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# APPLICATION OF VERTICAL FET FOR PULSE WIDTH MODULATION AUDIO POWER AMPLIFIER

by

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## Introduction

Pulse width modulation (PWM) amplifier due to its theoretically high efficiency of operation, has a promising future for high audio power amplification application in small and light package. Vertical FET (V-FET) developed for audio power amplification has such suitable characteristics for PWM application as wide area of safe operation (ASO), excellent high frequency response and good pulse response due to the absence of storage time in signal transfer.

This paper reports on a design of PWM amplifier with V-FET and analyzes its performance characteristics.

1. PWM amplifier, its theory of operation and characteristics:
   Fig. 1 shows the basic block diagram of PWM amplifier, in which the amplitude of input signal is converted to duration time of pulse by amplitude-time converting circuit. The converted pulse signal then controls the switching elements of the final power stage, the output of which is passed through a low-pass filter to the load. Since the final power stage is either in saturated ON mode or totally OFF mode, there is, theoretically no loss of energy in operation. Selection of the amplifier device can be done only from the stand point of its performance characteristics as switching device. On the other hand compared to conventional straight linear amplifier, the PWM amplifier requires an additional amplitude-time conversion circuit and low-pass network. Also, since it handles square wave signals, suppression of unnecessary RF radiation becomes necessary to make the amplifier commercially practical.

2. Construction of PWM amplifier
   Fig. 2 shows block diagram and waveform at each stage of the prototype PWM amplifier. A square wave generator is employed as the source of carrier. The carrier is converted to triangular wave through an intergrator circuit. This triangular shaped carrier and the audio input signal are added and then passed through a saturating high-gain amplifier to obtain square wave signal duration time of which is in direct proportion to the amplitude of the input audio signal. This signal is then amplified by a pulse power amplification stage and demodulated by a low pass filter. As shown the higher carrier frequency, the more

suppression of the carrier and sideband components is obtained with a given high-cut slope of the low-pass filter.

3. Amplitude-time converter
Fig. 3 shows the basic construction of the amplitude-time converter. In this case in order to return negative feedback signal to the audio input circuit the pulse width modulated output is put through a phase inverter and an intergrator circuit which increases frequency response by 6dB/oct. To compensate for this increase in frequency response and delay of phase by 90 degrees in the negative feedback network the input signal has to be passed through an intergrator circuit also. Since $\int A dt + \int B dt + \int C dt$ equals $\int (A+B+C)\, dt$ Fig. 3 can be reconstructed as Fig. 4. The entire block diagram of the PWM power amplifier is shown in Fig. 5. In order to lock the pulse width modulated signal with the carrier frequency the following condition has to be met: $I_c > I_s + I_F$

When the modulation is at 100% Is+IF becomes maximum and Is will be equal to IF. Therefore, it is necessary to satisfy the following condition to lock the signal with carrier up to 100% modulation: $I_c > 2\, I_F$

4. Amount of negative feedback
With RF removed from the diagram in Fig. 5 to make the amplifier open looped, the amplitude (Vi) of triangular shaped carrier obtained from square wave EC passed through an intergrator network is as follows: $V_i = \frac{1}{C}\int_0^{\frac{T}{2}} I_c\, dt = \frac{T\, I_c}{2C}$

Since Ic equals 2 IF to satisfy locking condition, this equation can be expressed as follows: $V_i = \frac{T V_1}{C R_F}$

The gain of this amplifier is the ratio between the potential of output pulse and amplitude of triangular carrier as shown below: $\frac{E_o}{E_i} = \frac{V_1}{\frac{T V_1}{C R_F}} = \frac{C R_F}{T}$ ------ (1)

Here if Es is put as Vs sin$\omega$t, the corresponding Ei is expressed as follows: $E_i = \frac{1}{C}\int I_s dt = \frac{V_s}{C R_s}\int \sin \omega t \cdot dt = \frac{V_s}{\omega C R_s} \cos \omega t$ --(2)

Therefore, the gain from ES to Ei is: $\frac{E_i}{E_s} = \frac{\frac{V_s}{\omega C R_s}}{V_s} = \frac{1}{\omega C R_s}$ ------- (3)

From the equations (1) and (2), the open loop gain of this amplifier is: $\frac{E_o}{E_s} = \frac{E_i}{E_s} \cdot \frac{E_o}{E_i} = \frac{1}{\omega C R_s} \cdot \frac{C R_F}{T} = \frac{R_F}{\omega R_s T} = \frac{1}{\omega T} \cdot \frac{R_F}{R_s}$ ---(4)

Here, since $\frac{R_F}{R_s}$ is the gain of this amplifier in closed loop condition, the amount of negative feedback $A_{NF}$ is obtained as follows: $A_{NF} = \frac{1}{\omega T}$

As shown in Fig. 5 the carrier frequency (fc) equals $\frac{1}{2T}$
And the input signal frequency (fs) equals $2\pi\omega$. Therefore:

$$A_{NF} = \frac{1}{\pi} \cdot \frac{f_c}{f_s} \quad ----(5)$$

From this the relationship between ANF and input signal frequency is as shown in Fig. 6. In this case the carrier frequency (Fc) is set at 500kHz.

5. V-FET as used for switching device

The carrier frequency for PWM amplifier should be as high as possible for the following reasons:

1. Higher carrier frequency allows higher suppression of the carrier and sideband components at the output with a given low-pass filter.
2. As shown in the equation (5), the higher the carrier frequency, the more amount of negative feedback for lower distortion.

However, as the carrier frequency is increased high speed switching capability of the device becomes a problem. High frequency carrier requires devices with good high frequency characteristics and high power capability. Since V-FET does not show storage time in pulse response and has excellent rise and fall time characteristics, it is especially suitable for this PWM application. Also, since both N- and P- channel devices are avialable, driving circuit for the V-FET final stage can be simple without requiring a phase inverter circuit.

Fig. 7 shows static characteristics of the V-FET's (at VGS=0). They show good linearity from forward range to reverse range, which means in equivalent circuit they can be regarded as pure resistive components having their saturation resistance (Ron) as the value when they are on. This allows dissipation within the V-FET of the return current generated at the inductive component in the low-pass network in the output circuit.

Fig. 8 shows the relationship between the switching waveform and the drain current waveform at no input signal. Fig. 9 also shows waveforms at zero input.

6. Operational trace at DC output

Fig. 10 and Fig. 11 (a) (b) show operation and waveforms of the PWM amplifier when it is supplying DC output (Vo) into the load (RL). Since Vo is the mean value during the period 2T, it is expressed as follows:

$$V_0 = \frac{1}{2T}\left\{V_1\,(T+t) - V_1\,(T-t)\right\} = \frac{t}{T}V_1 \quad ---(6)$$

By putting $\frac{t}{T}$ as $\alpha$ (degree of modulation), the equation (6) becomes $V_o = \alpha V_i$.

Therefore: $I_o = \frac{V_o}{RL} = \frac{\alpha V_i}{RL}$ ---- (7)

In this case the ripple current which flows in the low-pass inductor (L) is shown as follows (LI=ET):

$$\Delta I_o = \frac{1}{L}(V_i - V_o)(T + t) = \frac{V_i T}{L}(1 - \alpha^2) \text{ ---(8)}$$

7. Loss in switching

With the waveform as shown in Fig. 12 the loss during rise time (tr) and fall time (tf) is obtained as follows, assuming the rate of voltage change stays constant.

7,1. During tf

Since the current during this period is $I_o + \frac{\Delta I_o}{2}$ it is derived from the equations (7) and (8) and expressed as follows:

$$I_f = \frac{\alpha V_i}{RL} + \frac{V_i T}{2L}(1 - \alpha^2) \text{ ------ (9)}$$

As the change in voltage is $2V_i$, and its rate of change stays constant, the loss during fall time (Pf) is expressed as follows:

$$P_f = \frac{t_f}{2T} \cdot \frac{2V_i}{2} \cdot I_f = \frac{t_f V_i^2}{2T}\left\{\frac{\alpha}{RL} + \frac{T}{2L}(1 - \alpha^2)\right\} \text{---(10)}$$

Since the above equation (10) shows the Pf at a constant $\alpha$, it can be changed to the following equation when $\alpha$ becomes $\alpha_1 \sin X$ with sinusoidal signal. It is expressed as the mean amount of loss during the period when X=0 and $X = \frac{\pi}{2}$ :

$$P_{sf} = \frac{t_f V_i^2}{2T} \cdot \frac{2}{\pi}\int_0^{\frac{\pi}{2}}\left\{\frac{\alpha_1 \sin X}{R_L} + \frac{T}{2L}(1 - \alpha_1^2 \sin^2 X)\right\}dX = \frac{t_f V_i^2}{4\pi}\left\{\frac{4\alpha_1}{\pi R_L} + \frac{T}{L}\left(1 - \frac{\alpha_1^2}{2}\right)\right\} \text{ .....(11)}$$

7,2. During tr

The current during this period is $I_o - \frac{\Delta I_o}{2}$ . Therefore, from the equations (7) and (8), the following is obtained:

$$I_r = \frac{\alpha V_i}{RL} - \frac{V_i T}{2L}(1 - \alpha^2) \text{ ---- (12)}$$

Ir changes its direction as $\alpha$ varies from 0 to 1. The point where Ir=0 is obtained as follows:

$$0 = \frac{V_i T}{2L}\alpha^2 + \frac{V_i}{RL}\alpha - \frac{V_i T}{2L}$$

The positive value of $\alpha$ of the above equation is:

$$\alpha_o = \sqrt{\left(\frac{L}{R_L T}\right)^2 + 1} - \frac{L}{R_L T} \quad ----(13)$$

The loss during tr period is expressed as follows:

$$P_r = \frac{t_r V_1^2}{2T} \cdot \left| \frac{\alpha}{R_L} - \frac{T}{2L}(1-\alpha^2) \right|$$

For the sinusoidal signal, when $0 \leq \alpha_1 \leq \alpha_o$,

$$P_{sr} = \frac{t_r V_1^2}{2T} \cdot \frac{2}{\pi} \int_0^{\frac{\pi}{2}} \left\{ -\frac{\alpha_1 \sin X}{R_L} + \frac{T}{2L}(1-\alpha_1^2 \sin^2 X) \right\} dX = \frac{t_r V_1^2}{4T} \left\{ -\frac{4\alpha_1}{\pi R_L} + \frac{T}{L}\left(1-\frac{\alpha_1^2}{2}\right) \right\} \quad ---- (14)$$

when $\alpha_o \leq \alpha_1 \leq 1$, where $\sin^{-1}\frac{\alpha_o}{\alpha_1} = \theta_o$

$$P_{sr} = \frac{t_r V_1^2}{2T} \cdot \frac{2}{\pi} \left[ \int_0^{\theta_o} \left\{ -\frac{\alpha_1 \sin X}{R_L} + \frac{T}{2L}(1-\alpha_1^2 \sin^2 X) \right\} dX \right.$$

$$= \frac{t_r V_1^2}{\pi T R_L} \left[ \alpha_1 (2\cos\theta_o - 1) + \frac{T R_L}{2L} \left\{ \left(2\theta_o - \frac{\pi}{2}\right) - \frac{\alpha_1^2}{2}\left(2\theta_o - \frac{\pi}{2} - \sin 2\theta_o\right) \right\} \right] \cdot (15)$$

From the equation (7), the relationship between output current Iso and $\alpha$ (where $\alpha = \alpha_1 \sin X$ ) is expressed as follows:

$$I_{so} = \frac{\alpha_1 V_1}{R_l} \sin x$$

The effective output power in this case is obtained as follows:

$$P_{so} = \left( \frac{\alpha_1 V_1}{\sqrt{2} R_L} \right)^2 \cdot R_L = \frac{\alpha_1^2 V_1^2}{2R_L} \quad ----(16)$$

8. Loss in the saturation resistance (Ron)

In the resistive component in the switching element the following loss is generated due to the current as shown in Fig. 11 (b) when the output is Vo. In this case reduction in supply voltage at Ron is considerably smaller than $V_1$ and can be ignored. When the output waveform is $+V_1$ and $-V_1$, the loss is expressed as follows: (although the waveform of ripple current is as shown in Fig. 11(b), its integral value is equal to the waveform in Fig. 11(c).

$$I = I_o - \frac{\Delta I_o}{2} + \Delta I_o \cdot \frac{t}{2T} \quad ---- (17)$$

From the equations (7) and (8) ...

$$I = V_1 \left\{ \frac{1-\alpha^2}{2L}(t-T) + \frac{\alpha}{R_L} \right\} \quad ----- (18)$$

Therefore, the loss (Pr) is obtained:

$$P_R = \frac{1}{2T}\int_0^{2T} Ron \cdot I^2 dt = \frac{V_1^2 Ron}{2T}\int_0^{2T}\left\{ \frac{1-\alpha^2}{2L}(t-T) + \frac{\alpha}{RL}\right\}^2 dt$$
$$= V_1^2 Ron\left\{\frac{1}{3}\left(\frac{1-\alpha^2}{2L}\right)^2 T^2 + \left(\frac{\alpha}{RL}\right)^2\right\} \quad ----(19)$$

From this the loss under sinusoidal signal (Psr) is obtained by putting $\alpha$ as $\alpha_1 \sin X$ and taking mean value during 0 and $\frac{\pi}{2}$ period:

$$P_{SR} = V_1^2 Ron \cdot \frac{2}{\pi}\int_0^{\frac{\pi}{2}}\left\{ \frac{1}{3}\left(\frac{1-\alpha_1^2 \sin^2 x}{2L}\right)^2 T^2 + \frac{\alpha_1^2 \sin^2 x}{RL^2}\right\} dx$$
$$= \frac{V_1^2 Ron\, T^2}{L^2}\left\{\frac{1}{6\pi} + \alpha_1^2\left(\frac{L^2}{2RL^2 T^2} - \frac{5}{96}\right)\right\} \quad ----(20)$$

9. Comparison of amount of loss between class B amplifier and PWD amplifier.

The loss in a class B amplifier as shown in Fig. 13 is expressed as follows:

$$P_c = \frac{2V_1}{\pi}\sqrt{\frac{2P_o}{RL}} - P_o \quad ----(21)$$

where Po is effective power under sinusoidal signal.

Also the maximum power before clipping is obtained as follows:

$$P_{OM} = \frac{V_1^2}{2RL} \quad ----(22)$$

With the equations (21) and (22) Fig. 14 is drawn to show the relationship between Po/PoM and Pc/PsoM in which PsoM, equation (23) below, is the maximum non-clipping output of PWM amplifier with saturation resistance Ron and Psc is the loss in total defined as Psc=Psf+Psr+Psr; equations (11), (14), (15), (16) and (20).

$$P_{SOM} = \frac{V_1^2}{2RL} \cdot \frac{RL}{RL+Ron} = \frac{V_1^2}{2(RL+Ron)} \quad --- (23)$$

This equation shows that Pson is the output when $\alpha$ equals 1, cf equation (16), is devided by Ron and RL. Fig. 14 is obtained with the following conditions:

RL=8ohms, L-25$\mu$H, T=1$\mu$sec, tf=tr=0.1$\mu$sec Ron=1ohm

10. Conclusion

The performance characteristic of a prototype PWM amplifier with V-FET final stage are shown in Fig. 15, 16,17. They confirm that it is now practical to make a PWM amplifier for audio application by using the V-FET's. The actual measurement of the channel dissipation is close to the theoretical value as shown in Fig. 17. Also, from Fig. 14 it is possible to see that efficiency of the PWM amplifier is much higher than that of a class B amplifier operating in its ideal conditions in most of the power range. Class B efficiency becomes nearly equal to PWM efficiency only at the maximum output. Since peak factor of the audio signal is somewhere between 1/10 and 1/3, PWM amplifier reduces waste energy considerably.

In closing I wish to give my sincere appreciation to Dr. H. Nakajima and Messrs. M. Nagami and H. Nakamura of Sony Research Laboratory for their advice and guidance and to all of my colleagues who worked with me in developing the PWM amplifier.

FIG. 1

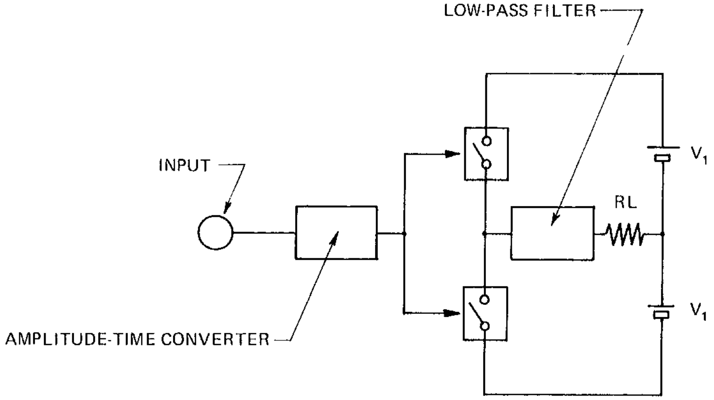


**FIG. 2** BLOCK DIAGRAM OF PWM AMPLIFIER AND WAVEFORMS AT EACH STAGE

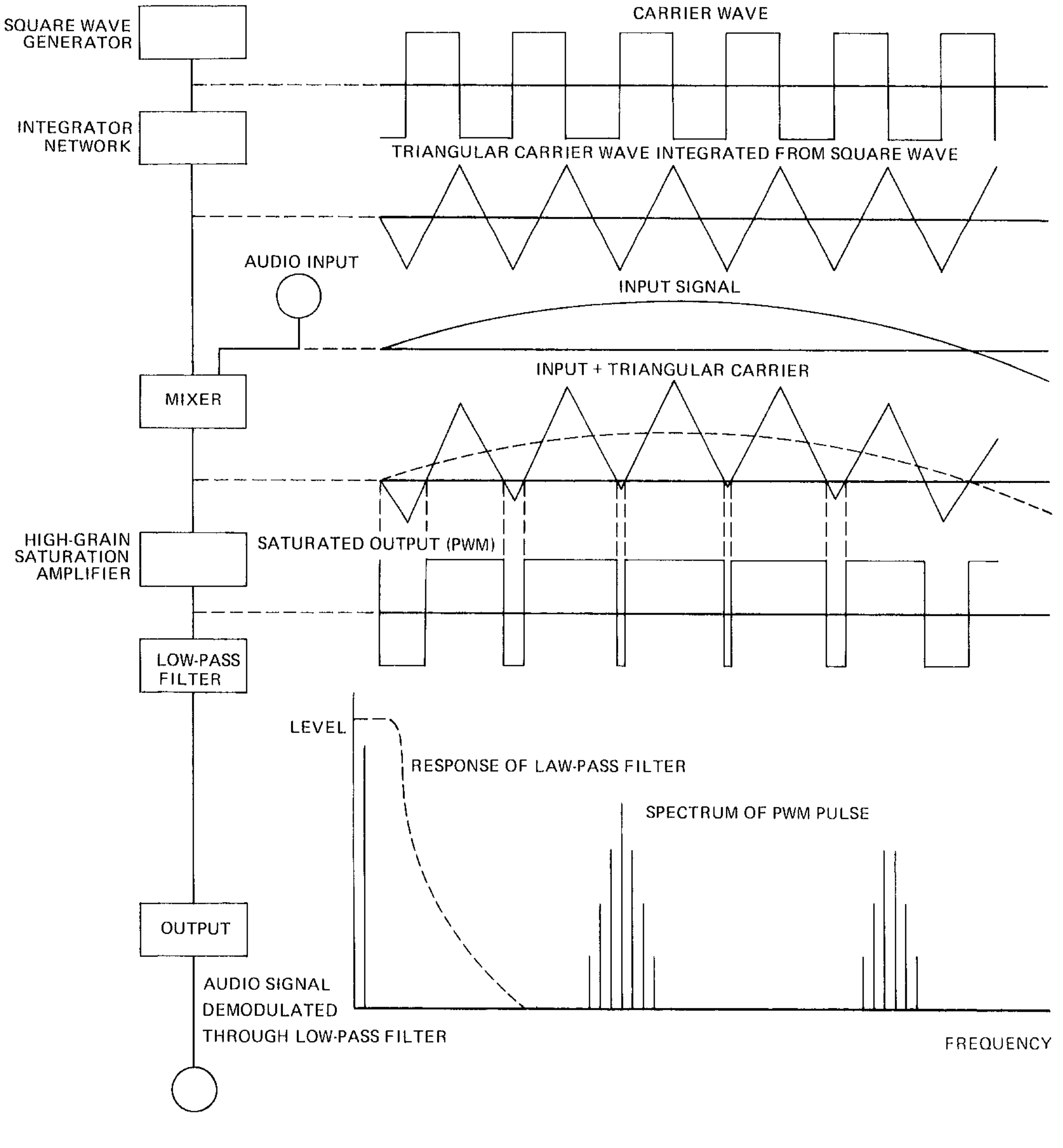


FIG. 3

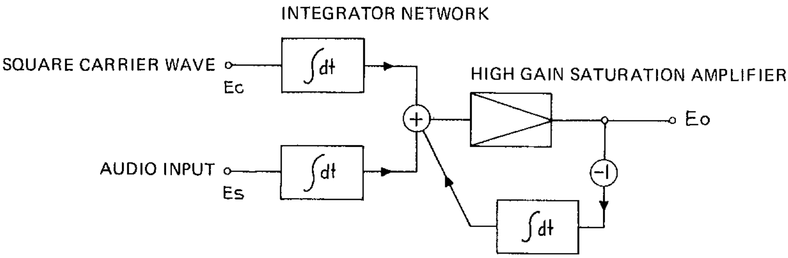



FIG. 4

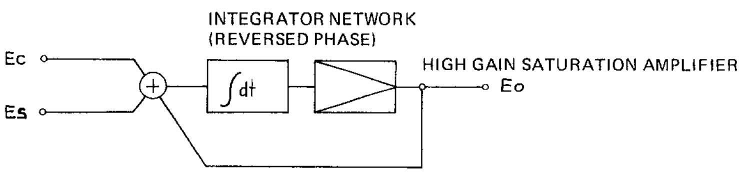



FIG. 5

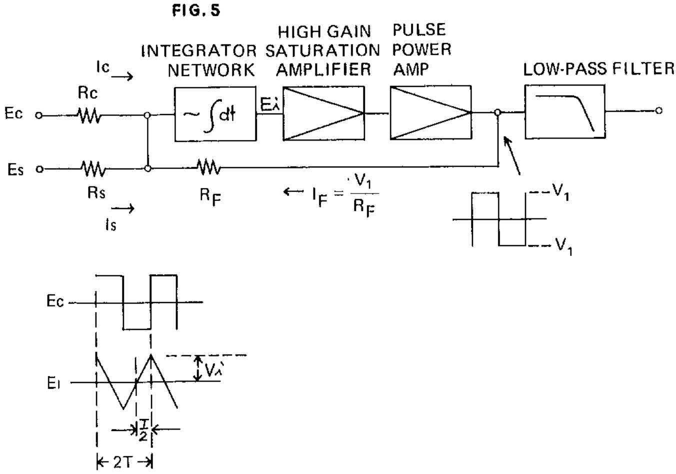


FIG. 6 NEGATIVE FEED BACK VS. FREQUENCY

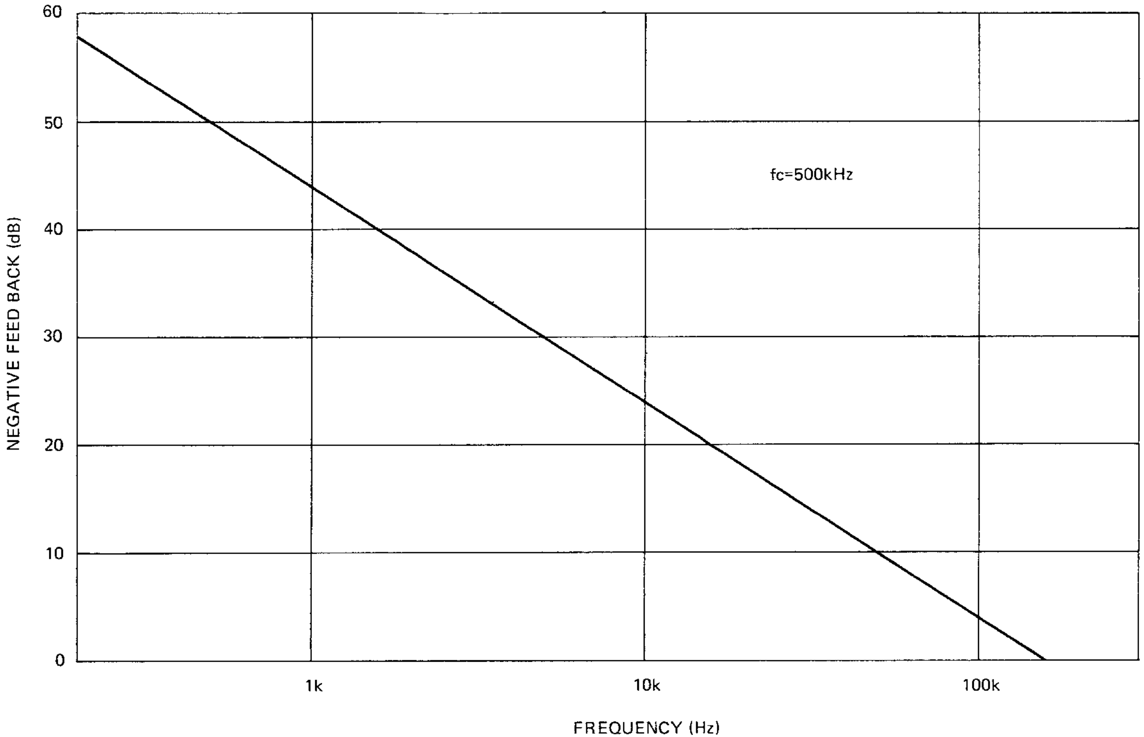
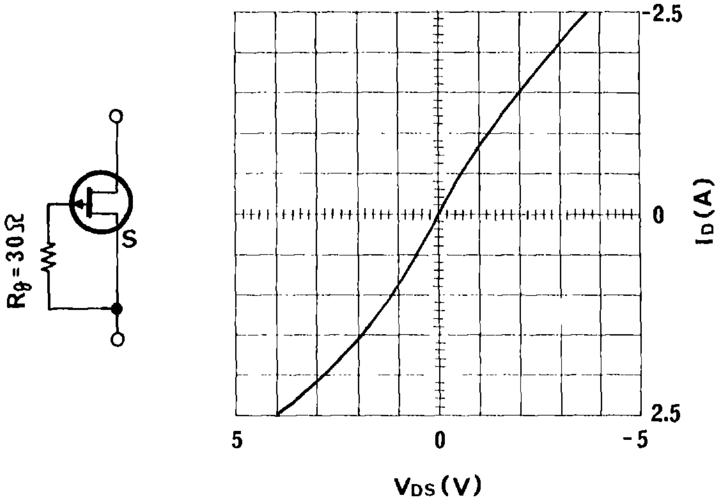


FIG. 7

2SJ28




2SK82

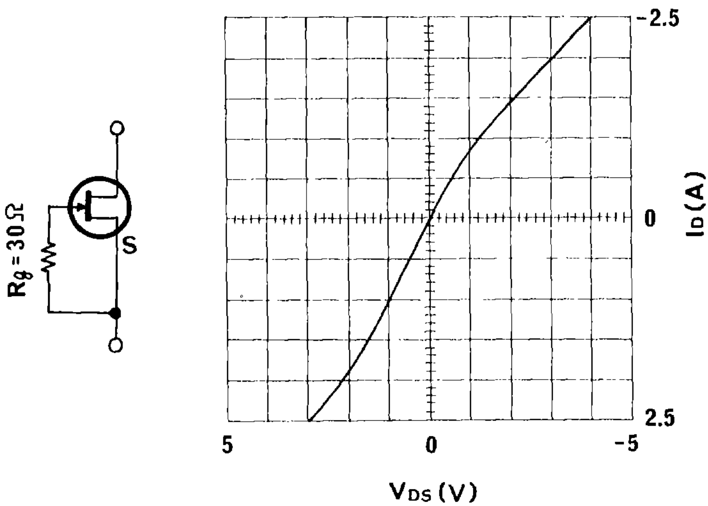
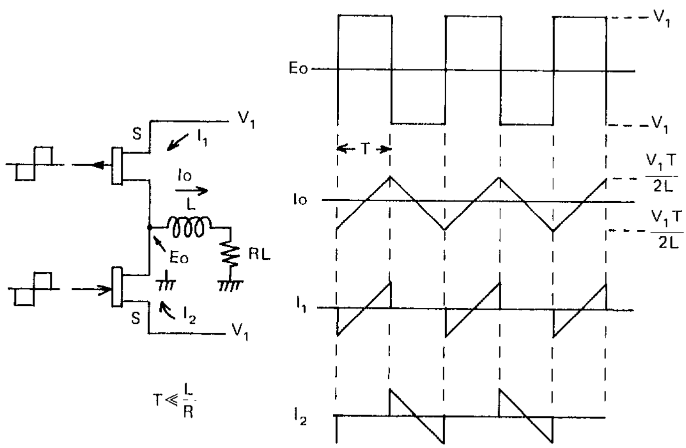
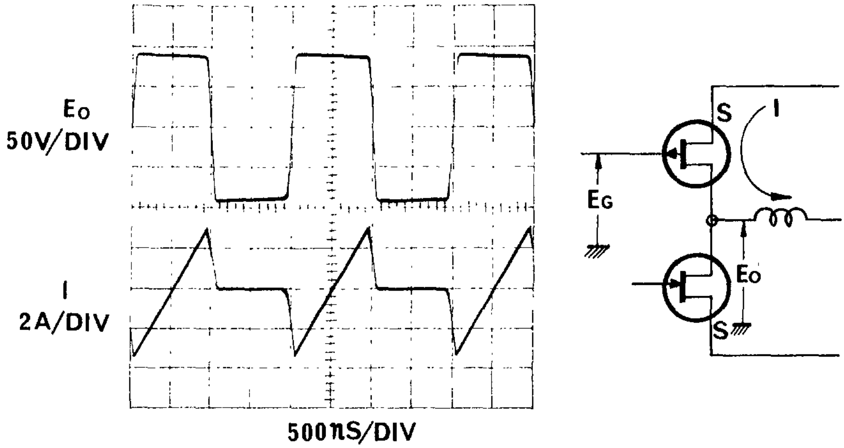


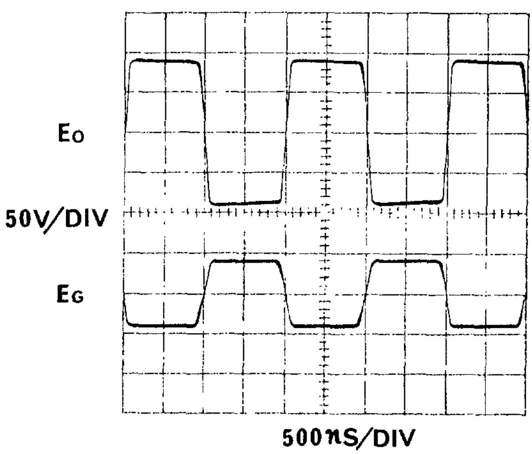
FIG. 8



S
$I_1$
$V_1$
Io
L
Eo
RL
$I_2$
S
$V_1$
$T \ll \frac{L}{R}$
Eo
$V_1$
$V_1$
T
Io
$\frac{V_1 T}{2L}$
$\frac{V_1 T}{2L}$
$I_1$
$I_2$

FIG. 9



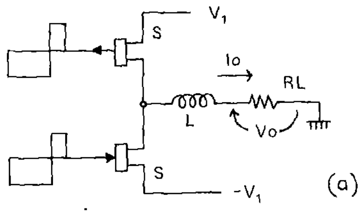



FIG. 10 (a)

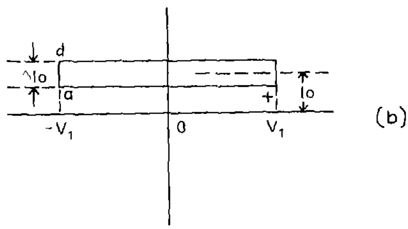



FIG. 10 (b)

FIG. 11

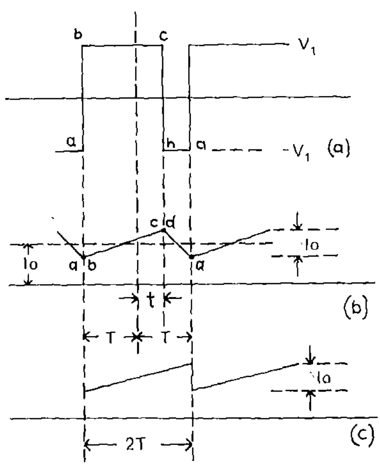


FIG. 12

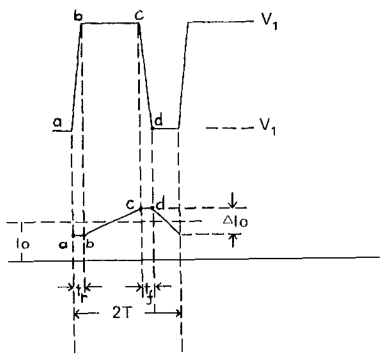



Fig - 12

FIG. 13

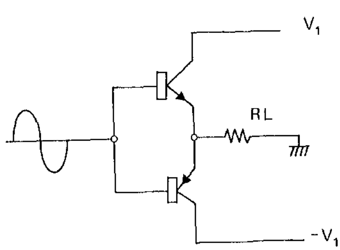



Fig - 13

**FIG. 14** DISSIPATION VS. OUTPUT (NORMALIZED BY MAXIMUM OUTPUT)

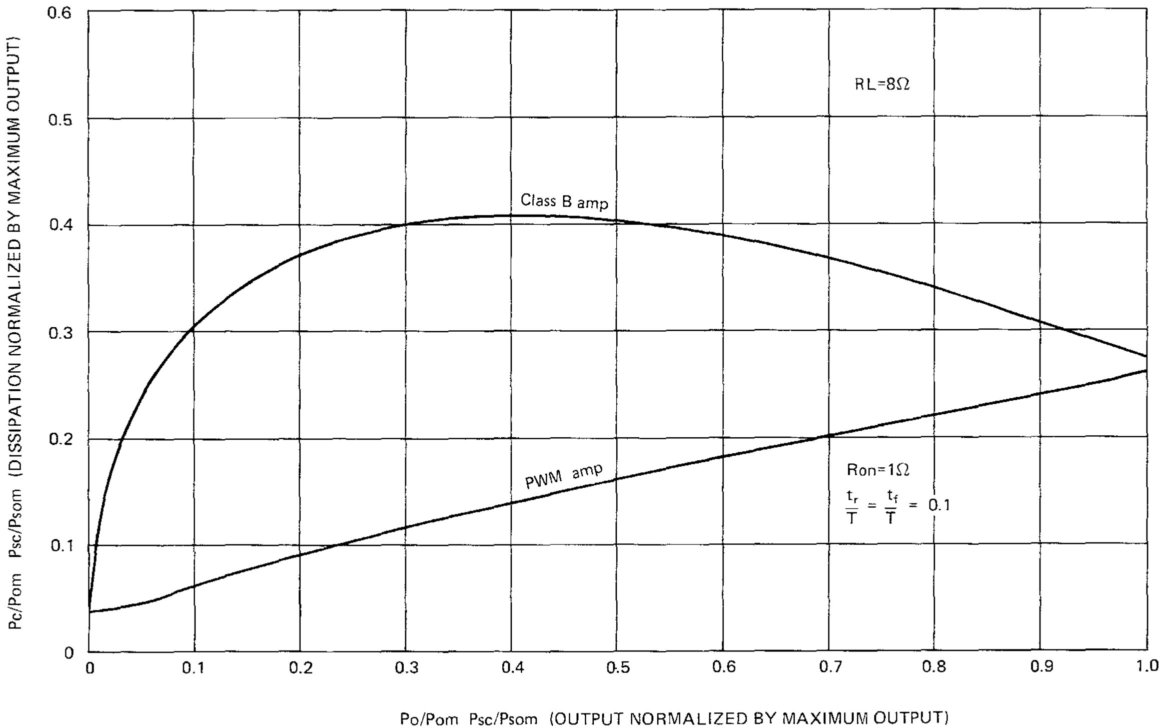


**FIG. 15** FREQUENCY RESPONSE

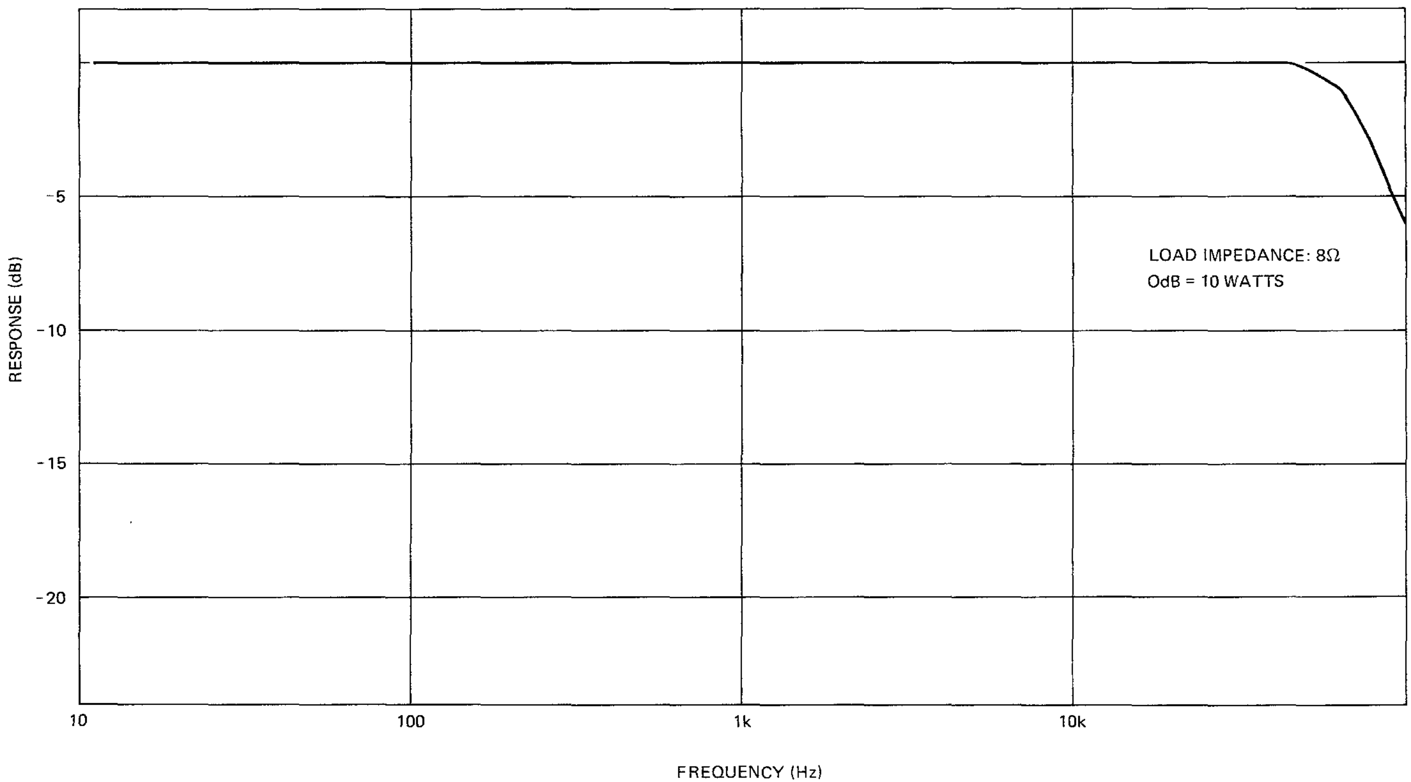


**FIG. 16** HARMONIC DISTORTION VS. OUTPUT

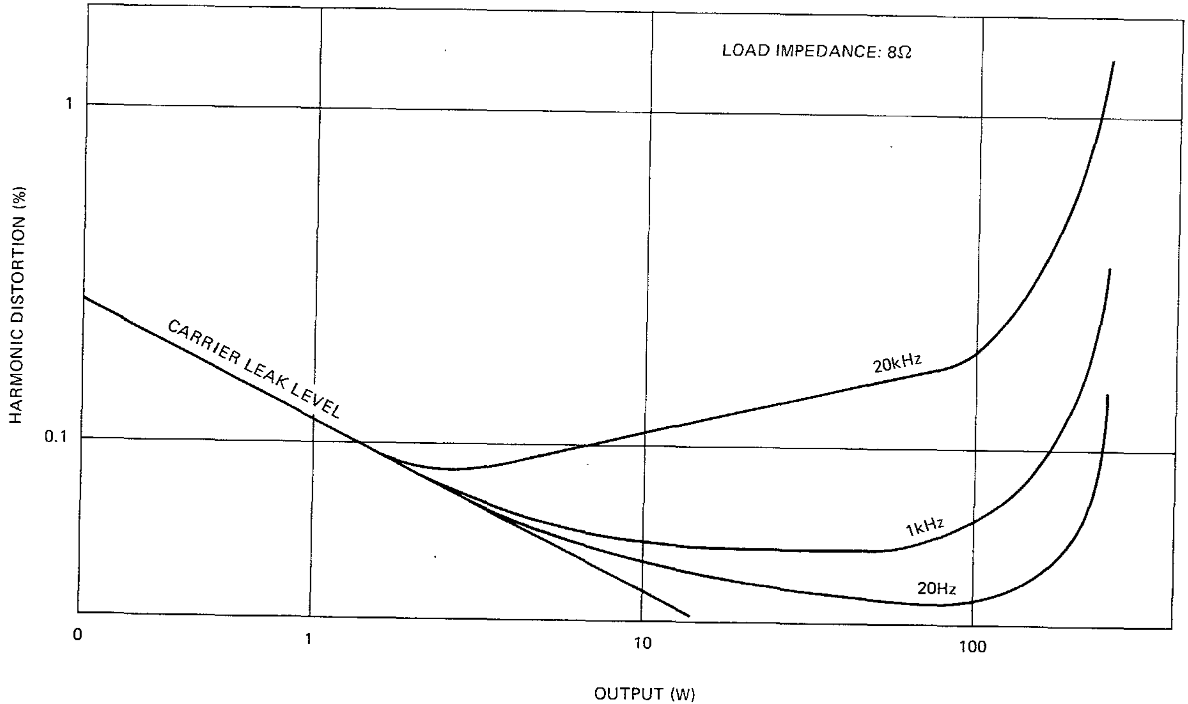


**FIG. 17** CHANNEL DISSIPATION VS. OUTPUT