MICROPHONE PRE-AMPLIFIER FOR USE IN DATA ACQUISITION SYSTEM

by

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SUMMARY

A miniature high gain pre-amplifier has been designed to amplify signals from a Shure Model 488B microphone to a level suitable for direct connection of the microphone plus pre-amplifier unit to the inputs of either the Ampex Model AR200 Tape Recorder or the Pemco Model 110 Tape Recorder/Reproducer.
## CONTENTS

1. INTRODUCTION ............................................. 3
2. DESIGN REQUIREMENTS ................................. 3
3. AMPLIFIER DESIGN CONSIDERATIONS ............... 4
4. AMPLIFIER REALIZATION .............................. 5
5. CIRCUIT EXTENSIONS .................................. 8
REFERENCES .................................................. 9
FIGURES .......................................................
1. INTRODUCTION

The standard Pemco Model 18 microphone is internally fitted with a small battery powered pre-amplifier, which provides approximately 200 millivolt peak or an equivalent of a 140 millivolt R.M.S. output level for normal speech. This microphone is frequently used by Flight Research Group for the recording of speech in conjunction with data recording on either the Ampex Model AR200 Tape Recorder or the Pemco Model 110 Tape Recorder/Reproducer. Both these machines are 7-channel types capable of recording in either Direct Record or Frequency Modulation (F.M.) modes. At present there are plug-ins to enable up to 4 channels of Direct and up to 7 channels of F.M. to be recorded on each of these recorders. Normally speech is recorded on a separate Direct Record channel but when 7 channels of F.M. recording are required it will be necessary to interrupt the data recording on one channel to record the speech. Hence it is imperative that the output level of the speech pre-amplifier be sufficient to accommodate both Direct and F.M. recording on either recorder. The maximum sensitivities (expressed as the minimum input level to provide 100% recording level) are tabulated below:

<table>
<thead>
<tr>
<th>Tape Machine</th>
<th>Maximum Sensitivities</th>
</tr>
</thead>
<tbody>
<tr>
<td>Ampex Model AR200</td>
<td>Direct Record: 0.6 volt R.M.S.</td>
</tr>
<tr>
<td></td>
<td>F.M. Record: 1.0 volt R.M.S.</td>
</tr>
<tr>
<td>Pemco Model 110</td>
<td>Direct Record: 0.1 volt R.M.S.</td>
</tr>
<tr>
<td></td>
<td>F.M. Record: 0.5 volt R.M.S.</td>
</tr>
</tbody>
</table>

From the above table it is obvious that the output of the standard Pemco Model 18 microphone pre-amplifier is sufficient to provide 100% recording level only when used with the Pemco Model 110 Recorder in the Direct Record mode. In order to cater for all possibilities an output level of 1 volt R.M.S. is required. Normally all channels are set to accept a 1 volt R.M.S. input and the signal conditioning equipment is adjusted to provide the required output. If the input levels are standardized in this manner channels may be readily changed in the field if any malfunction is encountered.

A replacement pre-amplifier has been designed for the Pemco Model 18 microphone. This pre-amplifier provides adequate output for both Direct Record and F.M. Record modes for either of the recorders.

2. DESIGN REQUIREMENTS

The unmodified Pemco Model 18 microphone comprises a microphone, a pre-amplifier with self-contained battery pack and a switching facility. The microphone itself is a Shure Model 488B which employs the variable reluctance principle. At normal speech level the output from the basic microphone is approximately 2.5 millivolts peak. In order to raise this output to a 1 volt R.M.S. or a 1.4 volt peak level a gain of approximately 560 is required. This figure is to be compared with the gain of 80 provided by the original pre-amplifier. Stability of gain under conditions of varying environment and with changing of the semiconductor parameters is important.

The microphone itself has an impedance of approximately 150 ohm. It is essential that the input impedance of the amplifier be high compared with the microphone impedance. Normally the load presented by the tape recorders at the output of the pre-amplifier is 10 kilohm or higher. The pre-amplifier must therefore be capable of providing at least 1 volt R.M.S. into an external 10 kilohm load. So that the output level does not change with different loads it is essential that the amplifier output impedance be kept low.

As the frequency response of the microphone is fairly limited a pre-amplifier response of 100 hertz to 5000 hertz is quite adequate.
Size limitations provide one of the most significant constraints on the design of this preamplifier. The original pre-amplifier is powered from a self-contained battery pack consisting of five Mallory Type RM400R mercury cells having a rated capacity of 80 milliampere hours. The nominal voltage of these cells is 1.35 volts, hence a supply rail of approximately 7 volts results. It is highly desirable that the newly designed pre-amplifier use the same miniature battery pack. To achieve a long running time without battery replacement it is essential that current drain be kept low, preferably less than 1 milliampere. The use of a single low voltage supply rail at low current drain precludes the use of many of the integrated circuits currently available.

The original pre-amplifier is mounted on a printed circuit board having an area of approximately 1.3 square inches for the mounting of components. To enable the new pre-amplifier to be a direct replacement for the old one it is required that the size of the printed circuit card be identical in both cases. Space considerations alone dictate that the new pre-amplifier must not use a larger area printed circuit board than the original one.

### 3. AMPLIFIER DESIGN CONSIDERATIONS

In order to achieve the required performance detailed in the previous section two or more amplifying stages are required (compared with one stage used in the original pre-amplifier). Conventional common emitter stages (see Fig. 1) are far too bulky to be used in the present instance. Most of the bulk is introduced by components used for biasing the transistors correctly. Resistors $R_1$, $R_2$ and $R_3$ (Fig. 1) are for D.C. biasing only and likewise bypass capacitor $C_5$ which is usually quite large (at least tens of microfarads in audio amplifiers) is for D.C. biasing purposes only. Interstage coupling capacitors $C_1$ and $C_3$ are used to prevent the D.C. biasing of one stage being upset by the preceding or the following stage.

In a multistage amplifier it is possible to simplify the biasing arrangements of individual stages and use overall feedback to stabilize the D.C. and the A.C. performance of the amplifier. In low level applications it is possible to use direct coupling between collector and base of successive stages. Laakmann¹ and Bertoya² discuss amplifiers which employ direct interstage coupling. To provide adequate stability and for ease of applying negative feedback a three-stage direct coupled amplifier has been adopted.

A layout of a typical amplifier is drawn in Fig. 2. Transistors $Q_1$, $Q_2$ and $Q_3$ are direct coupled to form a high gain amplifier. The D.C. collector voltage of the final stage is stabilized via the input resistor $R_a$, and the combined feedback resistance. For high gain amplifiers $R_a$ is usually much higher than $R_2$ and $R_3$ and hence these latter two resistors together with $R_a$ define the D.C. performance. If the base current of $Q_1$ is made much lower than the current flowing through $R_2$ then the output D.C. voltage will be given by

$$V_o = \left(1 + \frac{R_2 + R_a}{R_2}\right)V_s$$  \hspace{1cm} (1)

where $V_s$ is the base to emitter voltage of $Q_1$. Changes of $V_s$ with temperature (typically 2 to 3 millivolt per degree Centigrade for constant collector current) will cause some drift in the output voltage with temperature.

$C_3$ is chosen such that the transfer impedance of the D.C. feedback network is high compared with $R_a$ over the frequency range of interest. Defining $Z_T$ as the transfer impedance (output voltage divided by current fed back to base of first stage) of the D.C. feedback network we may write

$$Z_T = j2\pi fC_3R_2R_a$$  \hspace{1cm} (2)

over the bandwidth of interest.

$|Z_T| > R_a$ at $f_1$ where $f_1$ is the required lower half power frequency of the amplifier.

Provided that the open loop gain is made sufficiently high, the midband A.C. signal gain will be given simply by

$$A_M = \frac{V_o}{V_i} = \frac{R_a}{R_1}$$  \hspace{1cm} (3)

The low frequency performance of this amplifier will be a function of the input and output coupling capacitors $C_1$ and $C_3$, and capacitor $C_4$ used to render the D.C. feedback path ineffective at high frequencies. At low frequencies the gain is given by
Assume that the low frequency attenuation is shared equally by each of the poles of $A_L$.

Put

$$A_L = \frac{j2\pi f C_L R_1}{1 + j2\pi f C_L R_1} \frac{j2\pi f C_L R_L}{1 + j2\pi f C_L R_L} \frac{Z_T R_8}{R_6}$$

$$= -\frac{R_8}{R_6} \frac{1}{1 - \frac{j}{2\pi f C_L R_1}} \frac{1}{1 - \frac{j}{2\pi f C_L R_L}} \frac{1}{1 - \frac{jR_8}{2\pi f C_L R_6 R_8}}$$

(4)

At the lower half power frequency $f_1$,

$$|A_L| = -\frac{A_M}{\sqrt{2}}$$

and $(1 + x^2)^{1/2} = 2^x$

$$x = \sqrt{2^{1/2} - 1} = 0.51$$

$$C_1 = \frac{1}{2\pi f_1 R_1 x} = 3.2 f_1 R_1$$

(5)

$$C_2 = \frac{1}{3.2 f_1 R_L}$$

(6)

$$C_3 = \frac{R_8}{3.2 f_1 R_6 R_8}$$

(7)

It is possible to eliminate the D.C. feedback path components $R_7$, $R_a$ and $C_3$ if suitable D.C. biasing stability can be achieved by way of $R_a$. Under these conditions

$$V_a = \left(1 + \frac{R_a}{R_2}\right) V_x$$

(8)

and the low frequency performance is determined by $C_1$ and $C_2$. In this case we may write

$$A_L = \frac{A_M}{(1 - jx)^2}$$

$$|A_L| = -\frac{A_M}{1 + x^2}$$

$$1 + x^2 = 2^x$$

$$x = \sqrt{2^{1/2} - 1} = 0.64$$

$$C_1 = \frac{1}{4.0 f_1 R_1}$$

(9)

$$C_2 = \frac{1}{4.0 f_1 R_L}$$

(10)

The use of the separate D.C. feedback path conveniently isolates the D.C. and the A.C. requirements and hence places less constraints on the design. However, as size minimization was of utmost importance in the microphone pre-amplifier design the elimination of these components was highly desirable.

4. AMPLIFIER REALIZATION

For D.C. stabilization (assuming $R_7$, $R_a$ and $C_3$ of Fig. 2 are not used) it is desirable to minimize $R_a$. To meet the condition that the input impedance of the amplifier (nominally $R_1$) be much higher than the microphone impedance (150 ohm) $R_1$ is required to be at least $1K$. If $R_1$ is made equal to $1K$ $R_a$ is required to be $560K$ to meet the A.C. signal gain requirement.
To obtain good operating point stability it is sufficient that the base current of the first stage be much less than the current flowing through $R_3$ and that the base to emitter voltage of the first stage remain fairly stable.

If the base current of the first stage is negligible in comparison with the current flowing through $R_3$ we may write using equation (8):

$$ R_2 = R_b \frac{V_s}{V_s - V_e} $$

where $V_s$ is the base to emitter voltage of the first stage and $V_e$ is the quiescent output voltage.

Substituting $R_b = 560K$, $V_s = 0.50$ volt (typical) and $V_e = 2.85V$ (optimum bias value to be calculated later) gives $R_2 = 119K$. A value of $120K$ was chosen.

The very low base current for the first stage may be realized by operating a very high gain (low noise) transistor at low collector current. A Fairchild Type SE4010 low noise transistor was chosen for the input stage and as these transistors were small and relatively cheap they were also chosen for the other two stages. A collector current of 10 microampere was chosen for the first stage. At this value of collector current the Type SE4010 typically requires 160 nanoampere of base current at 25°C. The corresponding base to emitter voltage drop is 500 millivolt approximately. For this value of base to emitter voltage the current flowing through $R_2$ will be $4.2$ microampere approximately which is 26 times the base current of the first stage (at 25°C). Hence the condition that the base current of the first stage be much less than the current through $R_2$ is fulfilled.

For a given supply voltage the collector current of the first stage should be fairly stable. The small changes in base to emitter voltage of $Q_1$ with temperature will produce only slight variations in the current through $R_3$. Since the base current of $Q_3$ is made small in comparison with the current through $R_3$ it follows that the collector current of $Q_1$ will be reasonably stable with temperature.

Curves furnished by the manufacturer indicate that, for a SE4010 at constant temperature, a 5 times change in collector current will produce approximately a 10% change in base to emitter voltage. Hence the changes in base to emitter voltage due to collector current changes will be very small. Variations in the base to emitter voltage of $Q_1$ will be due mainly to temperature variations. For the SE4010 operated at 10 microampere collector current the drift in base to emitter voltage is typically $2.3$ millivolt/°C. increase in temperature. Hence to produce a 10% drop in base to emitter voltage a temperature rise of approximately 22°C. is required. It will be shown later that, for a 7V supply, full output (1V R.M.S.) may be obtained provided that the base to emitter voltage falls within the range

$$ 0.81V > V_e > 0.32V $$

It follows that the drift in the operating point with temperature is well within the limits allowable.

In order to keep the current drain down to a minimum, $R_6$ in the output stage was made equal to $6.8K$. Since the design requires a minimum of 1V R.M.S. into a 10K load it is important that the quiescent operating point of the final stage be chosen to give maximum output into a 10K load. The operating point should be chosen midway between cutoff and saturation limits of the output transistor. Referring to Fig. 3 let $V_{ce}$ be the quiescent output voltage and define $V_{2e}$ and $V_{2s}$ as the cutoff and saturation limits respectively. The measured value of $V_{2s}$ was approximately 0.4 volt. The condition for cutoff is given by

$$ \frac{V_{ce} - V_{2e}}{R_6} = \frac{V_{2e} - V_2}{R_L} $$

$$ V_{2e} = \frac{R_L}{R_6 + R_L} V_{ce} + \frac{R_6}{R_6 + R_L} V_s $$

But

$$ V_2 = \frac{V_{2e} + V_{2s}}{2} $$

For optimum biasing

$$ \frac{R_1 V_{ce}}{2(R_6 + R_L)} + \frac{R_s V_2}{2(R_6 + R_L)} + 0.2 $$
Substituting $fts = 5'871, RL = 10K$ and $V_{ee} = 7V$ gives

$V_2 = 2.85 \text{ volt}$

(compare with measured value of 2.80 volt).

The peak to peak output available into a 10K load is equal to $V_{ge} - V_{2s}$.

From equation (11) $V_{ge} = 5.33 \text{ volt}$, $V_{ge} - V_{2s} = 4.93 \text{ volt}$.

Hence the output voltage attainable is

$$\frac{4.93}{2\sqrt{2}} = 1.75 \text{ volts R.M.S. into } 10K$$

(compare with measured value of 1.73 volt).

It is of interest to know by how much the output D.C. voltage may drift before an undistorted 1 volt R.M.S. is no longer obtainable. There will be an upper and a lower limit for $V_2$.

The lower limit is given by

$$V_2 - V_{2s} = 1.41 \text{ volt}$$

$$V_2 \simeq 1.81 \text{ volt}$$

The upper limit is given by

$$V_{ge} - V_2 = 1.41 \text{ volt}$$

From equation (11)

$$V_{ge} - V_2 = \frac{R_L}{R_S + R_L} (V_{ge} - V_2) = 1.41$$

$$V_2 = V_{ee} - \frac{1.41(R_S + R_L)}{R_L}$$

$$= 4.62 \text{ volt}$$

The limits on the output D.C. voltage for 1 volt R.M.S. into a 10K load are therefore

$$V_2 = 2.85 + 1.77 \text{ volt}$$

$$- 1.04 \text{ volt}$$

The equivalent input drift (change in $V_2$) to produce these limits is given by

$$V_2 = 0.50 \pm 0.31 \text{ volt}$$

For the SE4010 operated at 10 microampere collector current the base emitter voltage drift is typically $-2.3 \text{ mv/}^\circ \text{C.}$ increase in temperature. Hence the amplifier should operate within specification over quite a high temperature range.

The circuit of the pre-amplifier is drawn in Fig. 4. In this circuit capacitor $C_4$ reduces loop gain at high frequencies and is required to prevent high frequency oscillation. However, because of this capacitor an undistorted 1 volt R.M.S. is not obtainable above about 40 kHz because of internal overload of the amplifier. If the signal level is reduced the distortion disappears. The bandwidth is more than adequate for the present application. A summary of the measured performance characteristics of the amplifier is given below:

<table>
<thead>
<tr>
<th>Characteristics</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Gain</td>
<td>560</td>
</tr>
<tr>
<td>Bandwidth (–3 db)</td>
<td>85 Hz–150 kHz</td>
</tr>
<tr>
<td>Input Impedance</td>
<td>1000 ohm</td>
</tr>
<tr>
<td>Output Impedance</td>
<td>12 ohm</td>
</tr>
<tr>
<td>Open Loop Gain ($\frac{V_2}{V_2}$)</td>
<td>70,000</td>
</tr>
<tr>
<td>Noise Measured at Output</td>
<td>2.7 mV R.M.S.</td>
</tr>
<tr>
<td>Maximum Undistorted Output (into 10K load)</td>
<td>$1.75V$ R.M.S.</td>
</tr>
<tr>
<td>Power Requirement</td>
<td>$7V$ at 0.65 mA (4.55 mW)</td>
</tr>
</tbody>
</table>
The pre-amplifier has been fitted on a printed circuit board and covers an area of approximately 1.3 square inches. The battery pack consists of five Mallory Type RM400R mercury cells having a rated capacity of 80 mA hours which should theoretically give 123 hours service at a current drain of 0.65 mA. These cells have a diameter of 0.455 inch and a height of 0.125 inch. From equation (14) it is possible to calculate the voltage to which the supply may drop before an output of 1 volt R.M.S. is no longer attainable. Assuming $V_2$ is closely independent of $V_{cc}$ we may write (using equation (14))

$$V_{cc} = V_2 + \frac{1.41(R_0 + R_L)}{R_L}$$

Substituting $V_2 = 2.85V$, it follows that $V_{cc}$ may drop to 5.23V before full output is no longer obtainable.

A photograph of the complete amplifier and the microphone is given in Fig. 5.

5. CIRCUIT EXTENSIONS

It is possible to utilize the same form of amplifier in quite a variety of applications. Input impedance can be readily changed by changing $R_1$ and gain can be readily changed by changing the ratio $R_6/R_1$. A pre-amplifier with an extended low frequency response has been drawn in Fig. 6 and its performance has been checked experimentally with the following results:

<table>
<thead>
<tr>
<th>Gain</th>
<th>1000</th>
</tr>
</thead>
<tbody>
<tr>
<td>Bandwidth</td>
<td>0.6 Hz—100 kHz</td>
</tr>
<tr>
<td>Input Impedance</td>
<td>1K</td>
</tr>
<tr>
<td>Output Impedance</td>
<td>20 ohm</td>
</tr>
<tr>
<td>Maximum Undistorted Output</td>
<td>1.7V R.M.S.</td>
</tr>
<tr>
<td>Power Requirement</td>
<td>7V at 1 mA (7 mW)</td>
</tr>
</tbody>
</table>

POSTAL ADDRESS:

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Author | Title
---|---
TYPICAL DIRECT COUPLED AMPLIFIER
OUTPUT STAGE
RELUCTANCE MICROPHONE 150 OHM SHURE MODEL 488 B

COMMON

V_{CC} + 7V (5 MALLORY TYPE 400 R)
MERCURY CELLS

R_3 560 K
R_4 47 K
R_5 6.8 K

C_1 2.2 M
R_1 1K
R_2 120 K

Q_1 SE4010
Q_2 SE 4010
Q_3 SE4010

C_2 1M

R_L 10 K

R_6 560 K

ALL RESISTANCE VALUES ARE IN OHMS
ALL CAPACITANCE VALUES ARE IN PICOFARADS

COMPLETE PRE-AMPLIFIER CIRCUIT
FIG. 5a  INTERNAL VIEW OF MICROPHONE SHOWING PRE-AMPLIFIER AT THE LOWER RIGHT

FIG. 5b  ASSEMBLED MICROPHONE
PRE-AMPLIFIER WITH EXTENDED LOW FREQUENCY RESPONSE

ALL RESISTANCE VALUES ARE IN OHMS
ALL CAPACITANCE VALUES ARE IN PICOFARADS