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LOUD SPEAKERS AS HIGH-QUALITY MICROPHONES

By
Peter J Baxandall
Malvern
Worcs

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LOUDSPEAKERS AS HIGH-QUALITY MICROPHONES

Peter J. Baxandall

Electro-Acoustical Consultant
Malvern, Worcs., England

Abstract: It is established, using the reciprocity principle, that if a voltage-driven loudspeaker has a flat axial frequency-response, then such a response will also be obtained when it is used as a microphone, provided the associated amplifier has zero input impedance and a V_{out}/I_{in} response rising with frequency at 20dB/decade throughout the audio spectrum.

The design of a suitable amplifier raises problems with regard to signal-to-noise ratio and the ability to handle the very large low-frequency input levels that are liable to be produced. Two different approaches are described, both leading to practical designs.

In conclusion, consideration is given to the use of the technique for subjective loudspeaker quality assessment, and for the measurement of frequency response and absolute sensitivity.

Introduction

It is well known that an ordinary loudspeaker will work as a microphone if connected to a flat-response voltage amplifier, but the quality of reproduction is found to be extremely poor.

The work described here started several years ago, merely as the result of idle curiosity - what sort of equalization would be required for good-quality results, how good would the quality turn out to be, and what would the signal-to-noise ratio be like?

Rather than determine the necessary equalization by measurement, thought was given to the basic theory involved, and it was soon appreciated that a direct application of the reciprocity principle[1, 2, 3] would lead to a very simple practical solution; so simple, in fact, that it could be tried very easily and quickly. The first-class sound quality immediately obtained aroused considerable further interest and led to thoughts about possible applications, not only to making interview-type recordings without causing "microphone shyness"*, but also in the fields of subjective loudspeaker quality assessment and performance measurements.

Basic Theory

Fig. 1(a) depicts a loudspeaker fed with a flat-frequency-response voltage input, represented by the horizontal frequency-response graph shown at the left. Since the loudspeaker is assumed ideal,

* But if this technique is used, I think the interviewee should be given the option of having the recording erased without being played, if he so wishes!

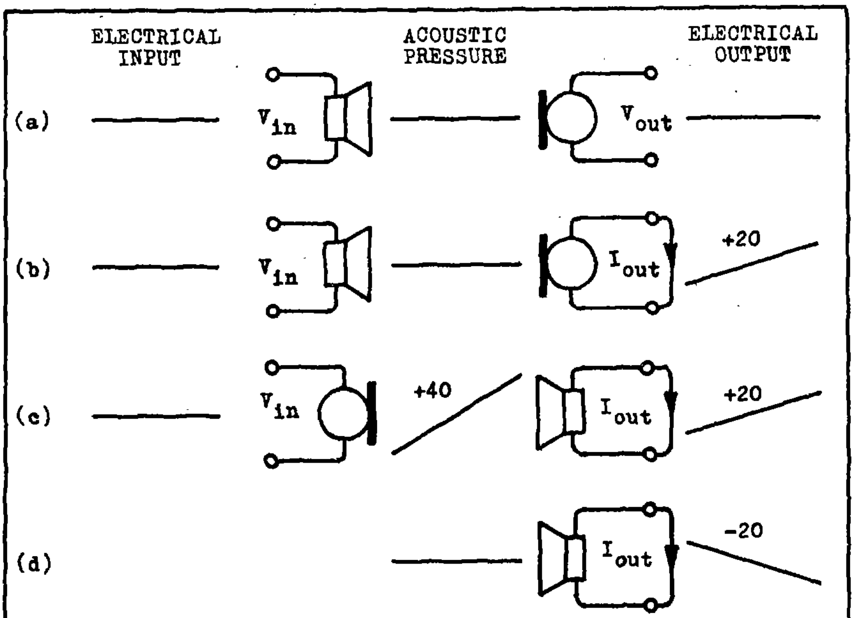

Fig. 1. Derivation of short-circuit-current response of an ideal loudspeaker used as a microphone. Slopes are in dB/decade.

the acoustic frequency response on axis is also flat, as represented by the middle graph. If the microphone has an ideal capacitance capsule of very small size, carrying a constant charge, then the open-circuit signal-output voltage from the capsule is constant with frequency, as shown by the right-hand graph. The internal electrical impedance of the capsule, if it is of the omnidirectional or pressure type, is in practice very nearly a pure capacitance, i.e. the motional-impedance component is negligible. With this assumption, which is made as a mere convenience to simplify the explanation, the a.c.-short-circuit output current of the capsule exhibits a response rising with frequency at 20dB/decade as shown in Fig. 1(b), for it is effectively due to a constant e.m.f. in series with a capacitance.

Thus a constant voltage applied to the left-hand terminal pair in (b) gives a short-circuit current from the right-hand terminal pair rising with frequency at 20dB/decade; or, in other words, the mutual admittance, = I_{out}/V_{in}, is proportional to frequency. The reciprocity principle[1, 2, 3], in the form here of interest, simply states that in any linear passive system of a reversible nature, having electrical input and output terminals, the mutual admittance (or impedance) is the same measured in either direction. Thus if V_{in} is applied to the microphone and the short-circuit output current from the loudspeaker is determined, it will be found to be the same in absolute magnitude and in frequency response (including phase) as for the conditions of Fig. 1(b). This reversed operation is shown in Fig. 1(c), where the microphone and loudspeaker positions have been interchanged to preserve the left-to-right signal-flow direction. The middle graph may be ignored for the moment.

What is now required is to establish the nature of the acoustic input to the loudspeaker in (c), and this is easily done because of the very simple behaviour of a diminutive pressure capsule, voltage driven. Being stiffness-controlled, the diaphragm amplitude is frequency-independent. But to radiate a constant output level at all frequencies, a diaphragm very small compared with the wavelength, mounted in an enclosed case, must vibrate with an acceleration independent of frequency[1, 2, 3, 4]. Since acceleration is the second differential coefficient of amplitude, it follows that a constant diaphragm amplitude will produce radiation in which the pressure rises with frequency at 40dB/decade, as shown by the middle graph in Fig. 1(c). It is thus established that if an ideal loudspeaker used as a microphone is placed in the path of an acoustic wave arriving axially, and if the acoustic wave pressure in free space rises with frequency at 40dB/decade, then the short-circuit output current from the loudspeaker rises with frequency at 20dB/decade. Therefore if the acoustic wave had constant pressure independently of frequency, the loudspeaker output current would fall with rising frequency at 20dB/decade.

Thus it may be concluded that to obtain a flat axial frequency response from an ideal loudspeaker used as a microphone, it is necessary to feed the loudspeaker into a special circuit having zero input impedance and a V_{out}/I_{in} response rising with frequency at 20dB/decade.

More generally, if a loudspeaker gives a certain frequency response with the ideal measuring microphone in a specified position, then when used as a microphone with the above-mentioned special circuit, it will give this same frequency response when excited by a point source with a frequency-independent output located at the position previously occupied by the measuring microphone.

Circuit Techniques

The first experiment used the purely passive arrangement shown in Fig. 2. The ferrite-cored transformer has a low primary inductance, such that the primary shunt reactance is small compared with 8 ohms even at 15kHz. The loudspeaker then sees

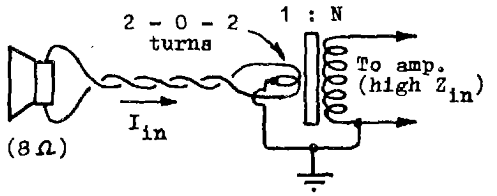

Fig. 2. Simple passive arrangement employing a differentiating transformer.

a reasonable approximation to a short-circuit load at all audio frequencies, and the secondary voltage is given by:-

$$|V_{sec}| = I_{in}\ \omega L_{pri}\ N \qquad . \quad . \quad . \quad . \qquad (1)$$

where N is the transformer step-up ratio.

Provided L_{pri} is constant throughout the audio spectrum, which is the case when a ferrite core is used (though it would be far from true if a normal Mumetal or Permalloy core were used), the required 20dB/decade rising frequency response is obtained.

In practice, care must be taken to avoid a resonance peak at high audio frequencies due to the inevitable secondary winding capacitance, cable capacitance and amplifier input capacitance. There is also the difficulty that many preamplifiers with negative feedback are liable to oscillate at RF when a tuned circuit, such as is constituted by the transformer secondary, is presented to the input, and radio-station interference troubles may also be experienced if the unmodified Fig. 2 arrangement is used. Suitable values of R and C shunted across the secondary will enable satisfactory results to be obtained with a given amplifier of high input impedance, but the difficulty in evolving a design on this basis for general use is that the associated amplifier input impedance is then an unknown quantity. A compromise must be struck between accuracy of frequency response, signal-to-noise ratio, and sensitivity to variations in loading.

Ideally, the following performance requirements should be satisfied:-

(a) $|Z_{in}|$ well under 1 ohm at all audio frequencies.

(b) Accurate 20dB/decade V_{out}/I_{in} response over the whole audio spectrum, independently of reasonable load variations.

(c) Ability to withstand high-level input currents, which may exceed 10mA rms at low frequencies.

(d) Noise performance such that the audible noise output when used with suitable associated equipment is mainly due to Johnson noise generated in the real part of the loudspeaker coil impedance.

(e) Very low non-linearity distortion.

(f) Low sensitivity to RF interference and mains hum.

Many possibilities were considered in the attempt to evolve an economical circuit design which would simultaneously satisfy all the above requirements, and space does not permit an account to be given here of all the ideas and experimental observations that now fill over 200 pages of laboratory notes. Instead, just two practical circuits will be described, one using a differentiating ferrite-cored input transformer, and the other employing a flat-response Mumetal-cored input transformer feeding a virtual earth, with differentiation carried out in a later part of the circuit.

Solution A

The great advantage of doing the differentiation right at the beginning by means of a transformer, is that the subsequent active circuits do not then have to handle the excessive signal levels generated by a loudspeaker-microphone when subjected to SPL's of 100dB or more at low frequencies. For this reason much thought was given, in the earlier stages of the work, to refining the Fig. 2 approach, and the circuit of Fig. 3 was evolved.

The reactance of the 6-turn primary winding rises to 1.36 ohms at 15kHz, providing a sufficiently low input impedance for most purposes with 8-ohm loudspeakers. For precise measurement applications, however, additional taps are available to give a 2-turn primary with a reactance at 15kHz of only 0.15 ohm. When this is used, there is, of course, some sacrifice of signal-to-

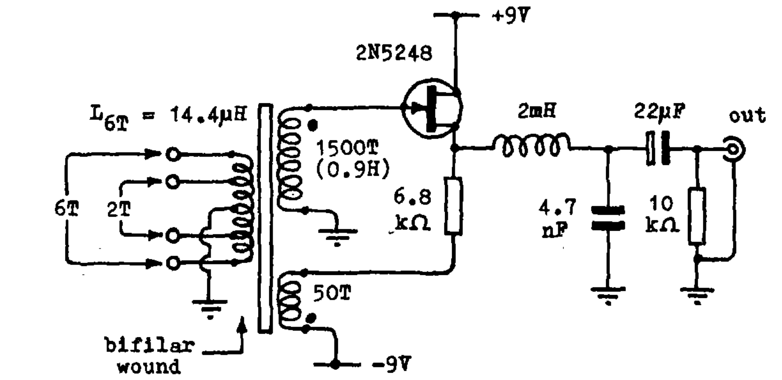

Fig. 3. Practical circuit 'A'.

noise ratio. The noise aspect of the design will now be considered in some detail.

Suppose initially that the secondary windings are disconnected from the remainder of the circuit, and that copper losses, core losses, and winding capacitance may be ignored. Then, looking into the 1500-turn winding, a pure inductive reactance is seen if nothing is connected to the primary, and a pure reactance does not generate a noise voltage at finite frequencies. If, however, an 8-ohm resistor is connected across the 6T primary, this will appear as a 500kΩ resistance across the 1500-turn winding, as shown in Fig. 4(a). This parallel CR combination may be

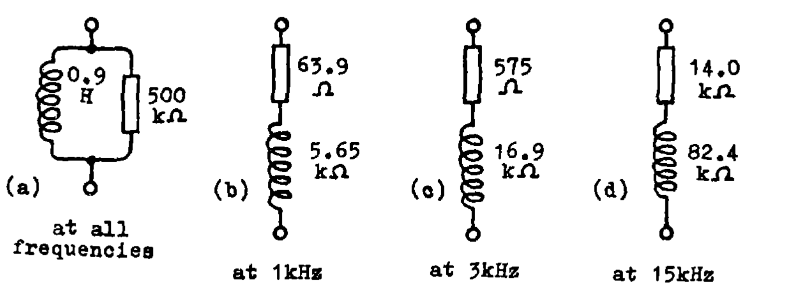

Fig. 4. Equivalent circuits.

expressed at any given frequency as an equivalent series combination[5], as shown in (b), (c) and (d) for three specific frequencies. Provided the FET current-noise generator may be neglected, which is normally justified, the overall noise performance, in the absense of transformer losses, is then determined by the ratio of the FET equivalent voltage-noise resistance, R_{N_V}, to the series resistance value at the frequency concerned, as shown in Fig. 4. It may be noticed that the series resistance values are almost exactly proportional to the square of the frequency, so that the corresponding noise voltages in unit bandwidth which they generate are closely proportional to frequency. This, of course, is what would be expected, for the 8-ohm resistor feeds an almost constant noise current in unit bandwidth to the primary, and this is subjected to a 20dB/decade rising frequency response.

Since R_{N_V} at highish audio frequencies is typically around 500 ohms for a reasonable FET running at about 1mA, it is clear from the Fig. 4 values that a noise figure of 3dB or better will be obtained above about 3kHz, but that below this frequency the noise output will be predominantly FET noise. In the practical design, which employs a 22mm pot core with an A_L value of 400nH for 1 turn, there is some further noise from the copper resistance of the 1500-turn winding, which is about 350 ohms, but the primary resistance, and the core-loss resistance, contribute negligible noise. The latter, expressed as a shunt resistance across the 1500-turn winding, is well over 10MΩ and reference to Fig. 4(a) shows that this will produce much less noise than that from the 500kΩ resistance representing the 8Ω across the primary. From this point of view, a more balanced design would employ a pot core with a higher A_L value, thus more nearly equalizing the noise contributions from the copper and core losses and giving less total noise, but this improvement would be accompanied by an increase in the non-linearity distortion when handling high SPL's at low frequencies, and was therefore not adopted.

A feature of the Fig. 3 circuit not yet considered is that involving the 50T winding. The purpose of this is to provide damping of the large response peak that would otherwise occur at about 40kHz due to resonance between the main secondary inductance and the inevitable shunt capacitance. This damping is achieved without introducing significant extra noise[6]. The amount of damping is equivalent to approximately 200kΩ across the 1500-turn winding, and results in a response, for constant primary current, that rises with frequency in a manner conforming within a fraction of 1dB to the desired 20dB/decade asymptote up to about 30kHz, and falls at 20dB/decade above about 50kHz. (To obtain this result it is necessary to employ a two-section bobbin with the 1500-turn winding in one section only; a single-section bobbin will give too high a winding capacitance.) The noise introduced by the damping arrangement is merely that caused by the passive presence of 6.8kΩ across the 50-turn winding, and it is equivalent to approximately 6MΩ across the main secondary.

For the noise from the FET current-noise generator to be negligible, it is necessary for the current-noise equivalent resistance[7, 8], R_{N_i}, to be high compared with 500kΩ, but with good FET's values of R_{N_i} of at least 100MΩ are normally met*.

Subjectively, the connection of an 8-ohm resistor across the 6-turn primary of the Fig. 3 circuit gives a considerable increase in the audible noise level, and it is evident that what is added is a rising spectrum of high-frequency hiss, leaving the lower-frequency components unaffected. In practical use, however, the latter part of the noise, which is due to the shortcomings of the circuit, tends to merge with room background noise, so that it is mainly the high-frequency hiss that determines the effective goodness of the system from a noise viewpoint. With a reasonably sensitive loudspeaker, the subjective noise performance is considerably better than that of many microphones.

The sensitivity provided by the Fig.3 circuit with a typical bookcase loudspeaker is in the region of $5\text{mVN}^{-1}\text{m}^2$ (0.5mV/μbar). It could be improved by putting more turns on the main secondary winding, and this would also improve the noise performance, though a redesign of the transformer, maybe with a larger size of pot core, to give a reduced value of secondary winding resistance would be necessary to obtain the maximum possible benefit.

Another way to increase the sensitivity of the complete circuit unit is, of course, to depart from using merely a simple FET follower, and a very satisfactory practical circuit was tried

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* FET current noise is usually given in $\text{pA}/\sqrt{\text{Hz}}$, but the use of the alternative R_{N_i} representation, originally proposed by Professor E. A. Faulkner in reference 7, gives a much more vivid sense of relative values, and is much to be recommended. The relationship is:-

$$R_{N_i} = 0.016/(\text{pA}/\sqrt{\text{Hz}})^2, \quad R_{N_i} \text{ being in M}\Omega$$

in which the FET output was taken from the drain and fed to a TDA1034 operational amplifier, with overall negative feedback from the output of the latter back to a 100-ohm resistor in the source circuit of the FET. This arrangement gave a voltage gain of 5. A later modification involved adding a small capacitor from the output back to the gate, thus producing an active negative capacitance across the transformer winding and raising the resonance frequency. With suitable readjustment of the resistive feedback damping, this technique extended the range of the accurate 20dB/decade rising frequency response upwards by about an octave. It should enable a higher value of secondary inductance to be adopted while still preserving the wanted response up to a given maximum frequency, but such a design was not actually tried, for attention had by then shifted largely to Solution B described below.

The RF filter shown in Fig.3 eliminated considerable radio interference trouble. The FET follower is very much less effective in demodulating such interference than is the single-ended junction transistor input stage of the subsequent amplifier. The filter resonance frequency is about 50kHz, the Q-value being in the region of unity with most of the damping provided by the resistive output impedance of the FET follower.

With no magnetic screening of the ferrite input transformer, mains hum is found to be quite negligible provided the input transformer is placed at least 1 metre from ordinary mains equipment, but at 0.5 metre it may be necessary to orientate it for minimum hum. A Mumetal can is therefore desirable for exacting applications. The induced hum voltage is not, of course, subjected to the 20dB/decade lifting response, so the hum sounds quite "normal".

The harmonic distortion introduced is mainly third-harmonic due to magnetic non-linearity in the input transformer. For a given input current, the distortion is not greatly frequency-dependent and is in the region of 0.1% at 10mA rms to the 6-turn primary. Though input currents will always be much lower than this at middle and high audio frequencies, the significant distortion is of the intermodulation type, these higher-frequency components being modulated by the inductance variations resulting from the large input currents which accompany high SPL's at bass frequencies.

Though the Fig. 3 type of circuit is economical and can give excellent results, the alternative solution described below is on the whole preferred. It does not involve the compromise between signal-to-noise ratio and lowness of input impedance that is inherent in the Fig. 3 approach, so that the extra input terminals there provided are no longer required. Lower distortion is also readily obtainable, though this is of doubtful real benefit.

Solution B

The essence of solution B is shown in Fig. 5(a). The transformer has a Mumetal core and is of generally similar design to that used in a normal microphone amplifier. For good noise performance the step-up ratio must be sufficient to raise the 8-ohm loudspeaker impedance to well over the R_{N_v} value for the operational amplifier - ideally[8] it should be raised to $\sqrt{R_{N_v}R_{N_i}}$. Since R_{N_v} for a good low-noise operational amplifier such as the TDA1034NB or NE5534AN is typically about 700Ω, with R_{N_i} in the 100kΩ region, it is necessary to step the loudspeaker impedance up to several kilohms.

The second requirement for good noise performance is that the value of R_1 should be several times the loudspeaker impedance referred to the secondary, otherwise the Johnson noise it generates will significantly worsen the noise figure[8]. R_2 is made much less than R_1 and contributes negligible noise.

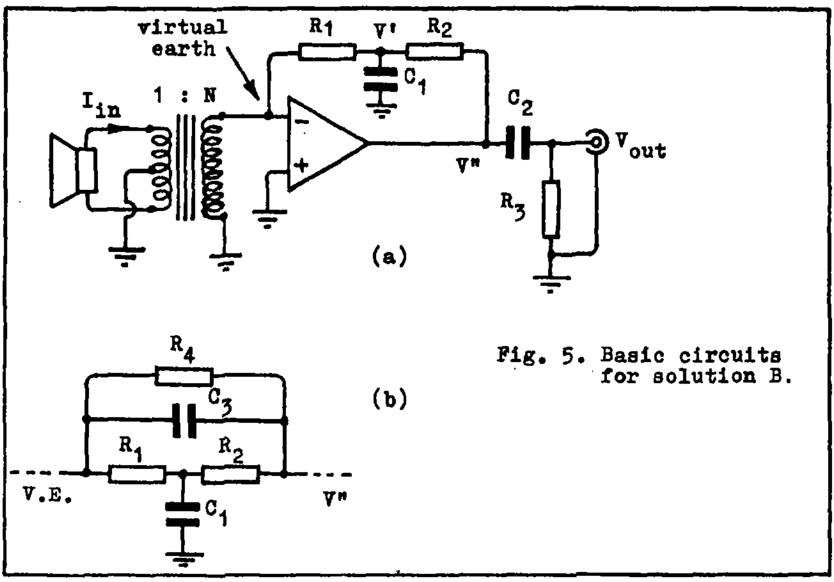

Fig. 5. Basic circuits for solution B.

With the step-up ratio and R_1 value already determined by the above considerations, the voltage V' at the point shown, for a given input current I_{in}, is also determined. The simplest way to introduce the required 20dB/decade rising frequency response would be to make C_1 very large and dispense with C_2. To maintain the 20dB/decade slope down to 20Hz, however, V" would have to be

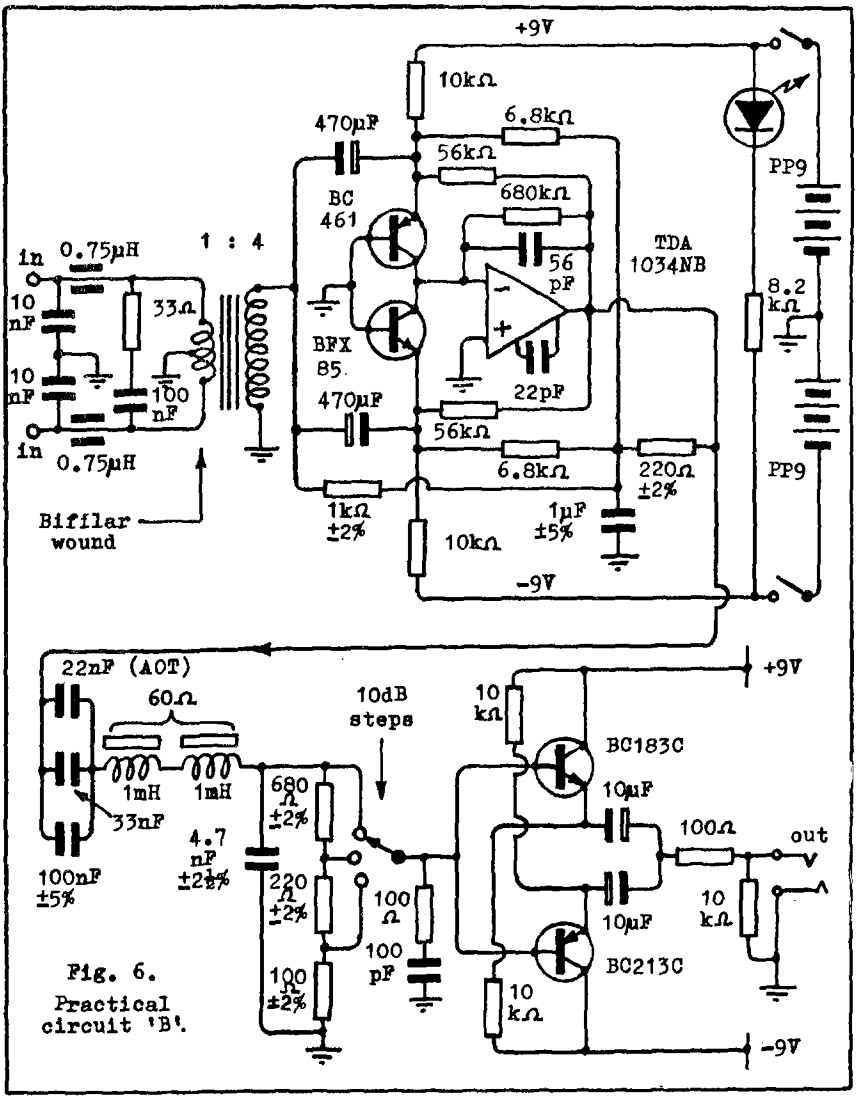

Fig. 6.
Practical circuit 'B'.

several times V' even at this low frequency, and overloading would then occur at an intolerably low SPL. The solution adopted is to make the corner frequency introduced by C_1 quite high, e.g. 1000Hz, so that it gives a lifting response only above this frequency. The required 20dB/decade slope at lower frequencies is provided by C_2 and R_3, the two time-constants being made accurately equal. With this arrangement the large value of V' liable to be reached at low audio frequencies is not accompanied, to any significant extent, by an even larger value of V", so that overload problems are much eased. By suitably choosing the value of the common time-constant, the sensitivity of the circuit may be tailored to suit requirements. The relevant equation is:-

$$V_{out} = \frac{I_{in}}{N}(R_1 + R_2)\omega T \qquad . \quad . \quad . \quad . \quad (2)$$

$$\text{where} \quad T = C_2R_3 = C_1\frac{R_1R_2}{R_1 + R_2} \qquad . \quad . \quad . \quad . \quad (3)$$

Considering the nature of equation (2), it is seen that the quantity $(R_1 + R_2)T/N$ must have the dimensions of inductance, and it is convenient to refer to it as the "equivalent inductance", L_{eq}, of the complete circuit. Thus:-

$$L_{eq} = \frac{(R_1 + R_2)T}{N} \qquad . \quad . \quad . \quad . \quad . \quad (4)$$

So for <u>any</u> circuit designed for operating a loudspeaker as a microphone on the basis described in this paper:-

$$V_{out} = I_{in} \times L_{eq} \qquad . \quad . \quad . \quad . \quad . \quad (5)$$

If the transformer ratio in Fig. 5 is 1 : 20, thus stepping 8 up to 3.2kΩ, which is about as low a value as is acceptable for low operational amplifier noise, then R_1 will have to be at least 10kΩ for its noise contribution to be adequately low. An input current of 10mA rms at low frequencies will then produce a voltage V' of 5V rms, V" being slightly higher. This is a higher level than the operational amplifier can produce on a convenient battery supply of ±9V, especially if the batteries are not new, and it is largely for this reason that the more elaborate practical design shown in Fig. 6 was finally evolved. Before describing the latter, however, a few further points relating to Fig. 5 remain to be considered.

First, feedback as shown, via a large lagging time-constant, is liable to produce oscillation in practice, and even if actual oscillation does not occur, the gain would continue rising with frequency up to frequencies far above the top of the audio band, which is most undesirable. The solution adopted is to replace the feedback network of Fig. 5(a) by that shown in (b). C_3 is chosen to give the desired asymptote for the 20dB/decade fall-off in response with rising frequency at very high frequencies, R_4 being chosen for an appropriate Q-value, somewhat less than unity, to give a satisfactory transition from 20dB/decade rising response to 20dB/decade falling response. The relationship between V" and the current I_{ve} supplied to the virtual earth is:-

$$\frac{V''}{I_{ve}} = -R_p \times \frac{1 + pT_1}{1 + \frac{1}{Q}pT + p^2T^2} \quad . \quad . \quad . \quad . (6)$$

in which $T_1 = C_1 \times \frac{R_1R_2}{R_1 + R_2}$ $\qquad R_p = \frac{R_4(R_1 + R_2)}{R_4 + (R_1 + R_2)}$

$$T = \sqrt{R_pC_3T_1} \qquad Q = \frac{T}{\frac{R_p}{R_4}T_1 + R_pC_3}$$

If the time-constant R_3C_2 in Fig. 5(a) is made accurately equal to T_1, then the overall response to a current supplied to the virtual earth becomes:-

$$\frac{V_{out}}{I_{ve}} = -R_p \times \frac{pT_1}{1 + \frac{1}{Q}pT + p^2T^2} \quad . \quad . \quad . \quad . (7)$$

This is the response of an ideal tuned circuit, or alternatively it may be regarded as the resultant of a 20dB/decade lifting response component, due to pT_1, in cascade with a second-order low-pass filter response.

Another feature of the Fig. 5(a) circuit is that there is very little zero-frequency feedback to stabilize the sit-points - there would be none in the absense of copper resistance in the transformer. A suitably large electrolytic capacitor in series with the secondary winding is therefore desirable.

If i.c. operational amplifiers could be obtained with much lower R_{N_V} values than 700Ω, there would be no need to step up to so high a secondary impedance. With a smaller step-up ratio, and all impedance values in the circuit lowered correspondingly, a larger value of current, I_{in}, to the primary could be handled before the occurrence of voltage overload at the operational amplifier output. However, because of the non-availability, as far as is known, of suitable operational amplifiers, a push-pull common-base preamplifier stage has been added in Fig. 6. This uses fairly large-area transistors, with low values of $r_{bb'}$, and their R_{N_v}'s appear in parallel, giving an effective total R_{N_v} value of about 15Ω only. The total R_{N_i} value is in the 3kΩ region[8], the impedance seen looking into the transformer secondary, with an 8Ω loudspeaker across the primary, being nominally 128Ω. The noise contributed by the operational amplifier itself is relatively negligible and the overall noise performance is therefore very satisfactory.

To help in achieving good HF stability, the feedback via the 56pF capacitor (corresponding to C_3 in Fig. 5(b), is taken not to the input stage emitters but to the i.c. input. The effect on the working frequency response is the same. The purpose of the 680kΩ resistor is merely to ensure that there is some feedback on the i.c. from the moment of switch-on and before the input stage has had time to build up current. Without this resistor, the switch-on output disturbance is much increased.

The resistor corresponding to R_2 in Fig. 5(b) is 220Ω, but R_1 is replaced by 1kΩ in parallel with a pair of 6.8kΩ resistors. The 56kΩ's provide the equivalent of R_4' in Fig. 5(b).

By taking some of the feedback via 1kΩ to the top of the secondary, instead of all to the emitters, the values required for the electrolytic capacitors are conveniently reduced. This may be understood as follows. On applying an imaginary a.c. voltage to the top of the secondary winding, a leading current flows to the electrolytic capacitors. This develops an output voltage from the operational amplifier and causes a current to be fed back via the 1kΩ resistor. This current has to be supplied by the imaginary voltage source just mentioned, and it is in phase with that supplied to the capacitors. The effective value of capacitance seen by the secondary winding is thus increased severalfold, maintaining the desired response slope of 20dB/decade accurately down to 20Hz without needing exorbitantly large electrolytic capacitor values.

The design value for the Q of the active tuned-circuit response mentioned above is 0.54, and in combination with a Q-value of nominally 1.31 for the passive filter involving the 1mH inductors, both having resonance frequencies of approximately 50kHz, an accurate fourth-order Butterworth response is obtained[9], i.e. the overall response of the circuit is that of a differentiator and a fourth-order Butterworth low-pass filter in cascade. The measured response of the Fig. 6 circuit is shown in Fig. 7.

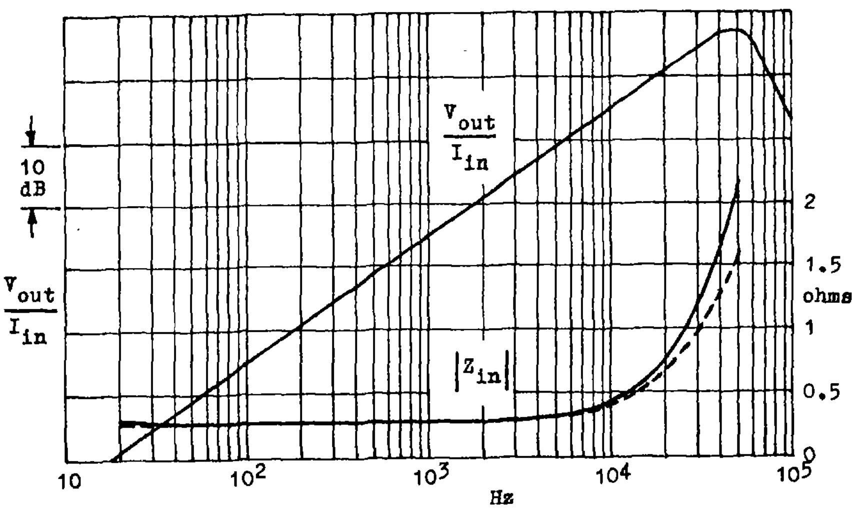

Fig. 7. Measured results for the Fig. 6 circuit. The broken-line curve was obtained with the transformer secondary short-circuited.

The choice of primary turns on the transformer is determined more by noise considerations than by frequency response, for the shunt resistance due to eddy currents in the core generates Johnson noise like any other resistance. The measured value for the eddy-current shunt resistance, referred to the secondary, is 1.1kΩ at 60Hz, but since it rises slowly with frequency due to the changing flux distribution within the thickness of each lamination, the value at high frequencies, which is more significant from a subjective noise viewpoint, is considerably higher, so that the eddy-current contribution to the audible noise is very small. It could be reduced further by using thinner laminations, but some reduction in the number of turns would then be justified, thus reducing the leakage inductance and hence the rise in impedance at high audio frequencies shown in Fig. 7.

The Fig. 6 circuit is found to be less susceptible than that of Fig. 3 to magnetic mains-hum pick-up in the absense of a Mumetal screening can for the transformer. The audible hum is quite negligible provided the transformer is at least 0.5 metre from ordinary mains equipment, so that a screening can is only marginally desirable. The two cylindrical 1mH inductors are mounted parallel to one another in such a manner as to give neutralization of hum pick-up.

The L_{eq} value for the Fig. 6 circuit, at maximum gain setting, is 37.5mH, which is over ten times the value for the circuit of Fig.3. To avoid the danger of overloading subsequent equipment when operating with high SPL's, a switched attenuator, with 10dB steps, has been incorporated. The output push-pull emitter-follower ensures that the accuracy of the time-constant corresponding to C_2R_3 in Fig. 5 is unaffected by the attenuator setting and/or the output loading.

The 100Ω and 100pF components on the input of the emitter-follower were found necessary for suppressing a tendency towards parasitic oscillation at a frequency in the region of 100MHz, due to a combination of wiring inductance in the base input and several pF's of shunt capacitance from emitters to earth. The alternative cure of a series base stopper resistor was not adopted as it would slightly worsen the noise performance at audio frequencies.

Fig. 8 shows measured overload characteristics for the Fig. 6 circuit at the maximum gain setting, with the d.c. supply set accurately to ±9V. The overload point is sharply defined, so that such a measurement is straightforward to make. It will be seen that an input of 20mA rms can be handled at low frequencies.

A very satisfactory dual version of the Fig. 6 circuit has been built using an NE5535N operational amplifier, which requires no compensating capacitor. The components, excluding the transformers, are mounted on a printed-circuit board measuring 70mm 96mm. The total current consumption, without the LED, is approximately 9mA.

The transformers use 0.015 ins. Mumetal (Permalloy C) laminations, ISCO No. 450, of size 1 ins. X ¾ ins. assembled, with ½ ins. stack. The centre limb is ¼ ins. wide. A single-section bobbin is used, with the 20T + 20T non-twisted-bifilar primary of 24 SWG (0.56mm) enam. sandwiched between two series-

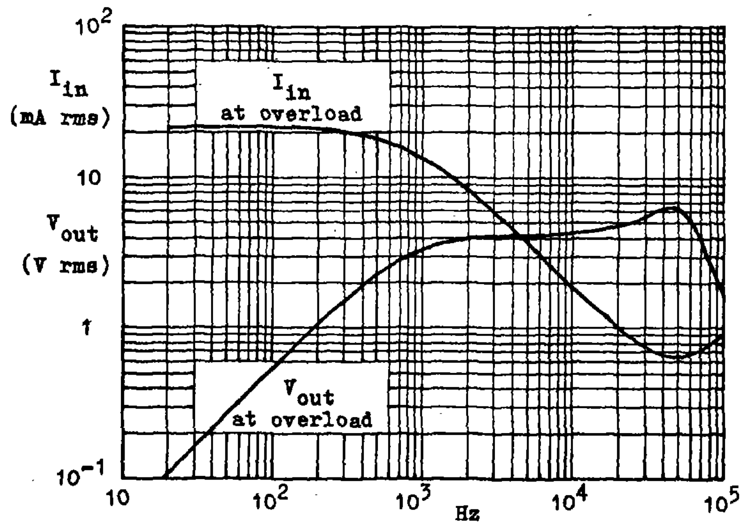

Fig. 8. Overload characteristics of Fig. 6 circuit.

connected 80T sections of 30SWG (0.315mm) enam. No interwinding insulation is used*.

In the initial experiments a readily-available commercial transformer was used - RS Components Ltd. stock no. 217-781 - with the primary and secondary sections parallel-connected to provide a ratio of 1 : 6.3. A very satisfactory performance was obtained, except that Z_{in} rose to nearly 2Ω at 15kHz, which is higher than really desirable.

The harmonic distortion introduced by the Fig. 6 circuit is much lower than that of Fig. 3, and is less than 0.01% measured at several frequencies throughout the audio band at levels about 3 dB below the overload point. The input transformer distortion, in contrast to that in a normal microphone amplifier, is greatly reduced by the shunt feedback.

* Some measured parameters are:- L_{sec} = 0.7H at low level, 60Hz. Total leakage inductance wrt sec. = 54μH. Shunt eddy-current resistance wrt sec. = 1.1kΩ at 60Hz. Copper resistances - pri. 0.165Ω, sec. 1.95Ω.

Applications

Most of the work done so far on this project has been concerned with the circuit-design aspects, and the following comments on applications are the result of preliminary thoughts and experiments only.

The technique enables a subjective quality assessment of a pair of nominally-identical loudspeakers to be made very straightforwardly and without the worry that microphone anomalies, particularly at very high frequencies, may lead to misleading conclusions. This is particularly advantageous for people not having access to measuring microphones of the highest grade.

Ideally, the loudspeaker used as the microphone should be in anechoic surroundings, with the sound source, e.g. voice, at the same distance* and orientation (usually on axis) as is the ear in relation to the other loudspeaker. Then the listener will receive a reproduction accompanied by twice the distortions introduced by a single loudspeaker, ignoring high-level non-linearities. By introducing a good tape recorder, quick A-B comparisons between the direct natural sound and the reproduced sound may be made. The same loudspeaker specimen may, of course, then be used both as the microphone and as the reproducing loudspeaker. It helps to have blank sections on the tape so that the background sounds are present all the time, particularly if the microphone-loudspeaker is used out-of-doors rather than in an anechoic chamber. The fact that the loudspeaker distortions are doubled helps to make them easier to assess subjectively.

When a first-rate loudspeaker is used as a microphone in the above manner, it is found that speech quality fully up to the standards of the best studio microphones is obtainable. This does not mean, however, that such results will be obtained in more normal circumstances of use, with off-axis sound sources and reverberant surroundings - indeed an ordinary loudspeaker must be regarded as a rather poor microphone when its directional

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* The axial frequency response of any normal loudspeaker is dependent to some extent on the measuring distance. This is an inevitable result for any non-point-source radiating system whose polar characteristic is frequency-dependent.

characteristics are taken into account. Nevertheless it is on the whole surprising what very good musical quality can be obtained, especially under stereo conditions, when the direct sound sources can be arranged to be more-or-less on axis for one or other of the two "microphones".

In testing line-source loudspeakers in the normal way, it is desirable to place the measuring microphone at a distance of at least 5 metres and preferably further. This almost inevitably means that the tests must be done out-of-doors. With an omni-directional microphone, because of traffic noise, birds etc., the loudspeaker must then be operated at quite a high level, causing considerable neighbour-nuisance in many circumstances. By using the line-source loudspeaker as a microphone, however, its own directional properties reduce ambient noise pick-up very considerably, especially if it is of the bi-directional type[10]. A subjective evaluation may then be made by simply talking at a suitably large distance and listening to the result on a good monitor loudspeaker.

The technique may be used, up to a point, for frequency-response measurements, but it is necessary to take certain effects into account when interpreting the results.

When a loudspeaker response is measured in the normal way, only those sound rays leaving the loudspeaker in the direction of the measuring microphone are significant, but with a large-area loudspeaker used as the microphone, rays leaving the sending loudspeaker in other directions become significant. Multiple reflections between the cabinets also introduce unwanted effects. Nevertheless, with ordinary loudspeakers, a reasonable approximation to the response as normally measured is obtained if the loudspeakers are placed facing eachother at a distance of 1 metre in anechoic surroundings, though with minor additional response fluctuations superimposed. The measurement may be extended above the upper frequency limit of normal measuring microphones.

The technique also enables the absolute sensitivity of a loudspeaker, for use either as a loudspeaker or as a microphone, to be determined in a conveniently straightforward manner.

On the assumption that the effective dimensions of the radiating elements are small compared with the wavelength, and that the spacing between the loudspeakers is large compared with these dimensions, it may be shown that:-

$$P = \sqrt{\frac{\rho V_{out} V_{LS}}{4\pi d L_{eq}}} \quad . \quad . \quad . \quad . \quad . \quad . \quad (8)$$

where P is the sound pressure in Nm^{-2} rms produced by the sending loudspeaker in free space at a distance d metres on axis, d being the effective spacing of the loudspeakers. V_{LS} is the voltage (volts rms) applied to the sending loudspeaker, and V_{out} (volts rms) is the output from the circuit unit, e.g. Fig. 6, L_{eq} being its equivalent inductance in henries. ρ is the density of air, $= 1.20kgm^{-3}$. (The approach used in deriving this result is similar to that on page 347 of reference 4.)

The loudspeaker sensitivity may be defined as Pd/V_{LS}, and equation (8) then yields:-

$$\frac{Pd}{V_{LS}} = \sqrt{\frac{\rho V_{out}\, d}{4\pi L_{eq} V_{LS}}} \quad . \quad . \quad . \quad . \quad . \quad (9)$$

One measurement then gives, from (8) or (9), the sound pressure at a distance d from the loudspeaker with a known voltage V_{LS} applied to it. The result can easily be expressed, if preferred, in terms of watts in 8 ohms for an SPL of 96dB at 1 metre, for example. A measurement of this type on a particular loudspeaker gave a result agreeing within 1 or 2 dB with the maunfacturer's independently-determined value.

It may further be shown that the sensitivity as a microphone, of the loudspeaker plus circuit unit, is given by:-

$$\frac{V_{out}}{P} = \frac{Pd}{V_{LS}} \times \frac{4\pi L_{eq}}{\rho} \quad . \quad . \quad . \quad . \quad (10)$$

and substituting for the loudspeaker sensitivity Pd/V_{LS} from (9) gives:-

$$\frac{V_{out}}{P} = \sqrt{\frac{V_{out} d \times 4\pi L_{eq}}{V_{LS}\,\rho}} \quad . \quad . \quad . \quad . \quad (11)$$

If a loudspeaker of the type in widespread use for high-quality sound reproduction some years ago, having much higher sensitivity than typical modern loudspeakers, is used as a microphone, fascinatingly high sensitivity is obtained, and considerably weaker sounds can be heard than is possible with good unaided human hearing - for example, a quiet wrist-watch has been detected at a range of nearly 2 metres!

Acknowledgement

The method of arriving at the reciprocal behaviour of a transducer by invoking the concept of an ideal omnidirectional capacitor microphone capsule, as in Fig. 1, was first brought to my attention by Peter Walker of the Acoustical Manufacturing Company, in an electrostatic loudspeaker context, and the value of this idea as an aid to clarity of thought is duly acknowledged.

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