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This paper deals with the practical design for output filter of switched mode amplifier. It has been designed by making resort of the experimental data. And the theoretical criteria have never been presented because of the various demands for it. A demand of continuous current flow into the filter is indispensable for the correct performance of this kind of amplifier. This demand is the major obstacle to design the filter and disturbs the clarification of its design criteria.

In this paper, sufficient condition of continuous current flow into the filter is derived and it is concluded that Butterworth filter is satisfied with all conditions or requirements for the output filter of the switched mode amplifier.

This results made it very easy to design filters for phase controlled rectifier and d.c. chopper circuits as well as switched mode amplifier.

1. Introduction

Recently, the switched mode linear[1] or nonlinear amplifier[2] using switching devices such as transistors and thyristors is used for high power, high efficiency amplifier. The switched mode amplifier has very high efficiency but its output voltage contains undesired frequency components in the vicinity of the integral multiples of switching repetition frequency besides the desired output signal frequency components in the low frequency range. It is necessary to insert a proper low pass filter between the amplifier and its load in order to remove the undesired components.

This output filter requires various demands such as continuous input current flow, choke input, small power dissipation and flat frequency response in the pass band by which the frequency response of overall amplifier is determined. In designing the filter, it is the most troublesome condition that the input current of the filter must continuously flow to yield the correct voltage waveform accross its input terminals.

In this paper, the continuous current flow condition is studied on the view of designing the filter which makes the amplifier realize the wanted output voltage waveform, which is not only amplified but has high fidelity to input voltage waveform. In the latter part, it is shown that Butterworth filter is satisfied with required conditions.

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2. Continuous current condition

The fundamental construction of the switched mode amplifier is shown in Fig. 1. Transistors or thyristors may be used for unilateral switching devices S_1, S_2.

While the switch S_1 or S_2 is conducting, the anode source voltage appears accross input terminals $a-a'$ of the filter. As soon as the switch S_1 or S_2 stops conduction, the wheeling diode D_W begins to conduct and the voltage of the point

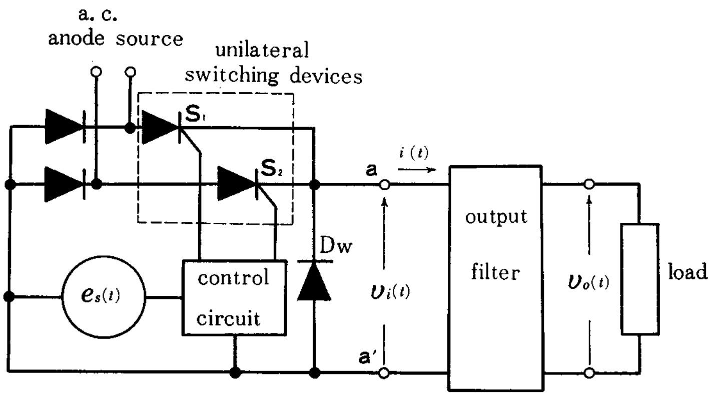
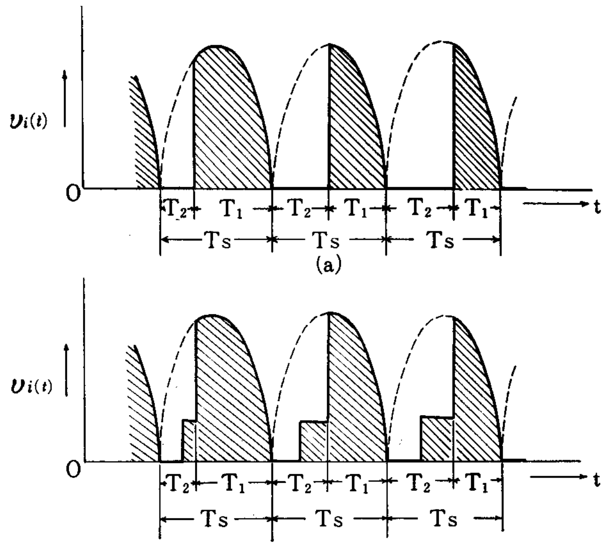


Fig. 1 Fundamental construction of switched mode amplifier



(b)

(a) for continuous input current flow
(b) for discontinuous input current flow
T_1 : conduction interval of switching devices
T_2 : cut-off interval of switching devices
$T_s = \frac{2\pi}{\omega_s}$: period of switching repetition

Fig. 2 Input voltage waveforms of filter

a falls to zero. For example, the sinusoidal voltage is used as the a.c. anode source voltage, the input voltage of the filter is shown as a train of pulses occured every half cycle of a.c. anode source voltage as shown in Fig. 2 (a). The control circuit generates the signal which controls conduction intervals of the switches S_1 and S_2 according to the wanted relation between the input signal voltage and the average value of output voltage in every half cycle.

In order to perform correctly above operation, at the instance of the switch S_1 or S_2 cutting off, the wheeling diode D_W must begin to conduct and the input current of the output filter must be continuously flowing into the filter in the direction of the arrow indicated in Fig. 1.

If the input current of the filter changes its direction, neither switches S_1, S_2 nor wheeling diode D_W conducts, and uncorrect voltage waveform appears accross input terminals of the filter $a-a'$ as shown in Fig. 2 (b)[3]. This abnormal voltage disturbs the correct relation between input and output voltage, for example the linear relation between input signal voltage $e_S(t)$ and output voltage $v_o(t)$ of the filter in the case of switched mode linear amplifier.

The condition of the continuous current flow can be replaced by the condition that the input current in the direction of the arrow in Fig. 1 is non-negative when the pulse train, which is obtained by suitable PWM of anode source voltage for required relation between input and output voltage, is applied to the input terminals of the filter $a-a'$. The continuance of the current into the output filter can be examined by means of non-negativity of the input current $i(t)$.

3. Necessary and sufficient condition for continuous current flow

Using the Fourier transform $V_i(j\omega)$ of the input voltage $v_i(t)$ accross $a-a'$, and the input admittance $Y(j\omega)$ of the filter, the Fourier transform $I(j\omega)$ of the input current $i(t)$ is given by

$$I(j\omega)=Y(j\omega)\cdot V_i(j\omega) \tag{1}$$

The input current $i(t)$ is given as the inverse transform of the right hand side in Eq. (1).

$$i(t)=\frac{1}{2\pi}\int_{-\infty}^{\infty} Y(j\omega)\cdot V_i(j\omega)e^{j\omega t}d\omega \tag{2}$$

Non-negativity of the right hand side in Eq. (2) is the necessary and sufficient condition for continuous current flow, but to judge by it is very difficult, moreover it is not so general judging condition since it is necessary to determine the Transformed input voltage $V_i(j\omega)$.

On the other hand, the input current $i(t)$ in Eq. (2) can be expressed in the time domain as shown in Eq. (3) using the impulse response $y(t)$ of the input admittance $Y(j\omega)$, and $y(t)$, $Y(j\omega)$ are connected one another with the Fourier and inverse Fourier transformation in Eqs. (4), (5).

$$i(t)=\int_{-\infty}^{\infty} y(t-\tau)\cdot v_i(\tau)d\tau \qquad (3)$$

$$Y(j\omega)=\int_{-\infty}^{\infty} y(t)\cdot e^{-j\omega t}dt \qquad (4)$$

$$y(t)=\frac{1}{2\pi}\int_{-\infty}^{\infty} Y(j\omega)\cdot e^{j\omega t}d\omega \qquad (5)$$

From Eq. (3), the necessary and sufficient condition in *strict sense* may be said that the impulse response $y(t)$ is non-negative, since the input voltage $v_i(t)$ is non-negative pulse train as shown, for example, in Fig. 2(a), where "*strict sense*" means that any non-negative input voltage $v_i(t)$ is allowable for continuous current flow. It is not so easy to design the filter whose impulse response of input admittance $y(t)$ is non-negative, and the non-negativity of $y(t)$ is not practical condition to design the filter for continuous input current flow.

As mentioned above, some of conditions for continuous input current flow are theoretically deduced, but they are not so practical and not convenient for designing the output filter of the switched mode amplifier.

3-1. Practical condition for continuous input current flow

Some of input voltages, for example, the pulse train as shown in Fig. 2(a), do not require non-negativity of the impulse response $y(t)$ of the input admittance $Y(j\omega)$ for the condition of continuous input current flow. In that case, practical and easy-design condition is discussed as follows. Usually, it is more convenient to deal with the filter function in the frequency domain than in the time domain. The input admittance $Y(j\omega)$ is expressed in Eq. (6) by the input conductance $G(\omega)$ and susceptance $B(\omega)$.

$$Y(j\omega)=G(\omega)+jB(\omega) \qquad (6)$$

If the filter circuit is constructed with passive components, $G(\omega)$ and $B(\omega)$ are individually even and odd function. From Eqs. (5), (6) $y(t)$ is deduced as follows:

$$\begin{aligned} y(t) &=\frac{1}{2\pi}\int_{-\infty}^{\infty} \{G(\omega)+jB(\omega)\}\, e^{j\omega t}d\omega \\ &=\frac{1}{2\pi}\int_{-\infty}^{\infty} \{G(\omega)+jB(\omega)\}(\cos\omega t+j\sin\omega t)d\omega \\ &=\frac{1}{2\pi}\int_{-\infty}^{\infty} [\{G(\omega)\cos\omega t-B(\omega)\sin\omega t\}+j\{G(\omega)\sin\omega t+B(\omega)\cos\omega t\}]d\omega \end{aligned} \qquad (7)$$

In Eq. (7), the real part of the right hand side is even function and imaginary part is odd function.

$$y(t)=\frac{1}{\pi}\int_{0}^{\infty} \{G(\omega)\cos\omega t-B(\omega)\sin\omega t\}\, d\omega \qquad (8)$$

From the law of causation,

$$y(-t)=\frac{1}{\pi}\int_0^\infty \{G(\omega)\cos\omega t+B(\omega)\sin\omega t\}\,d\omega=0 \qquad (t>0) \tag{9}$$

Applying the relation of Eq. (9) to Eq. (8),

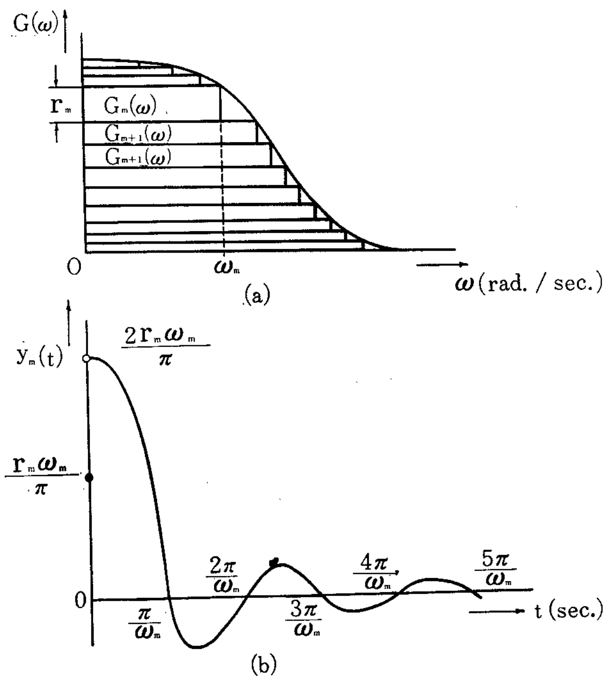
$$y(t)=\frac{2}{\pi}\int_0^\infty G(\omega)\cos\omega t\omega d=-\frac{2}{\pi}\int_0^\infty B(\omega)\sin\omega t d\omega \qquad (t>0) \tag{10}$$

From Eq. (10), the condition of the continuous input current flow can be discussed by using characteristics of either input conductance $G(\omega)$ or susceptance $B(\omega)$. It is assumed that Eqs. (11), (12) are the sufficient condition for the output filter of continuous input current flow.

$$G(\omega)\begin{cases}\geqq 0 & (\omega\lessapprox\omega_S)\\ \approx 0 & (\text{otherwise})\end{cases} \qquad (\omega\geqq 0) \tag{11}$$

$$\frac{d}{d\omega}G(\omega)\leqq 0 \qquad (\omega\geqq 0) \tag{12}$$

Where, ω_S represents the angular frequency of switching repetition, and equals to two times angular frequency of the a.c. anode source. Eq. (11) means the pass band of $G(\omega)$ is sufficiently narrower than ω_S, and is mainly required for the continuous input current. Eq. (12) means $G(\omega)$ is monotonously reduced to zero, and required for realizing the wanted overall frequency response of the switched mode amplifier.



(a) approximation of $G(\omega)$
(b) expression of $G_m(\omega)$ in the time domain
Fig. 3 Characteristics of $G(\omega)$ and $G_m(\omega)$

It can be said that if only the conditions of Eqs. (11), (12) are met, conditions for the output filter such as continuous input current flow, choke input, flat frequency response in pass band are satisfied.

The input conductance $G(\omega)$ may be approximated by sum of rectangles $G_m(\omega)$ as shown in Fig. 3 (a) where r_m and ω_m represent individually the amplitude and maximal frequency of the m th rectangle $G_m(\omega)$. Using Eq. (10) the inpulse response $y_m(t)$ corresponding to $G_m(\omega)$ is given by,

$$y_m(t) = \frac{2}{\pi}\int_0^\infty G_m(\omega)\cdot\cos\omega t d\omega = \frac{2}{\pi}\cdot r_m\cdot\frac{\sin\omega_m t}{t}\cdot U(t) \tag{13}$$

where $U(t)$ represents unit step function and $y_m(t)$ is shown in Fig. 3 (b). The impulse response $y(t)$ corresponding to $G(\omega)$ is given as sum of $y_m(t)$.

$$y(t) = \sum_{m=1}^\infty y_m(t) = \sum_{m=1}^\infty \frac{2}{\pi}\cdot r_m\cdot\frac{\sin\omega_m t}{t}\cdot U(t) \tag{14}$$

The continuance of input current flow for several cases will be proved by using Eq. (14) under the condition of Eq. (11) and Eq. (12).

3-1-1. Constant conduction interval of switches

In the case that the conduction interval T_1 in Fig. 2 (a) is regarded as constant, input voltage $v_i(t)$ of the output filter becomes periodic function with its primary period $T_S = \frac{2\pi}{\omega_S}$, and is expressed by Fourier series.

$$v_i(t) = \sum_{n=0}^\infty \epsilon_n(a_n\cos n\omega_S t + b_n\sin n\omega_S t) \tag{15}$$

where,

$$\left.\begin{aligned} a_n &= \frac{1}{T_S}\int_0^{T_S} v_i(t)\cdot\cos n\omega_S t dt \\ b_n &= \frac{1}{T_S}\int_0^{T_S} v_i(t)\cdot\sin n\omega_S t dt \\ \epsilon_n &= \begin{cases} 1 & (n=0) \\ 2 & (\text{otherwise}) \end{cases} \end{aligned}\right\} \tag{16}$$

Input voltage $v_i(t)$ is non-negative function and its average value a_0 is non-negative.

$$a_0 \geqq 0 \tag{17}$$

Using Eqs. (13), (14), input current $i(t)$ is given by convolution of $y(t)$ and $v_i(t)$.

$$\begin{aligned} i(t) &= \int_{-\infty}^\infty v_i(\tau)\cdot y(t-\tau)d\tau = \int_{-\infty}^\infty v_i(t-\tau)\cdot y(\tau)d\tau \\ &= \int_{-\infty}^\infty \left\{\sum_{n=0}^\infty \epsilon_n a_n\cos n\omega_s(t-\tau) + \epsilon_n b_n \sin\omega_S(t-\tau)\right\} \\ &\times\left\{\sum_{m=1}^\infty \frac{2}{\pi} r_m \frac{\sin\omega_m\tau}{\tau} U(\tau)\right\} d\tau \end{aligned}$$

$$=\sum_{m=1}^{\infty}\frac{2}{\pi}\mathrm{a}_0 r_m\int_0^{\infty}\frac{\sin\omega_m\tau}{\tau}d\tau$$

$$+\sum_{m=1}^{\infty}\sum_{n=1}^{\infty}\frac{4}{\pi}\cdot a_n r_m\int_0^{\infty}\frac{\cos n\omega_S(t-\tau)\sin\omega_m\tau}{\tau}d\tau$$

$$+\sum_{m=1}^{\infty}\sum_{n=1}^{\infty}\frac{4}{\pi}\cdot b_n r_m\int_0^{\infty}\frac{\sin n\omega_S(t-\tau)\sin\omega_m\tau}{\tau}d\tau \tag{18}$$

The integration of the first term in Eq. (18) becomes $\frac{\pi}{2}$, the second and the third terms' reduce to zero under the condition of Eq. (11).

Eq. (18) results as follows:

$$i(t)=a_0\sum_{m=1}^{\infty}r_m = a_0G(0)\geqq 0 \tag{19}$$

From Eq. (19), it may be proved that the condition of Eqs. (11), (12) is the sufficient condition of continuous input current flow of this case.

3-1-2. Variable conduction interval

In the case of variable conduction interval T_1 of switching devices, the average value of input voltage $v_i(t)$ accross terminals $\boldsymbol{a}-\boldsymbol{a}'$ in Fig. 1 varies with time according to input signal voltage $e_S(t)$ of the switched mode amplifier. By way of example, if the variance is sinusoidal, variable average value $s(t)$ of input voltage $v_i(t)$ is given as follows:

$$s(t)=1+M\sin\alpha t \qquad 0\leqq M\leqq 1 \tag{20}$$

where $s(t)$ is represented with relative amplitude and M is relative amplitude of fractuation. Under assumption that the angular frequency α of fractuation is so low that frequency components of modulation product is negligibly decrease in the pass band of $G(\omega)$**.

Input current into the output filter is expressed as the convolution of $y(t)$ and $s(t)$.

$$i(t)=\sum_{m=1}^{\infty}\int_{-\infty}^{\infty}\{1+M\sin\alpha(t-\tau)\}\frac{2}{\pi}\cdot r_m\cdot\frac{\sin\omega_m\tau}{\tau}U(\tau)d\tau \tag{21}$$

Non-negativity of the right hand side of Eq. (21) is shown as follows:

$$i(t)=\sum_{m=1}^{\infty}\int_{-\infty}^{\infty}\{1+M\sin\alpha(\mathrm{t}-\tau)\}\frac{2}{\pi}\cdot r_m\cdot\frac{\sin\omega_m\tau}{\tau}\cdot U(\tau)d\tau$$

$$=\sum_{m=1}^{\infty}\int_0^{\infty}\{1+M\sin\alpha(t-\tau)\}\frac{2}{\pi}\cdot r_m\cdot\frac{\sin\omega_m\tau}{\tau}d\tau$$

$$\geqq\sum_{m=1}^{\infty}\frac{2}{\pi}\cdot r_m\cdot(1-M)\int_0^{\infty}\frac{\sin\omega_m\tau}{\tau}d\tau$$

** The assumption is satisfied when the angular frequency ω_s of switching repetition is chosen about 14 times of α.

$$\geqq(1-M)\cdot\sum_{m=1}^{\infty}r_m=(1-M)\cdot G(0)\geqq 0 \tag{22}$$

From Eq. (22), even in the case of variable conduction interval, the conditions of Eqs. (11), (12) can hold true for the continuance of input current of the filter. When the fractuation is not sinusoidal, input current may not be always continuous, but the input voltage $v_i(t)$ of the filter can be regarded as periodic function, since input signal voltage $e_S(t)$ of the switched mode amplifier is limited in low frequency range compared with switching frequency $\frac{\omega_S}{2\pi}$, and input current $i(t)$ continues in most cases.

3-2. Reactance filter

If the output filter is constructed with reactance components, power is dissipated only in the load resistance in the output circuit. This kind of filter is called reactance filter being commonly used for the output filter of the switched mode amplifier. In this case, input conductance $G(\omega)$ is connected in simple relation with the transfer function $S(j\omega)$ of the output circuit, and Eqs. (11), (12) are expressed by means of transfer function $S(j\omega)$. From the difinition of $S(j\omega)$,

$$S(j\omega)=V_0(j\omega)\ /\ V_i(j\omega) \tag{23}$$

where $V_0(j\omega)$ indicates Fourier transform of the output voltage $v_0(t)$ accross load resistance. The power of the frequency components $\frac{\omega}{2\pi}$ flowing into the filter circuit is represented in next Eq. (24).

$$w_i=G(\omega)\cdot|V_i(j\omega)|^2 \tag{24}$$

In the output circuit, power is dissipated only in the load resistance R_L and it is given as follows:

$$w_L=|V_0(j\omega)|^2\ /\ R_L \tag{25}$$

From equating w_i and w_L,

$$G(\omega)=|V_0(j\omega|^2\ /\ (|V_i(j\omega)|^2\cdot R_L)=|S(j\omega)|^2\ /\ R_L \tag{26}$$

Using Eq. (26), condition of Eqs. (11), (12) is rewritten as follows:

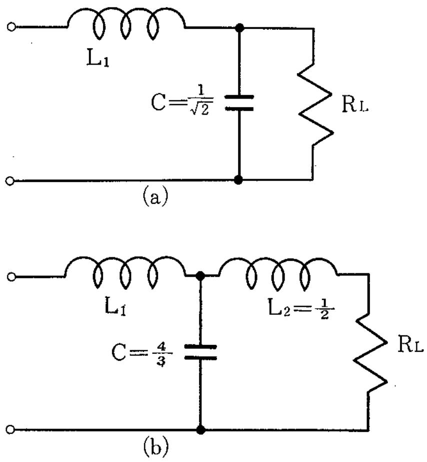
$$|S(j\omega)|^2\ /\ R_L\quad\begin{cases}\geqq 0 & \omega\ll\omega_s\\ \approx 0 & \text{otherwise}\end{cases}\quad\omega\geqq 0 \tag{27}$$

$$\frac{d}{d\omega}\,|S(j\omega)|^2\ /\ R_L\leqq 0 \tag{28}$$

Conditions of Eqs. (27), (28) is easy to judge and the most desirable, since the designing conditions wanted previously may be satisfied such that the pass band of $S(j\omega)$ can be flat, decreases undesirable frequency components caused by time modulation and the filter circuit can be constructed with choke input as well as small power loss and continuous input current flow. Moreover, Butterworth filter, which is designed to have its cut-off angular frequency ω_{CB} sufficiently low band compared with ω_S is located on the boundary of Eqs. (27), (28). Above things suggest Butterworth filter is had only to aim at in designing the output filter of the switched mode amplifier.

4. Experiments

Experiments were performed with the switched mode linear amplifier using sinusoidal a.c. anode source voltage in Fig. 1. The filter used in these experiments are single and double $L-C$ reactance filter shown in Fig. 4 (a) and (b). In the case of reactance filter, it is convenient to make Butterworth filter a standard since it is on the boundary of Eq. (28) each value of filter components is indicated by the value which is normalized to cut-off angular frequency ω_{CB} and load resistance R_{LB} of Butterworth filter. In experiments, allowable region of variable point (L_1, R_L) is examined and shown.



(a) single *L-C* filter circuit
(b) double *L-C* filter circuit
Fig. 4 Filter circuits used in experiments

4-1. Constant conduction interval

The hatched part in Fig. 5 represents the region in which the point (L_1, R_L) of the single $L-C$ filter shown in Fig. 4 (a) is satisfied with Eq. (28). In these experiments, boundary points of continuous input current flow are examined when conduction interval T_1 is time invariant, namely the input signal voltage $e_s(t)$ of the switched mode amplifier is constant.

○, □, △, ☆ indicate boundary points of continuous input current in cases that conduction intervals $T_1=\frac{1}{6}\frac{\pi}{\omega_s},\ \frac{2}{3}\cdot\frac{\pi}{\omega_s},\ \frac{\pi}{\omega_s},\ \frac{3}{2}\cdot\frac{\pi}{\omega_s}$ (sec.), where the cut-off angular frequency ω_{CB} of standard Butterworth filter is chosen as $\frac{1}{7}\omega_s$. For the load resistance R_L above these points, input current did not continue.

●, ■, ▲, ★ indicate similar points of above, but the cut-off angular frequency ω_{CB} equals $\frac{1}{10}\omega_s$.

The hatched part in Fig. (6) represents the region in which the point (L_1, R_L) of double $L-C$ filter shown in Fig. 4 (b) is satisfied with condition in Eq. (28).

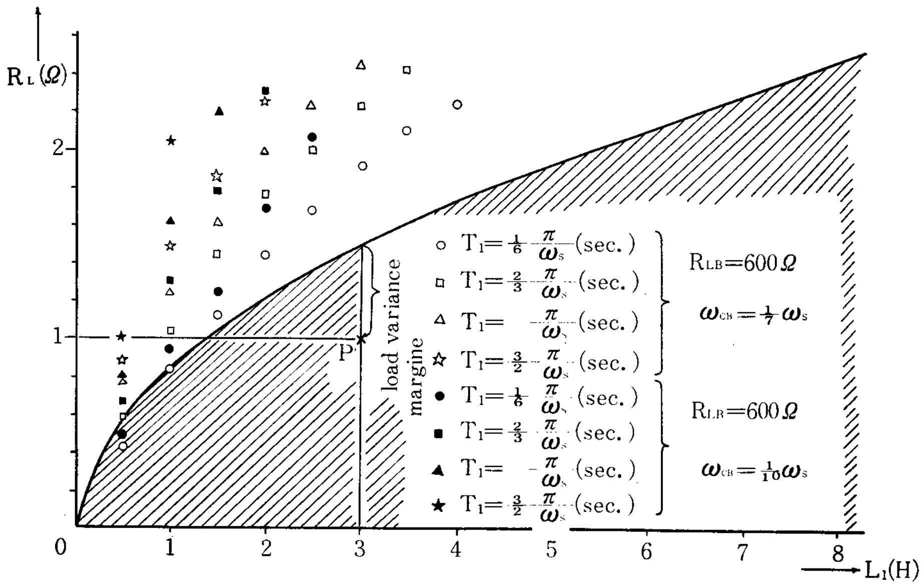


Fig. 5 Allowable range of (L_1, R_L) for correct operation of single L-C filter (case of constant conduction interval) (values of R_L, L_1 are normalized to R_{LB} and ω_{CB}).

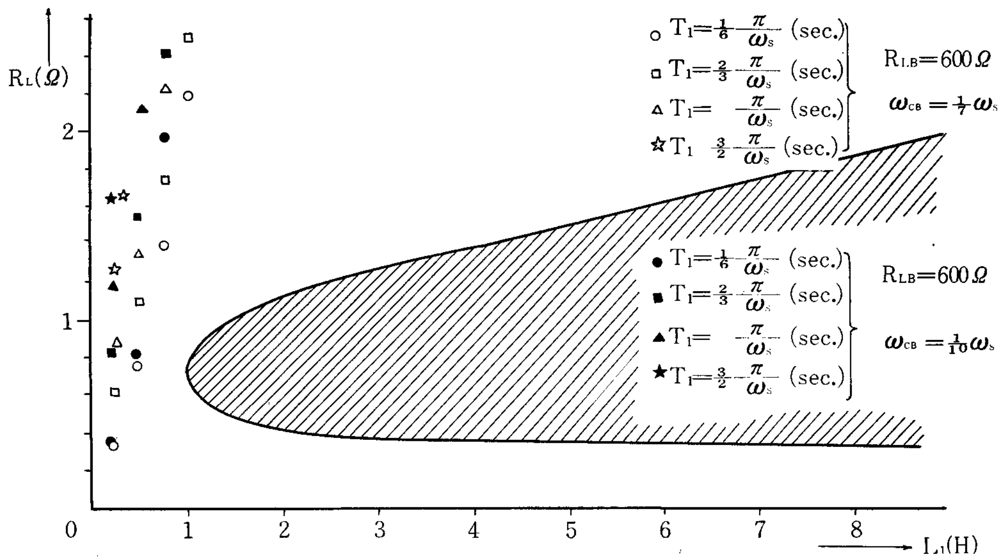


Fig. 6 Allowable range of (L_1, R_L) for correct operation of double L-C filter (case of constant conduction interval) (values of R_L, L_1 are normalized to R_{LB} and ω_{CB})

○, □, △, ☆ indicate boundary points of continuous input current of double $L-C$ filter in individual cases that conduction interval $T=\frac{1}{6}\cdot\frac{\pi}{\omega_s}$, $\frac{2}{3}\cdot\frac{\pi}{\omega_s}$, $\frac{\pi}{\omega_s}$, $\frac{3}{2}\cdot\frac{\pi}{\omega_s}$ and ω_{CB} is chosen as $\frac{1}{7}\omega_s$. ●, ■, ▲, ★ indicate similar points of above without $\omega_{CB}=\frac{1}{10}\omega_s$.

4-2. Variable conduction interval

The experiments are performed in this case that conduction interval is time variant according to input signal voltage $e_s(t)$ of the switched mode amplifier which varies sinusoidally with angular frequency α coincident to ω_{CB}. T_{max} represents the maximum conduction interval of T_1, and it corresponds to the maximum (or minimum) amplitude of input signal voltage $e_s(t)=M'\sin\alpha t$ and switch S_1 and S_2 may start to conduct in time interval $\left[\frac{2\pi}{\omega_s}-T_{max} : T_{max}-\frac{\pi}{\omega_s}\right]$.

In Fig. 7, ○●, □■, △▲, ☆★ indicate boundary points of continuous input current of the single $L-C$ filter circuit in Fig. 4 (a) in individual cases that the maximum conduction interval $T_{max}=2\cdot\frac{\pi}{\omega_s}$, $1.8\frac{\pi}{\omega_s}$, $1.7\cdot\frac{\pi}{\omega_s}$, $1.6\frac{\pi}{\omega_s}$ (sec.) and input current became discontinuous for the load resistance R_L above these points, where white and black points correspond to $\omega_{CB}=\frac{1}{15}\omega_s$ and $\omega_{CB}=\frac{1}{20}\omega_s$ respectively.

In Fig. 8, boundary points ○●, □■, △▲, ☆★ of the double $L-C$ filter circuit in Fig. 4 (b) are shown in the similar manner of Fig. 7.

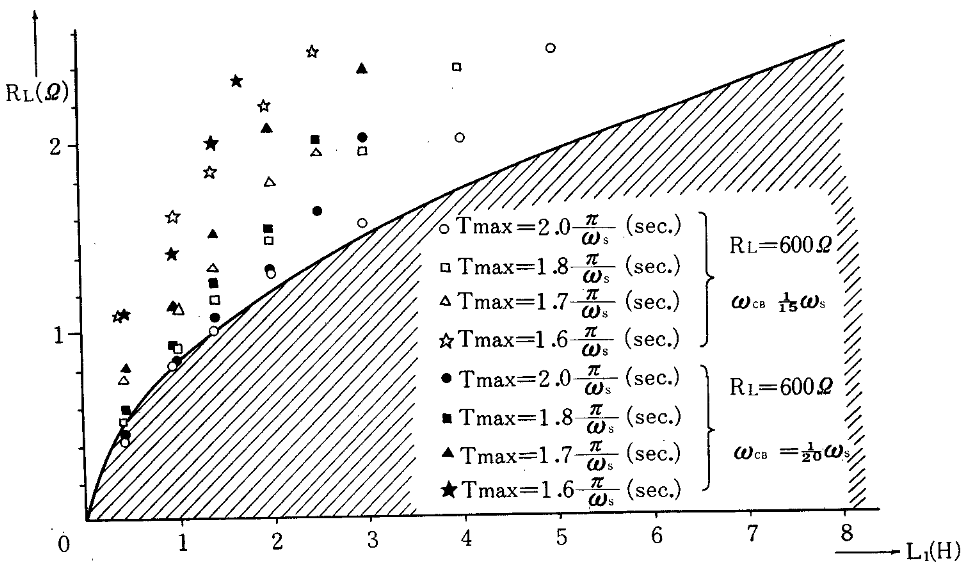


Fig. 7 Allowable range of (L_1, R_L) for correct operation of single L-C filter (case of variable conduction interval) (values of R_L, L_1 are normalized to R_{LB} and ω_{CB}).

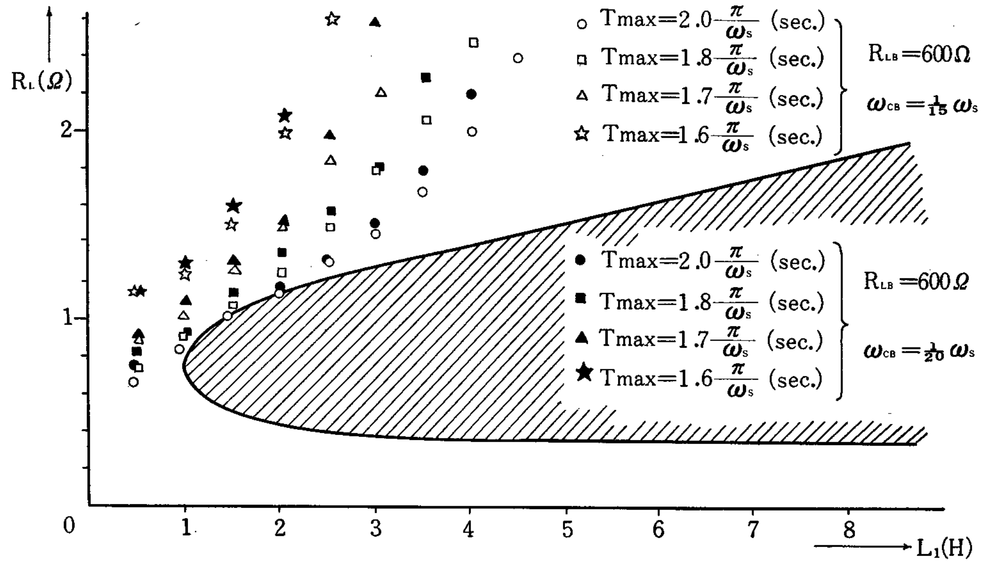


Fig. 8 Allowable range of (L_1, R_L) for correct operation of double *L-C* filter (case of variable conduction interval) (values of R_L, L_1 are normalized to R_{LB} and ω_{CB})

4-3. Experimental results

Results from Fig. 5 to Fig. 8 represent that the conditions of Eqs. (11), (12) are sufficient condition for continuous input current flow since boundary points obtained by experiments are not included in the hatched region in which Eq. (12) is satisfied in each case. Eqs. (11), (12) can serve as the practical condition of continuous input current flow and it is very easy to use.

In addition, for the variable load resistance, continuance condition of input current is satisfied if input choke L_1 is so properly chosen that the operation point of (L_1, R_L) is in the hatched region, for example, if the point (L_1, R_L) is chosen at P (3, 1) as in Fig. 5, the allowable value of the load resistance is given as follows:

$$0 \leqq R_L \leqq 1.5$$

Non-normalized load resitances and input choke are also given.

$$0 \leqq R_{L(\text{non-normalized})} \leqq 1.5 R_{LB}$$

$$L_{1(\text{non-normalized})} = \frac{3R_{LB}}{\omega_{CB}}$$

In each way, Fig. 5 or Fig. 6 can be a design chart even in the case of variable load resistance.

5. Conclusions

It is the necessary and sufficient condition for the continuous current flow into the output filter that the impulse response of the input admittance has to be non-negative provided that the non-negative voltage is applied to it. Especially, for the switched mode amplifier that the input voltage of the output filter is repeated almost periodically, it is sufficient for continuous input current that the input conductance $G(\omega)$ of the filter circuit has narrower pass band compared with angular frequency ω_s of switching repetition and monotonously decrease with respect to ω. And it was confirmed with experimental results.

Generally, the output filters, which is considered to have no power dissipation, are used for switched mode amplifier. When such filters are employed, the transfer function $S(j\omega)$ of the output circuit can be served for judging the continuous current flow condition in place of $G(\omega)$. Butterworth filter can have the characterstics of no power loss in itself and maximally flat response in its pass-band. It is confirmed that well-known Butterworth filter fulfils all conditions for this output filter.

This may be concluded that the output filter of the switched mode amplifier has only to be designed for Butterworth filter with much lower cut-off angular frequency than ω_S. This results make it very easy to design the output filter of the switched mode amplifier which has been difficult because of the continuance condition of input current flow.

Moreover, the results derived without limitation to the input voltage waveform of the filter may be applied not only to the switched mode amplifier but to the phase controlled rectifier, d.c. chopper circuits and other switched mode circuit.

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