rate of each copper losses and the iron loss for the total losses. It is obvious that the copper losses, which consist of the losses in the primary, the tuning and the secondary circuits, account for about 50 percent of all losses.

Three-phase conversion characteristics. The characteristic of the exciting input power of frequency $2f$, P_{in} , the oscillating output power of frequency f , P_{out} , and the conversion efficiency, η , in the single-phase to three-phase converter is shown in Fig. 16. In this case, the ohmic loss

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