100 WATT SUPER AUDIO AMPLIFIER USING NEW MOS DEVICES

Tohru Sampei, Shin-ichi Ohashi
Consumer Products Research Center

Shikayuki Ochi
Central Research Laboratory

Hitachi, Ltd.
Tokyo, Japan

ABSTRACT

A 100 watt audio amplifier was recently developed using a new Power MOSFET which was developed by our MOS Device Group. The Power MOSFET has several advantages over bipolar transistors. It has good frequency response, no carrier storage delay, thermal stability, no secondary breakdown and high input impedance.

The MOSFET amplifier delivers a continuous power output of 100 watts at 8 ohms from 5 Hz to 100 kHz with no more than 0.01 per cent total harmonic distortion which is about ten times better than ordinary bipolar transistor amplifiers, and in addition, the high input impedance and thermal stability of the MOSFET reduces the size of total circuit about 30 per cent.

1. INTRODUCTION

At present, bipolar transistors are widely used as power devices of audio amplifiers because in the past they have been the only semi-conductor power devices available. Such devices require a wide Area of Safe Operation for greater reliability and a large gain bandwidth product for large amounts of negative feedback at high frequencies. Bipolar transistors, which have positive temperature coefficients of collector currents, have secondary breakdown, and bipolar transistors which are minority carrier devices lack frequency response and carrier storage delay.

In light of these drawbacks associated with transistors, a MOSFET for power applications is being developed and this research is progressing significantly, especially with regard to high frequency and high speed switching applications.*(1) In audio power applications, Power MOSFET, a majority carrier device, is expected to have several advantages over bipolar transistors:

(1) Good frequency response because of fast carrier speed,
(2) High speed switching due to the absence of minority carrier storage,
(3) Thermal stability and no secondary breakdown because of their drain current, negative temperature coefficient,
(4) High power gain because of high input impedance.

An available Power MOSFET presently has a maximum rating of 90 V, 2 A, but in an audio amplifier which delivers a continuous power output of 100 watts at 8 ohms or 4 ohms, a power device of 140 V, 7 A is needed.

Our MOS Device Group has concentrated on obtaining a greater power handling capability for the MOSFET and developed the p channel MOSFET of 100 V, 20 A.*(2)

A complementary Power MOSFET pair (n channel and p channel, with 160 V, 7 A capacity) has been developed. The breakdown voltage capability is now improved by using an ion implanted offset gate structure.

A 100 watt amplifier with the complementary MOSFET pair has been designed, and it has low distortion, wide power bandwidth, high reliability and simple circuitry. In this paper the responses, switching delays, and total harmonic distortion of source followers and darlington connected emitter followers are compared, and the circuit construction of an amplifier with a complementary MOSFET pair is also described.

Received July 29, 1977

2. FEATURES OF THE POWER MOSFET

Figure 1 shows a cross-section of the Power MOSFET. High breakdown voltage capability is achieved by using an ion implanted offset striped gate structure.

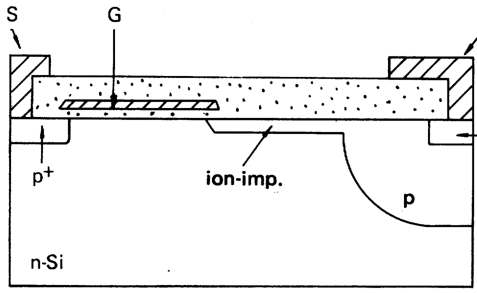
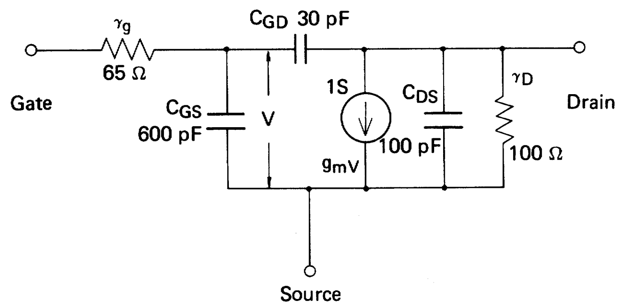


Fig. 1 Cross-section of Power MOSFET

The offset gate structure reduces the electric field around the gate electrode, but the channel is not constructed in the region which is not covered with the gate electrode. Ions then are implanted in this region so this high resistance region divides the voltage between the drain and source.

A capability for handling large currents, and large transconductance are achieved with wide channel width and short channel length. The channel width is 40 cm for the n channel; 50 cm for the p channel. The channel lengths are 9 μm for the n channel, 8 μm for the p channel. The equivalent circuit of the Power MOSFET is shown in Figure 2. In this circuit, parasitic capacitance and leak resistance are neglected. Since the gate is made with poli-silicon, the intrinsic gate resistance is relatively larger than the metal gate structure.

Additional features of the Power MOSFET are summarized in Table 1.



γ_g: intrinsic gate resistance
C_{GS}: gate to source capacitance
g_m: transconductance
C_{GD}: gate to drain (feedback) capacitance
C_{DS}: drain to source capacitance
γ_D: drain output resistance

Fig. 2 Equivalent circuit of MOSFET

Table 1 Features of Power MOSFET

Item	Symbol	Conditions	n channel	p channel
Drain to Source Breakdown voltage	V (BR) DSX	I_{DS} = 10 mA	160 V	−160 V
Gate to Source Breakdown voltage	V (BR) GSS	I_{GS} = ±100 μA	±14 V	±14 V
Maximum Drain Current	I_D		7 A	−7 A
Maximum Power Dissipation	P_D		100 W	100 W
Threshold voltage	V_{GS} (off)	I_{DS} = 100 mA	0.8 V	−0.8 V
Transconductance	Gm	V_{GS} = 3 V	1 S	1 S
On-resistance	Ron	I_D = 5 A	1 Ω	1 Ω
Input capacitance	Cgs	V_{GS} = 5 V	600 pF	900 pF
Chip size			4.5 x 4.5 mm²	5 x 5 mm²

3. FREQUENCY RESPONSE

The Power MOSFET, which is a majority carrier device, has a better frequency response than a bipolar transistor. Figure 3 shows transconductance of the n channel MOSFET as a function of frequency for various values of external gate resistance Rg that are measured using the circuit in the inset. For example, in the case of Rg = 0, cut-off frequency fc is 3 MHz. The cut-off frequency fc is limited by the 600 pF gate to source capacitance and by the 65 Ω intrinsic gate resistance.

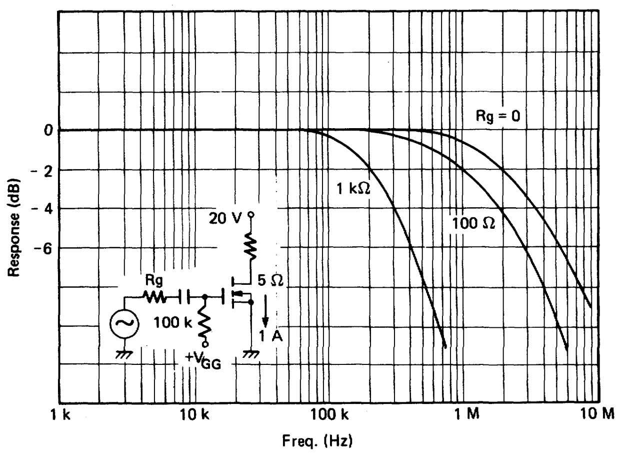


Fig. 3 Frequency response of transconductance

In the output stages of audio amplifiers, the B class emitter follower push-pull operation is generally used. Figure 4 shows the frequency response of the emitter and source follower. The frequency response of the source follower is 10 times greater than that of the emitter follower of an equivalent semi-conductor size.

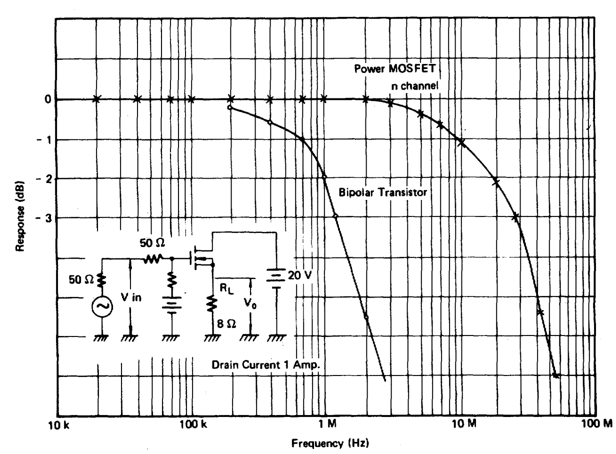


Fig. 4 Frequency response of source follower

Figure 5 shows the source follower and its equivalent circuit. Figure 5 (b) can also be applied to the emitter follower circuit. For a MOSFET the gate leak resistance R_2 can be neglected, but in bipolar transistors, R_2 cannot be neglected.

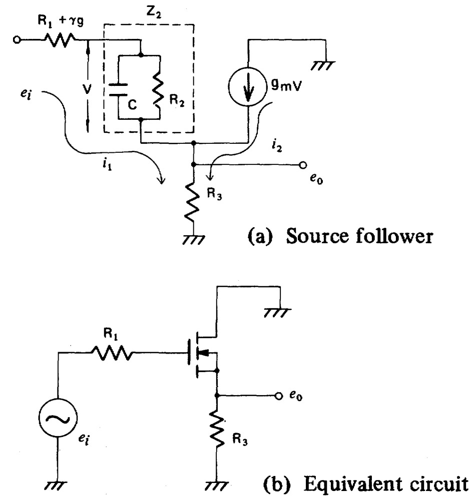


Fig. 5 Source follower and its equivalent circuit

The transfer function $\frac{e_o}{e_i}$ is derived:

$$\frac{e_o}{e_i} = \frac{1}{1 + \frac{R_1 + Z_2}{R_3 (1 + Z_2 \, gm)}} \quad (1)$$

For a Power MOSFET the parameters are $R_1 = 50\ \Omega + 65\ \Omega$, $R_2 = \infty$, $C = 600\,pF$, $gm = 1S$, $R_3 = 8\ \Omega$. The calculated cut-off frequency (- 3 dB), which is calculated by equation (1), ia approximately 20 MHz so it is roughly equal to the measured result in Figure 4. For emitter followers, the parameters change: $R_1 = 50\ \Omega + 5\ \Omega$, $R_2 = 7.5\ \Omega$, $C = 0.18\ \mu F$, $gm = 8S$, $R_3 = 8\ \Omega$. The calculated cut-off frequency is 1 MHz, and the results are roughly equal to the measured result.

In an emitter follower amplifier with more than a few watts output, the darlington arrangement is generally used because it can be driven by a small signal amplifier stage.

The transfer characteristics of the darlington connected emitter follower are rather complicated, but they can be roughly estimated by the transfer characteristics of single emitter followers. The phase shift of a darlington-connected emitter follower stage is the sum of phase shifts of each emitter follower stage. For a two stage darlington arrangement, the total phase shift is $-\pi$ radians. For three stage, the phase shift is $-\frac{3}{2}\pi$ radians. Figure 6 shows frequency responses of gain and phase shift: the phase shift of the darlington connected emitter follower is larger than that of the source follower.

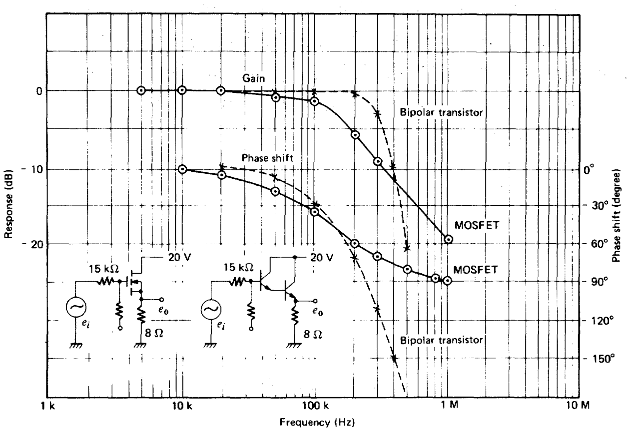


Fig. 6 Frequency response of gain and phase shift

In audio amplifiers, negative feedback is applied to improve non-linear distortions. Phase shifts of amplifiers above the unity loop gain should be kept below $-\pi$ radians. If the phase shift exceeds $-\pi$ radians, the frequency response should be decreased by frequency compensation. Then, in the audio amplifier with darlington connected emitter followers, frequency response should be limited.

In contrast, with MOSFET source followers, it is not necessary to use the darlington arrangement because of the high input impedance. Consequently, the maximum phase shift is $-\frac{\pi}{2}$ radians, and little frequency compensation is needed.

In conclusion, the frequency response of the amplifier with a Power MOSFET is definitely improved, the amplifier has large amounts of negative feedback at high frequency (40 dB at 300 kHz) and distortion is lowered.

In the source follower output stage, one complicated problem is parasitic oscillation due to a lead inductance to gate contact and negative resistance caused by capacitance load. Figure 7 shows the source follower's equivalent circuit, and in this figure, L is lead inductance to gate contact C_L is capacitance load which contains drain to source capacitance, stray capacitance of speaker code etc.

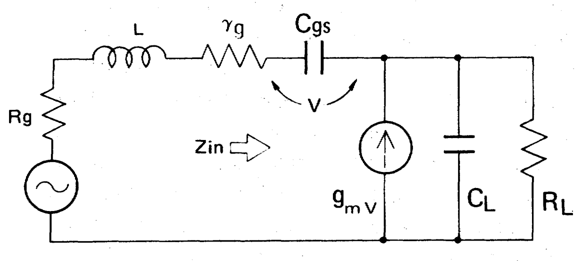


Fig. 7 Equivalent circuit of source follower

Input impedance Zin is given by:

$$Zin = \gamma g + sL + \frac{1}{sCgs} + \frac{R_L}{1 + sC_L R_L}\left(1 + \frac{gm}{sCgs}\right)$$

$$= \gamma g + sL + \frac{1}{sCgs} - \frac{sC_L R_L{}^2}{1 + \omega^2 C_L{}^2 R_L{}^2}$$

$$- \frac{sgm R_L}{(1 + \omega^2 C_L{}^2 R_L{}^2)\,\omega^2 Cgs} + \frac{R_L}{1 + \omega^2 C_L{}^2 R_L{}^2}$$

$$- \frac{C_L R_L{}^2 gm}{(H\omega^2 C_L{}^2 R_L{}^2)\,Cgs} \quad \ldots\ldots\ldots\ldots\ldots \quad (2)$$

under condition of:

$$\gamma g + \frac{R_L}{1 + \omega^2 C_L{}^2 R_L{}^2} -$$

$$\frac{C_L R_L{}^2 gm}{(1 + \omega^2 C_L{}^2 R_L{}^2)\,Cgs} < 0 \quad \ldots\ldots\ldots\ldots \quad (3)$$

where input impedance Zin has a real negative part. If the sum of Rg and the real part of Zin is negative, this circuit oscillates. Here, the imaginary part of the input impedance Zin should be zero as a condition of oscillation. Oscillation frequency fosc is given by:

$$sL + \frac{1}{sCgs} - \frac{sC_L R_L{}^2}{1 + \omega^2 C_L{}^2 R_L{}^2}$$

$$- \frac{sgm R_L}{(1 + \omega^2 C_L{}^2 R_L{}^2)\,\omega^2 Cgs} = 0 \ \ldots \ (4)$$

$$1 \gg \omega^2 C_L{}^2 R_L{}^2, \quad 1 \gg \omega^2 C_L Cgs R_L{}^2$$

$$fosc \fallingdotseq \frac{1}{2\pi}\sqrt{\frac{1 + gm R_L}{LCgs}} \quad \ldots\ldots\ldots\ldots\ldots \quad (5)$$

For example, at L = 50 nH, fosc is about 87 MHz calculated by equation (5), and practically speaking, fosc is 60 MHz.

Preventive measures for this parasitic oscillation are:

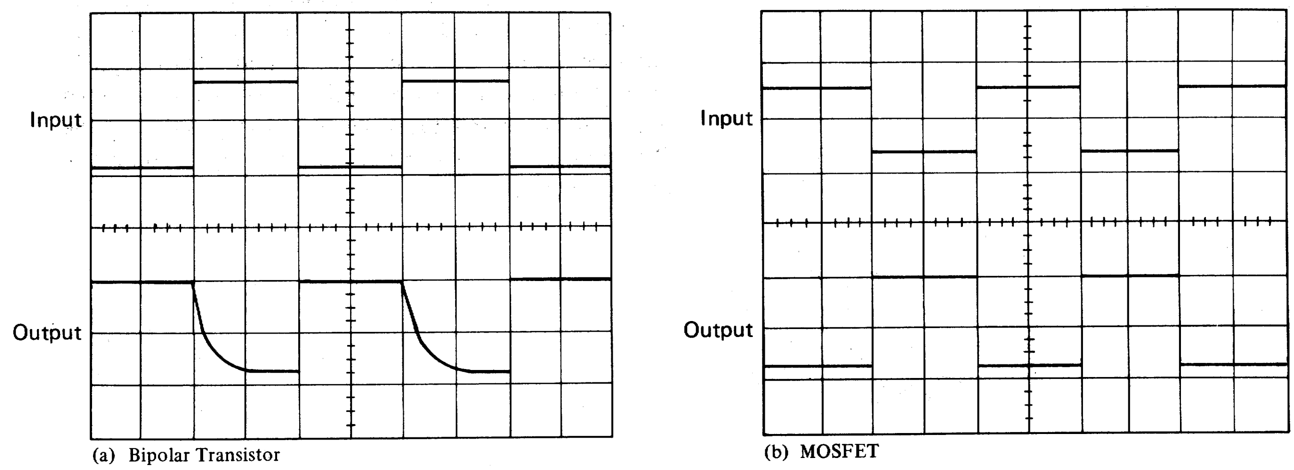
(1) an inductance between source and load

(2) a large Rg

These measures are introduced in equation (2).

4. SWITCHING DELAY AT HIGH FREQUENCY

In a bipolar transistor audio-amplifier, switching delay at high frequencies is another restriction of output power. Figure 8 (a) shows an example of bipolar transistor switching response. The long switching delay of 2 μsec, due to carrier storage in the base region, is recognized. Figure 8 (b) shows that in a Power MOSFET, the rise and fall time is about 50-100 μsec which is more than 20 times faster than that of a bipolar transistor.

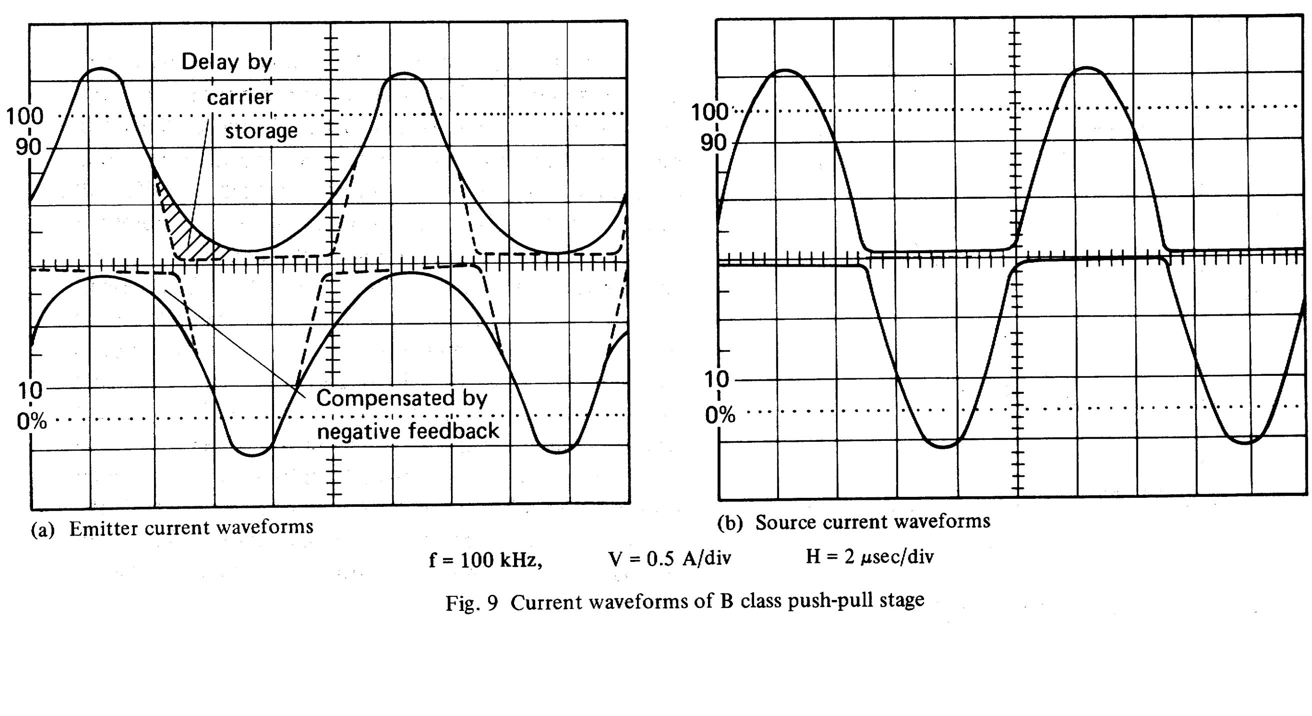


f = 50 kHz, H = 5 μs/div V = 5 V/div (Input), V = 10 V/div (Output)

Fig. 8 Switching response of bipolar transistor and MOSFET

Figure 9 (a) shows the actual current wave form for each half cycle of a bipolar, B class, push-pull stage. In this graph the dotted wave form is the presumed wave form when the transistor has no delay, and it is equal to the response of the Power MOSFET shown in Figure 9 (b). When the wave form is stretched by the carrier storage delay, the other half cycle's wave form is also stretched in such a way that the composite full wave form approaches to the correct sinusoidal wave form by negative feedback. As a result, in both transistors in the B class, push-pull stage useless current begins to simultaneously flow near the cross-over region, and power consumption is increased. This phenomenon becomes significant at high frequencies and brings the output stage into thermal runaway.

As illustrated in Figure 9 (b), the Power MOSFET does not have this delay, so high power operation at high frequencies is possible.



f = 100 kHz, V = 0.5 A/div H = 2 μsec/div

Fig. 9 Current waveforms of B class push-pull stage

Figure 10 shows total harmonic distortion of MOSFET amplifier compared with bipolar transistor amplifiers. Output power of the bipolar transistor amplifier is limited by thermal runaway at high frequencies.

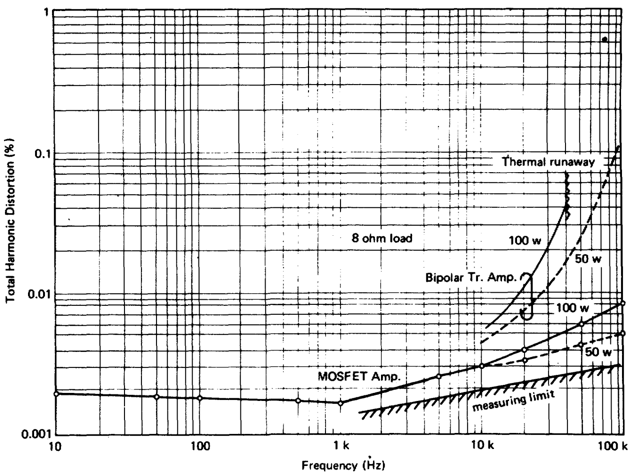


Fig. 10 Total harmonic distortion vs. frequency

5. CIRCUIT DESCRIPTION

Figure 11 shows the driving power of the Power MOSFET required for 100 watt output as a function of frequency compared with that of a bipolar transistor. The Power MOSFET requires little driving power, only a small amount of current to charge the gate to its source capacitance. This advantage makes the driving circuit simple.

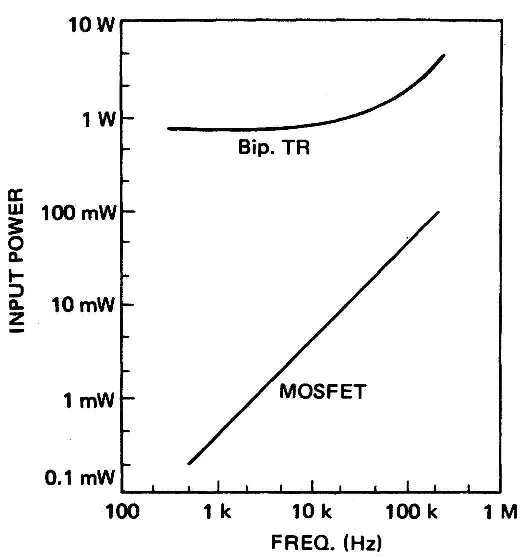


Fig. 11 Driving power required for 100 W output

The thermal stability of the Power MOSFET is better than that of bipolar transistors, and Figure 12 shows the transfer characteristics at various temperatures. At a large drain current, the temperature coefficient is negative. If the current density increases at one particular point on a chip, its temperature rises, then the current decreases due to the negative temperature coefficient. In this way the current flows equally throughout a chip. As a result, secondary breakdown does not occur and the Area of Safe Operation of the Power MOSFET is wider than that of a bipolar transistor of equal chip size.

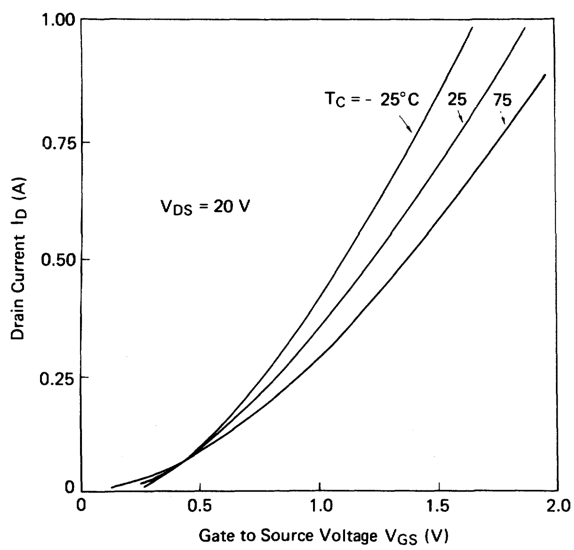


Fig. 12 Transfer characteristics at various temperatures

At a small drain current, the temperature coefficient is positive and the drain current is independant of temperature at a current level of 0.1 A. In the B class push-pull amplifiers, the n and p channel MOSFET act complementary as source followers for each signal's half cycle. The transfer characteristics of source followers change from linear to non-linear of square low as drain current decreases. It is best to bias the DC quiescent current for a B class push-pull stage at the cross-point between the linear and non-linear of square low region to minimize cross over. In our MOSFET this optimum bias current is conveniently equal to the drain current which is independent of temperature. Then, the temperature compensating network, indispensable in bipolar transistor amplifiers can be eliminated. As a result, the size of the total circuitry is reduced about 30 per cent compared with that of ordinary bipolar transistor circuits.

Figure 13 is a circuit diagram of an audio amplifier using a Power MOSFET. The first stage is a differential amplifier formed by Q_1-Q_2 and the second stage is a differential amplifier which has an active collector load (current mirror formed by Q_5 and D_1) for the push-pull action. This second stage differential amplifier directly drives the power stage formed by a complementary Power MOSFET pair. In ordinary bipolar transistor amplifiers, the power stage is constructed with a complementary pair of darlington connected transistors.

The amplifier uses two Power MOSFET pairs in a push-pull arrangement. One hundred watt amplifiers with Power MOSFET pairs can be designed if using a large heat radiator is feasible. But generally with a high power amplifier, a large heat radiator is more expensive than the device itself.

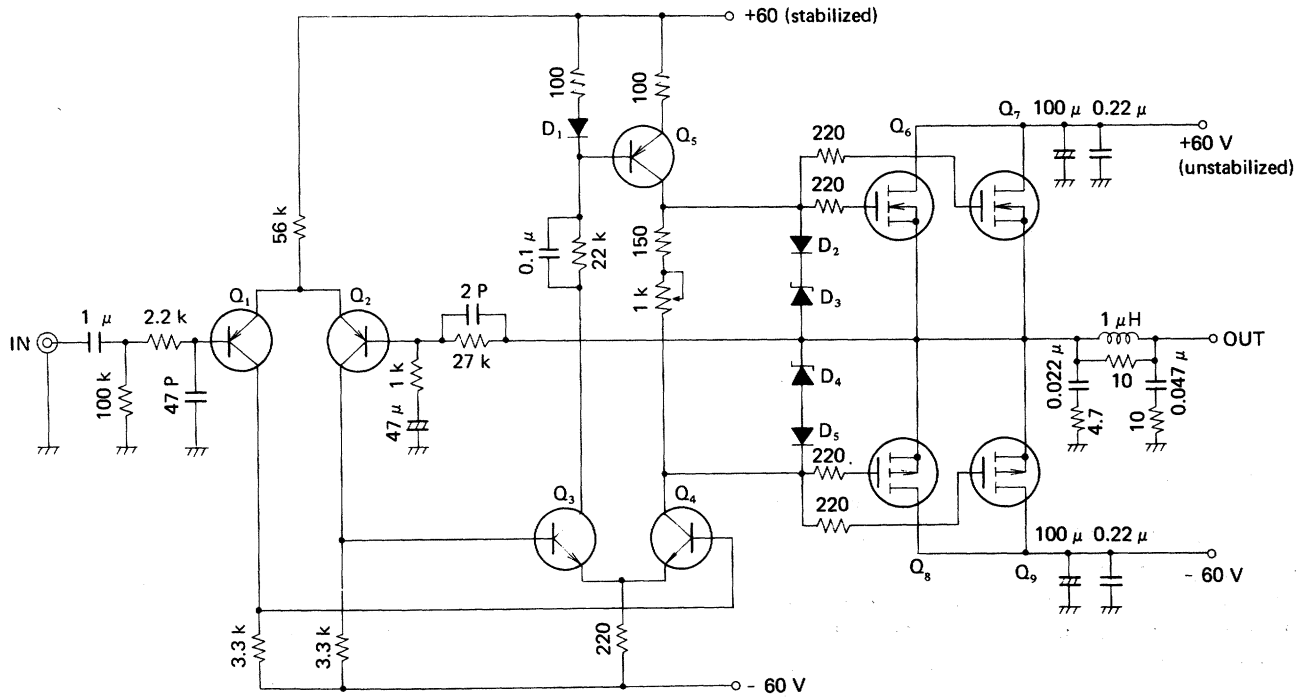


Fig. 13 Circuit diagram of the 100 watt MOSFET amplifier

Distortion is usually reduced by negative feedback, but it has been found that some distortion is never reduced by negative feedback. This residual distortion is induced by magnetic coupling between power supply wiring and an output line or feedback loop.

In the B class push-pull operation, half-cycles of output current flow through each output transistor, and this current contains even harmonic components. Most of the residual distortion, therefore, is this even harmonic component. This residual distortion can be cancelled with a negative coupled coil inserted in the output line or feedback loop. Especially at high frequencies, this counter-measure is very effective, and the total harmonic distortion is improved from 0.01 per cent to 0.003 per cent as shown in Figure 14.

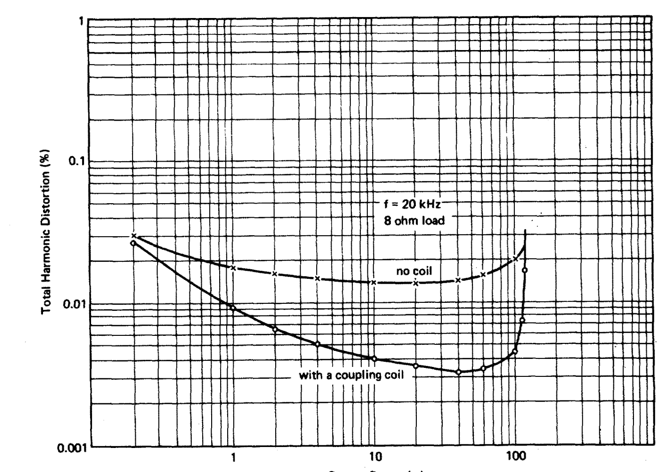


Fig. 14 Reducing harmonic distortion with a coupling coil

As a result, the MOSFET amplifier delivers a continuous power output of 100 watts at 8 ohms from 5 Hz to 100 kHz with no more than 0.01 per cent total harmonic distortion which is about ten times better than ordinary bipolar transistor amplifiers as illustrated in Figures 10 and 15.

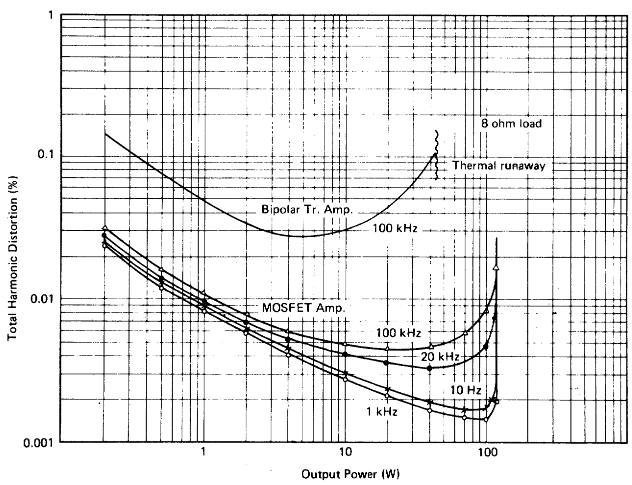


Fig. 15 Total harmonic distortion vs. output power

6. CONCLUSION

A 100 watt audio amplifier was developed using the new Power MOSFET developed by our MOS Device Group.

Since the phase shift of the source follower is less than that of the darlington connected emitter follower, the MOSFET amplifer has good frequency response. Consequently, low distortion is achieved with large amounts of negative feedback at high frequency.

The fast switching speed of the Power MOSFET can produce high power at a high frequency.

After studying the stability of source follower circuits, it was discovered that a coil inserted in output, and resistors inserted in each gate contact are effective.

The high input impedance of the Power MOSFET enables reduction of the size of driver circuits, and thermal stability can eliminate the temperature compensating networks, consequently, the total size of the circuits can be reduced by about 30 per cent.

Distortion induced by magnetic coupling between power supply wiring and output lines, can be cancelled by a negative coupling coil inserted in the output lines, and low distortion, wide power bandwidth, high stability and simple circuitry are achieved. The MOSFET amplifier delivers a continuous power output of 100 watts at 8 ohms from 5 Hz to 100 kHz with no more than 0.01 per cent total harmonic distortion which is about ten times better than ordinary bipolar transistor amplifiers.

ACKNOWLEDGEMENTS

Special appreciation for their helpful comments and support is expressed to Dr. M. Nagata, Mr. T. Okabe, Mr. I. Yoshida of the Central Research Laboratory, Mr. M. Ohta, Mr. S. Iguchi of the Semi-conductor Division; Mr. K. Maruyama of the Toyokawa Works; Dr. T. Sugimoto, Dr. G. Miyazaki and Mr. Y. Ohta of the Consumer Products Research Center, Hitachi, Ltd.

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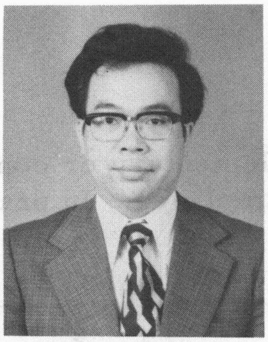
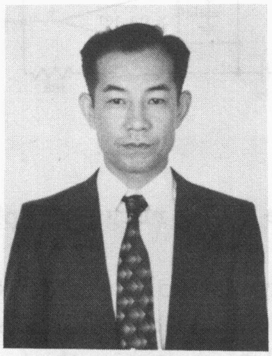
BIOGRAPHIES



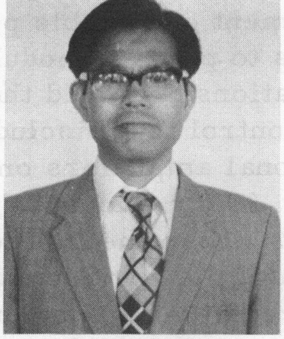
Tohru Sampei

Tohru Sampei received the B.S. degree in 1968, the M.S. degree in 1970, both from Waseda University, Japan. Since 1970, he has been a researcher in the Consumer Products Research Center, Hitachi, Ltd., where he has been engaged in the development of advanced stereo receivers. He is a member of the Institute of Electronics and Communication Engineers of Japan and the Acoustical Society of Japan.



Shin-ichi Ohashi

Shin-ichi Ohashi received the B.S. degree from University of Electro-communication in 1956. In 1960, he joined the Central Research Laboratory, Hitachi, Ltd., where he engaged in the research of DC amplifier for industrial use. From 1969 to 1973 he was a senior researcher in Consumer Products Research Center, Hitachi, Ltd., where he engaged in the development of advanced stereo receivers. Since 1974, he has been engaged in the development of circuits for white goods in addition to stereo receivers, as the manager of the Fourth Department. He is a member of the Institute of Electronics and Communication Engineers of Japan and the Institute of Electronics and Communication Engineers of Japan and the Institute of Electrical Engineers of Japan.



Shikayuki Ochi

Shikayuki Ochi received the B.S. degree from Osaka University, Japan, in 1956. In 1956 he joined the Central Research Laboratory, Hitachi, Ltd., where he has been working in the field of semi-conductor devices and integrated circuits for linear applications. He is a member of the institute of Electronics and Communications Engineers of Japan and the Institute of Electrical Engineers of Japan.