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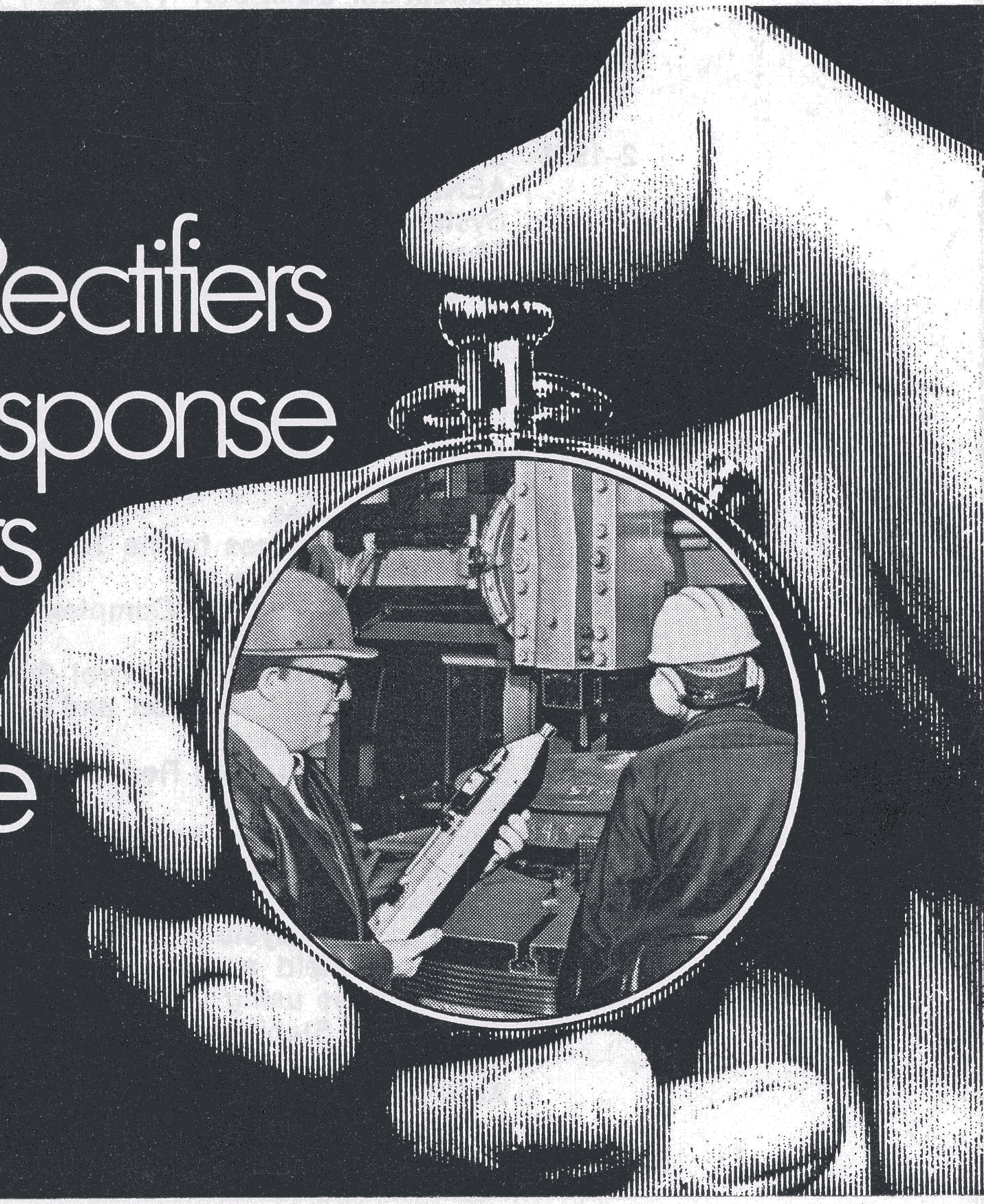
To Advance Techniques in Acoustical, Electrical and Mechanical Measurement



- RMS-Rectifiers
- Acoustic Response of Theaters
- Noise Dose

BRÜEL & KJÆR

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RMS – Rectifiers

by

J. Austin Hansen

ABSTRACT

During recent years Brüel & Kjær have released several devices which have, among other features the common trait that they contain some RMS detecting circuit.

This paper briefly explains the basic principles of these circuits and outlines the problems involved when large dynamic range, high crest factor capability and long averaging times must be achieved.

As these demands are contradictory, the advantages and shortcomings of several circuit configurations, such as idealized diodes and anti-distortion circuits, are discussed with reference to the instruments in which they are used.

The paper concludes that the optimum solution, both from a technical and an economic point of view, is to use circuits optimized for each group of instruments rather than one standardized circuit.

SOMMAIRE

Au cours des dernières années, Bruel & Kjær a produit de nombreux appareils ayant en commun, parmi d'autres caractéristiques, un circuit détectant la valeur efficace.

Cet article explique brièvement les principes fondamentaux de ces circuits et souligne les problèmes soulevés lorsqu'on veut avoir une grande dynamique, de longs temps d'intégration et lorsqu'on veut que le système admette des facteurs de crête élevés.

Comme ces exigences sont contradictoires, on discute les avantages et désavantages de plusieurs configurations de circuits, tels que les diodes idéalisées et les circuits anti-distorsion, en faisant référence aux appareils dans lesquels ils sont utilisés.

L'article conclut que la solution optimale, au point de vue technique comme au point de vue économique, est d'utiliser des circuits optimisés pour chaque groupe d'instruments plutôt qu'un seul circuit standard.

ZUSAMMENFASSUNG

In den vergangenen Jahren hat Firma Brüel & Kjær eine Reihe von Geräten herausgebracht, die neben anderen Eigenschaften das gemeinsame Merkmal haben, daß sie Schaltungen zur Erzeugung eines Effektivwert-Analogsignals enthalten.

Nach einer kurzen Erläuterung der grundlegenden Schaltungsprinzipien werden die Probleme behandelt, die sich aus den Forderungen nach einem großen Pegelbereich,

einem großen Scheitelfaktorbereich und langen Mittelungszeiten ergeben. Diese Forderungen sind teilweise einander entgegengerichtet, und es werden die Vor- und Nachteile mehrerer Schaltungsvarianten – mit idealisierten Dioden oder Entzerrungsnetzwerken –, jeweils in Bezug auf die entsprechenden Geräte, diskutiert.

Die Untersuchungen führen zu der Schlußfolgerung, daß es die technisch und wirtschaftlich günstigste Lösung ist, für jede Gerätegruppe eigens optimierte Schaltungen einzusetzen anstelle einer einheitlichen Schaltung.

Introduction

Several Brüel & Kjær instruments contain RMS-detectors, based on a principle first described by C.G. Wahrmann and further developed to optimize their properties for different applications.

This paper endeavours to describe briefly their function and to discuss the choice of optimal electronic circuits for the various demands on the detectors.

The basic principle

The principle employed is the same for all the detectors treated in this paper. A diagram of the basic circuit is shown in Fig. 1. Its function has been treated completely in the literature and it will only be discussed briefly in the following.

Assumptions:

1. The diodes are ideal, i.e. they conduct as soon as the anode voltage exceeds the cathode voltage.

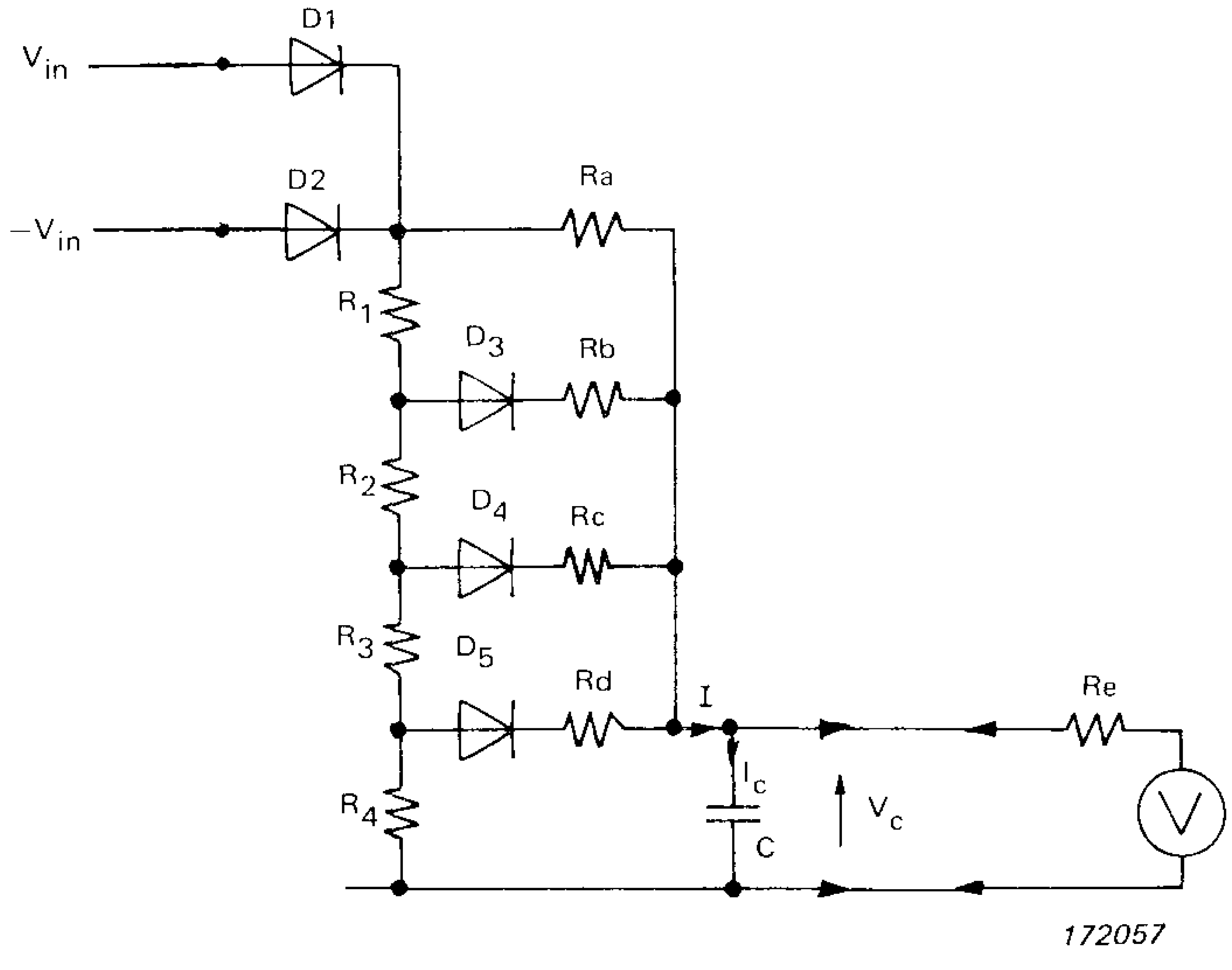

Fig. 1. Schematic diagram of the basic detector

2. The diodes are active, i.e. they draw no input current and they supply an output voltage equal to the input voltage when they "conduct".

3. A constant voltage V_c is maintained over the capacitor.

If the input voltage is initially zero, a negative current I_o must flow through $R_a + R_{1\text{-}4}$ resulting in a bias on D_1 and D_2 of kV_c, $k<1$.

If then the numerical value of the input voltage increases, I will remain constant and equal to I_o (see Fig.2), until the voltage reaches the value kV_c. Then one of the diodes D_1 or D_2 conducts and I increases linearly with the increase in input voltage (with a slope of the curve equal to $1/R_a$).

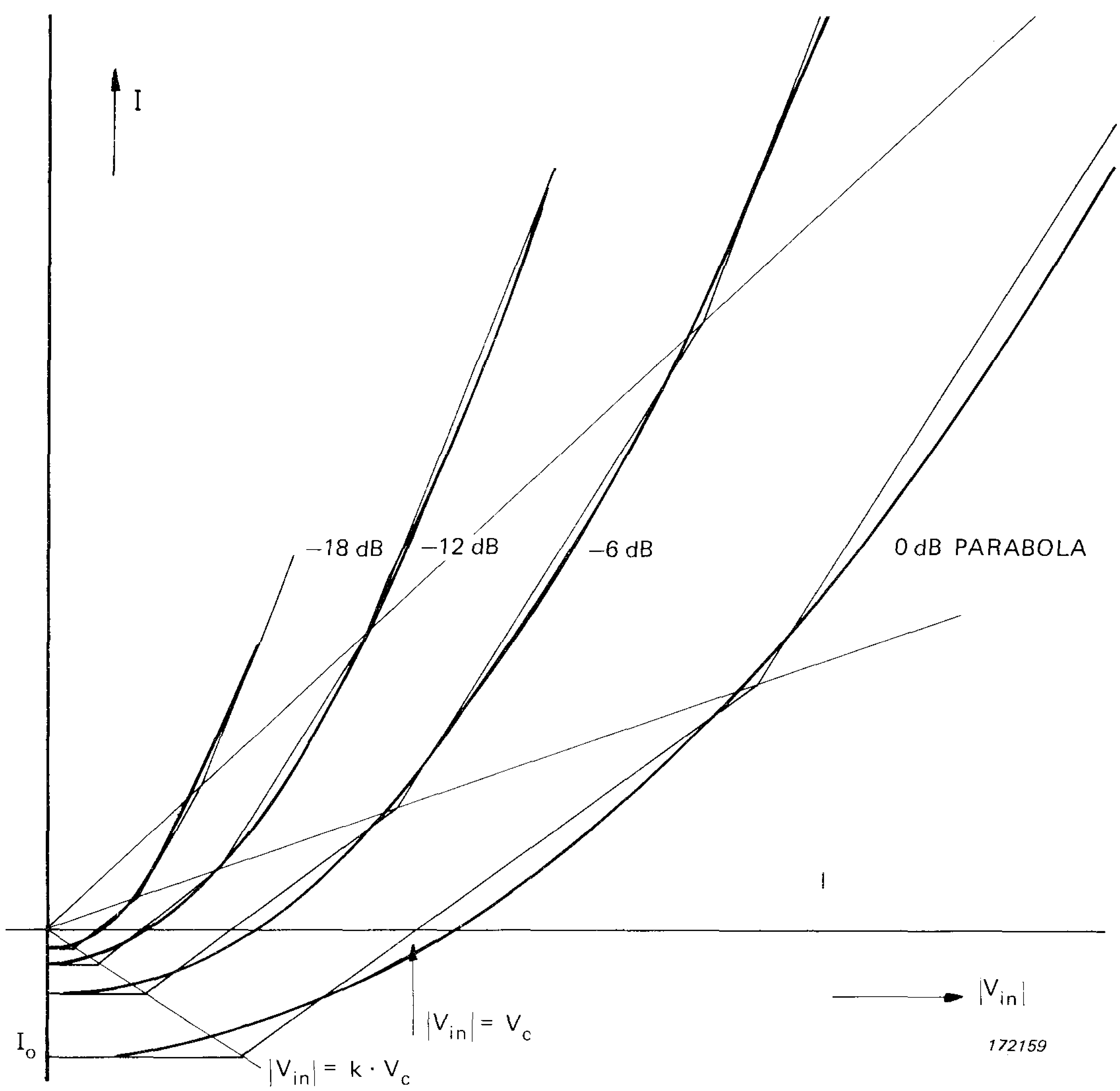

Fig.2. The I/ | V_{in} | polygons

When the input voltage $|V_{in}| = V_c$ then $I = 0$ as indicated on the figure. If the voltage continues to increase it will reach a value where D_3 becomes conducting, whereby the slope of the curve changes to $1/R_a + (1/R_b)\ (R_{2\text{-}4}/R_{1\text{-}4})$. A further increase in voltage may cause D_4 and thereafter D_5 to conduct.

The resulting $I/|V_{in}|$ curve is a polygon, which, by proper selection of the resistors, can be made to approximate a parabola, slightly displaced along the current axis.

As the numerical value of the input voltage is used in the circuit, only one half of the parabola is necessary. If the capacitor voltage is changed the polygon will also change in such a way that any point on the new polygon can be found by drawing straight lines from the origin to points on the old polygon and then multiplying the distance from the origin by the factor by which the capacitor voltage is changed.

The charge of the capacitor determines its voltage. The charge depends on the current drain I_m through the meter and on the current I from the square law input circuit, the constant of which is changed inversely proportional to the capacitor voltage. This feed-back provides the root-extraction necessary to obtain a capacitor voltage linearly dependent on the RMS value of the input voltage. The meter resistance and the capacitor, together with the other resistances, determine the integration time of the total rectifier circuit.

The fact that the polygon is displaced along the ordinate (current axis) has no effect on the function of the circuit, however, it has the important advantage that the first change of slope of the polygon is achieved at an input voltage which is considerably lower than the capacitor voltage. The negative current, for low input voltage, however, represents a reduction in the current available for the meter which should be considered when the meter sensitivity is selected.

Limitations to the basic circuit

Obviously the above is only valid if the input signal exceeds a value above which the forward voltage drop across the diodes can be ignored. If standard silicon diodes are used this source of error cause perceptible deviations from the ideal performance when the input voltage is below approximately 5 volts RMS. The error will appear as a nonlinearity in the dynamic characteristics as shown in Fig.3.

It is important to notice that the deviation is dependent on the shape of the input voltage as indicated by the several curves in the A-region.

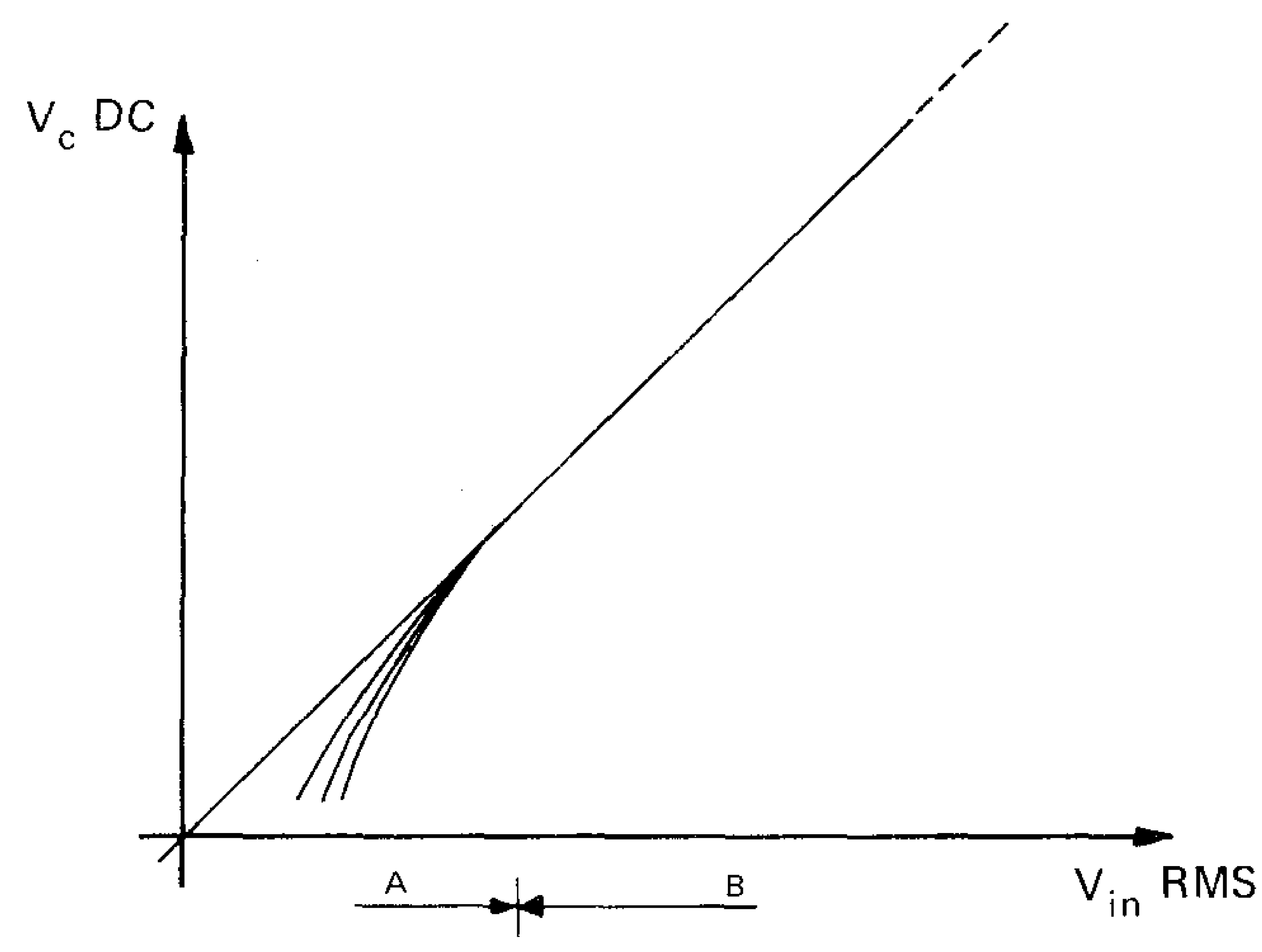

Fig.3. Dynamic characteristics for RMS detector with normal diodes. In the B-region the required approximation is fulfilled, but not in the A-region

If germanium diodes are used the lowest usable input voltage will be decreased by a factor of approximately 3. Unfortunately the higher (and very temperature dependent) reverse current will constitute an extra discharge path for the capacitor and thus affect the performance.

The highest crest factor which can be measured correctly with the circuit is determined by the design of the parabola. The higher the crest factor and greater the accuracy required of the parabola approximation the more diodes would be required. The circuit will have the same crest factor capability in the entire dynamic range unless special measures are taken to increase it at lower signal levels.

At levels higher than full scale meter deflection the tolerable crest factor will normally decrease due to overloading of the preceding amplifier stages, and reach unity at a level which is the normal crest factor times the full scale level. (In some RMS measuring systems the crest factor capability automatically increases as the input level is decreased).

The averaging time is determined by the capacitor and the resistors connected to the diodes and across the capacitor. To obtain long averaging times both the capacitor and the resistors must be made large. The maximum capacitor value is limited by the available space since electrolytic capacitors will normally not be usable. The resistors cannot be made arbitrarily high without affecting the performance of the diodes. Since the component values are then limited it may be necessary to modify the basic circuit to obtain long averaging times.

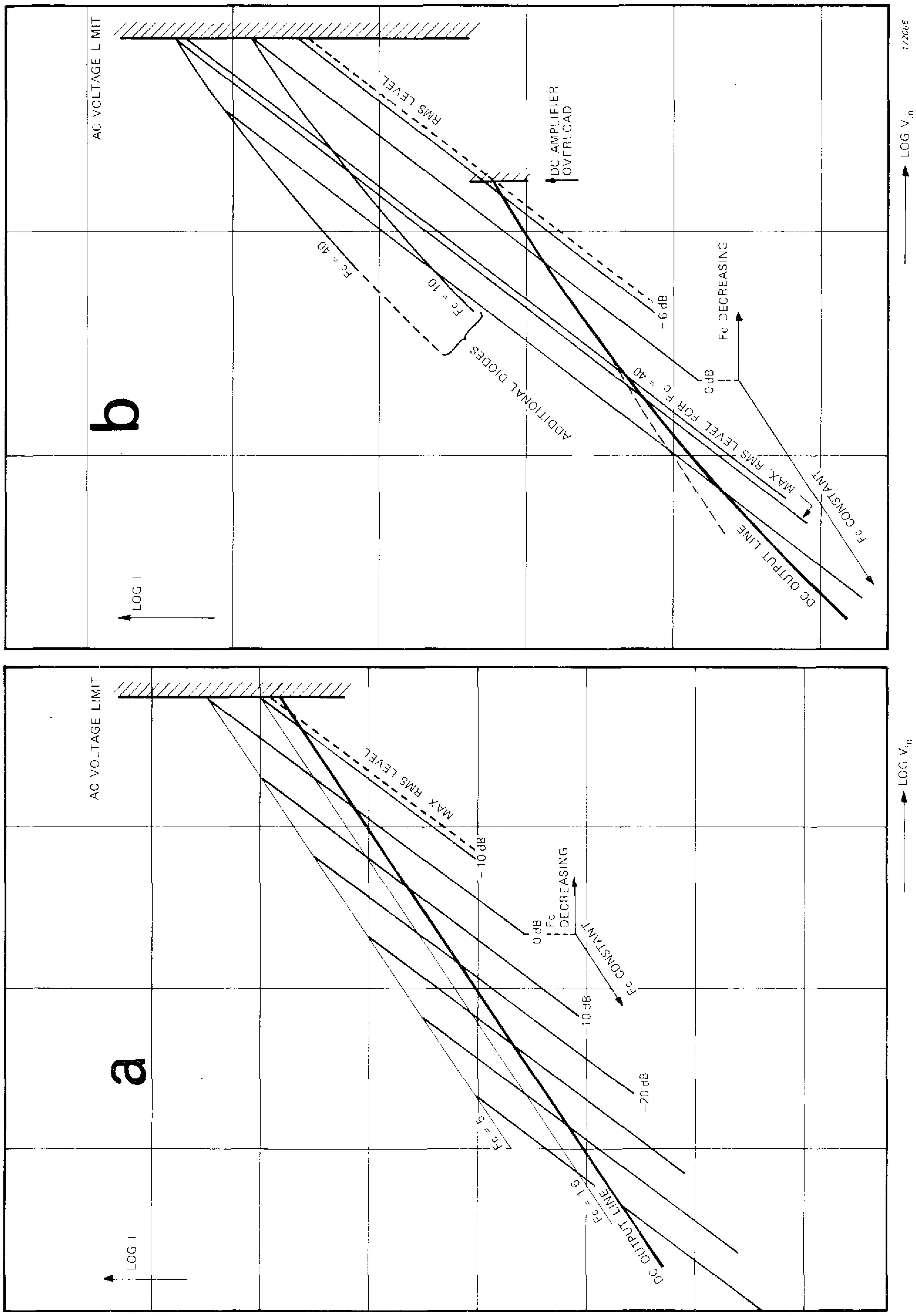

Fig.4. Log-log diagram of rectifier parabolas
a) large dynamic range (case 2) low crest factor
b) large dynamic range (case 1) high crest factor

Definition of dynamic range
Two definitions of the dynamic range can be used dependent on whether it is used for the indication on some read out device or for some analogue output.

Case 1:
The dynamic range is the ratio between the largest and the smallest signal of any shape (not exceeding the permissible crest factor at full scale level) for which there exists no deviation from proportionality between the input signal and the *reading* greater than some specified value.

Case 2:
The dynamic range is the ratio between the largest and the smallest signal of any shape (not exceeding the permissible crest factor at full scale level) for which there exists no deviation from proportionality between the input signal and *the analogue output signal* greater than some specified value.

It is clearly seen that the second definition is the most restrictive. The difference between the definitions may be more clearly seen from Fig.4a and 4b where the parabolas are depicted with log-log scales. These figures show the performance of two different RMS-converters.

Fig.4a indicates a detector with a large dynamic range after the case 2 definition, but with a relatively low crest factor capability. The shaded lines indicate limitations in voltage from the amplifiers.

Fig.4b indicates a converter which has some nonlinearity on the DC-voltage which may, however, be compensated by a suitable calibration of the scale on the indicating device making the case 1 definition usable. Fig.4b further shows how increased crest factor capability is achieved at lower deflections by using additional diodes, and how the output voltage may be limited by a DC amplifier after the converter.

The DC output voltage is found on the input voltage scale vertically below the crossing of the DC output line and the parabola which is valid for the existing input signal.

An example of Fig.4a performance is the Measuring Amplifier Type 2607 and of Fig.4b Type 2606 and 2204.

Methods to obtain a wide dynamic range
The diodes indicated in the diagram of Fig.1 may take many forms. They can be normal passive PN-junctions, or transistors, or they can be active

circuits with diode properties. If they are perceived as ordinary PN junctions the diagram of Fig.1 cannot be used directly because the charging current will have to be drawn through the voltage divider. In this case the divider must be split as it was done in the Measuring Amplifiers, Types 2409, 2603/04 and other instruments. The use of ordinary diodes has the disadvantage that the amplifier must be capable of delivering a high peak current when a signal of high crest factor is measured. One further disadvantage is that the load is voltage dependent which makes it adverse to use a transfer capacitor.

Transistors as diodes

The diodes can be replaced by transistors. In this case the high peak currents will be drawn from the supply voltage and the load seen from the amplifier can be linear, and larger than otherwise possible.

It is possible to extend the dynamic range of this circuit slightly by forward biasing the transistors e.g. by applying a small positive voltage at both inputs and/or by lifting the "bottom" of the voltage divider to a slight positive potential. One disadvantage is that with zero input signal, a small forward current will still flow to the capacitor and, consequently, the output will never be zero.

Compensating pre-amplifier

The drawback mentioned above can be overcome by using more advanced techniques: The amplifier which feeds the RMS converter is designed to amplify those parts of the input wave, to which too little weight is ascribed, more than the rest of the cycle. Then some of the nonlinearity of the converter is compensated for. An amplifier with these properties is shown in Fig.5.

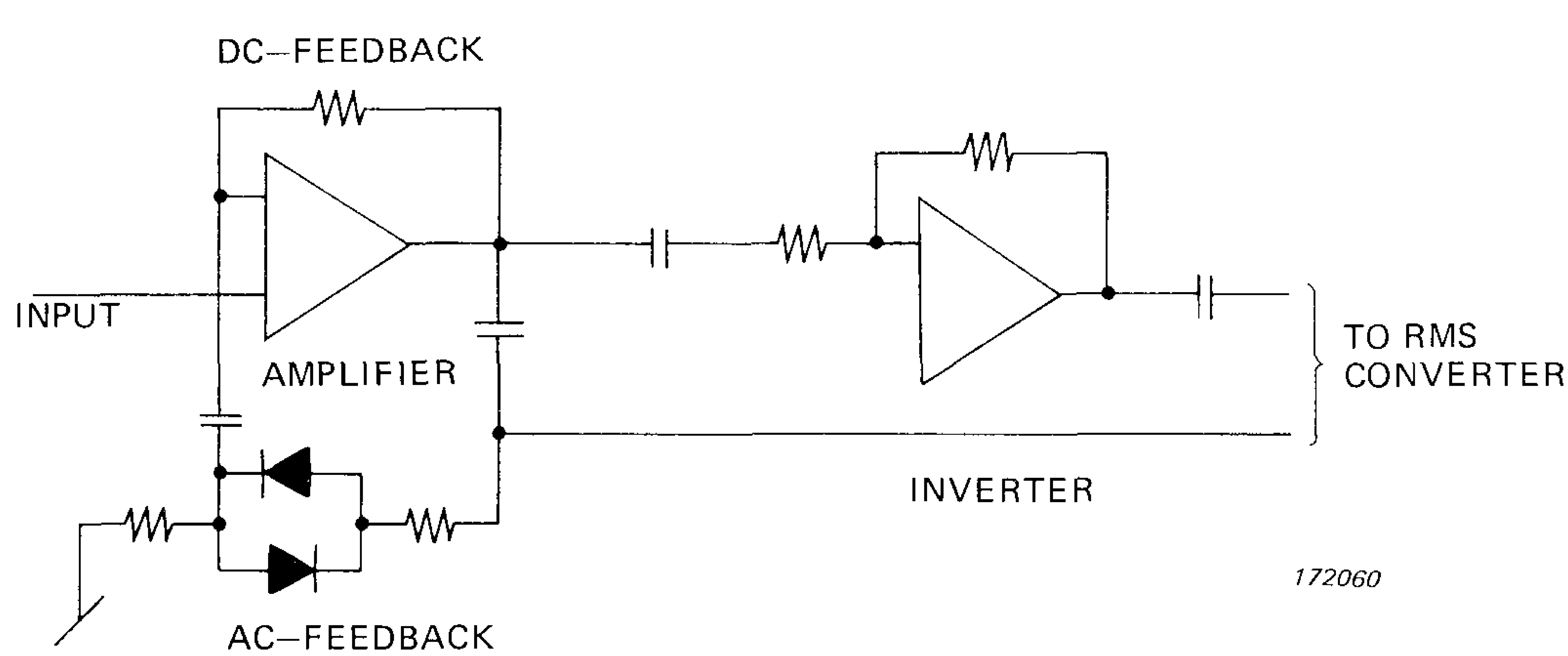

Fig.5. Compensating preamplifier

When the output voltage is approaching zero the impedance of the AC feed back diodes will increase and the gain will be close to the open loop value at the point where zero is reached. This counter-distortion technique involves the disadvantage that the amplifier voltage cannot be used as the AC output from the instrument. In the Sound Level Meter Type 2204 where the technique is used the problem is solved by shorting the diodes and disconnecting the meter when an AC output is required. In the Measuring Amplifier, Type 2606 and in other instruments which incorporate its circuits an extra output amplifier is used, by means of which the AC output and the corresponding meter reading can be obtained simultaneously.

In the above mentioned instruments, a dynamic range of 12 dB (case 1 definition) is obtained with an accuracy of ± 0.5 dB and for crest factors up to 10. If the tolerable deviation is increased to ± 1 dB the dynamic range is 20 dB.

In the design of the parabola special measures are taken to make the measurable crest factor increase with decreasing levels. This makes it possible to measure crest factors up to 40 at 12 dB below full scale meter deflection with a tolerance of ± 2 dB. By designing the parabola in this way the important advantage achieved is that erroneous measurements are impossible to perform within the upper 12 dB without overload indication.

Idealized diodes

A further improvement can be achieved if instead of en-bloc compensation the diodes are individually compensated. This leads into the field of active idealized diodes which may, but need not necessarily contain a diode or another rectifying PN junction. An example is shown in Fig.6.

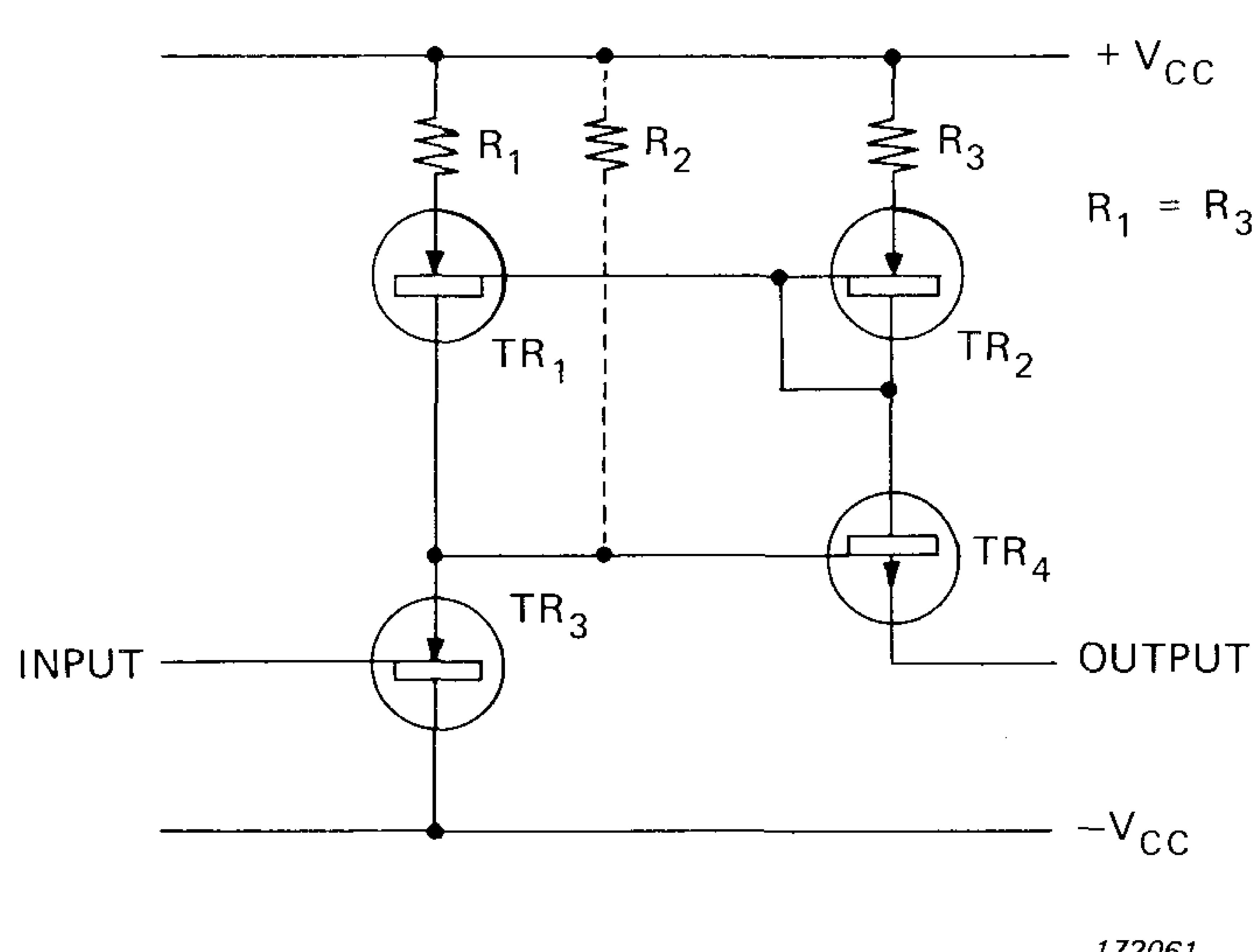

Fig.6. Idealized diode

The principle of this circuit is based on the assumption that if the currents flowing through two different PN junctions are kept equal the voltages across them will also be equal. The current in TR_4 is determined by the input voltage and by the load resistor. The same current produces a voltage drop across R_3 and the diode connected TR_2. Hereby the current in TR_1 and TR_3 are fixed to be equal to that in TR_2 and TR_4 since $R_1 = R_3$. It is seen that in the current range where the above assumption is valid, the two basis-emitter voltages of TR_3 and TR_4 will compensate for each other.

A very large resistor R_2 must be connected as shown, to ensure that TR_3 will always conduct slightly even if the idealized diode is reverse biased. The active idealized diode is used in the exciter control, Type 1026 to produce an RMS detector with a dynamic range (both case 1 and 2) of 20 dB with an accuracy of 1%. Crest factors up to 5 can be measured with errors not exceeding 2%. The frequency range of the diode is limited to below 20 kHz.

It is important to note that if "ideal" diodes are used it may be reasonable to isolate the specification of the tolerance of the linearity and the parabola approximation, especially, if as in this case the linearity error is smaller than the error arising from the approximation to the parabola.

The discussion so far has dealt only with what could be called open loop methods to obtain a widening of dynamic range since they all use one element to perform the desired action and another to compensate for the nonlinearity.

Closed loop idealized diodes

The remainder of this section will be devoted to a discussion of closed loop systems where the acting and compensating element is the same, if the functions can both be pointed out. These circuits are all different attempts to develop a circuit with diode properties.

One must realize that the forward voltage drop which may be negative or, in other words, the resolution is not the only quality of interest. It is important that no current should flow in the reverse direction (equivalent to the leakage current of normal PN-diodes). Furthermore the diodes used in the full-wave rectifier must be able to withstand a reverse voltage greater than twice the maximum peak value which can be delivered by the amplifiers.

For the other diodes it is sufficient if they can withstand the maximum capacitor voltage. The diode shown in Fig.6 is not able to stand a reverse voltage greater than the B-E breakdown potential of TR_4. The reverse current is also determined by TR_4.

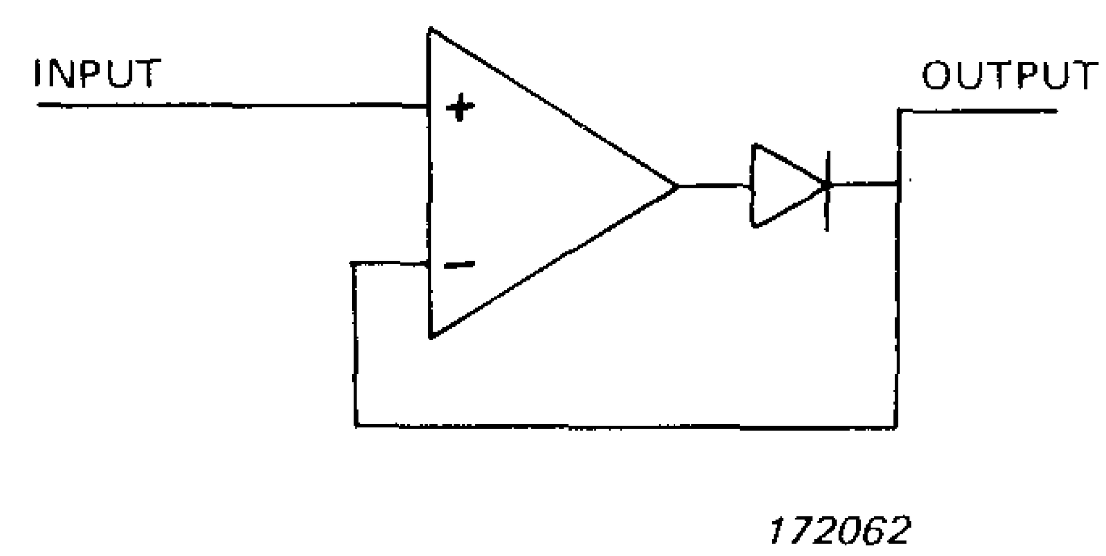

Fig.7. Idealized diode with operational amplifier

One very obvious way to make an ideal diode is shown in Fig.7. However, this circuit suffers from several disadvantages. If a normal operational amplifier is used, a current of twice the normal bias current must flow into the inverting input when the diode is off. An input overvoltage protected operational amplifier cannot be used since this safeguard is normally performed by two PN diodes across the inputs. This will cause the total diode to show normal forward biased PN diode characteristics in the reverse direction. Therefore an unprotected amplifier must be used which will limit the maximum signal peak to approximately 3 volts, for the full-wave rectifiers. Finally the frequency range is limited which can be seen from the following: when the diode is conducting the circuit has unity gain and the frequency compensation circuit must be selected to fit this case. When the diode is off the amplifier is in open loop condition with unity gain compensation which leads to very slow response. Besides the amplifier is driven heavily into "saturation" which causes delay. As a result the output will be cut at the leading edge of the halfcycle as indicated in Fig.8. This effect increases at higher frequencies. If the open loop gain is reduced, the cut-off frequency can be increased and the frequency range will be widened but at the same time the "linearization" of the PN diode will also be reduced.

If 2 volts peak is available and a detector with a dynamic range of 50 dB and crest factor capability of 10 is demanded, the resulting diode must have a resolution of less than 100μV since the smallest signal has a RMS value of less than 1 mV. A gain of more than 5000 would be necessary. The unavoidable nonlinearity (case 2) can be compensated by a DC offset adjustment in

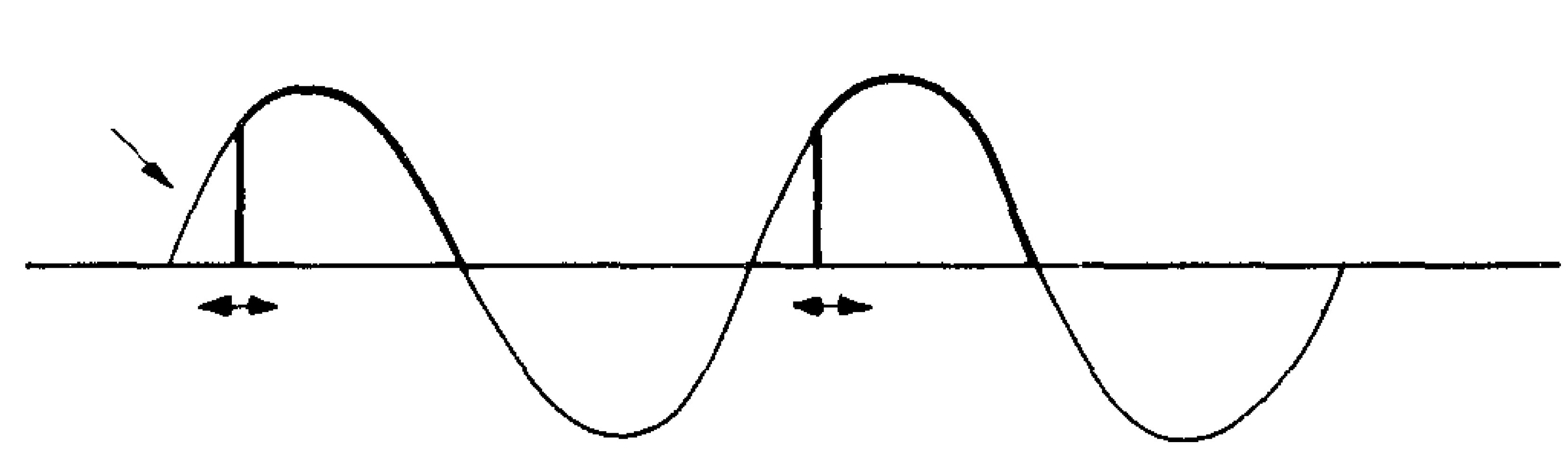

Fig.8. Distortion of the signal in the circuit of Fig.7

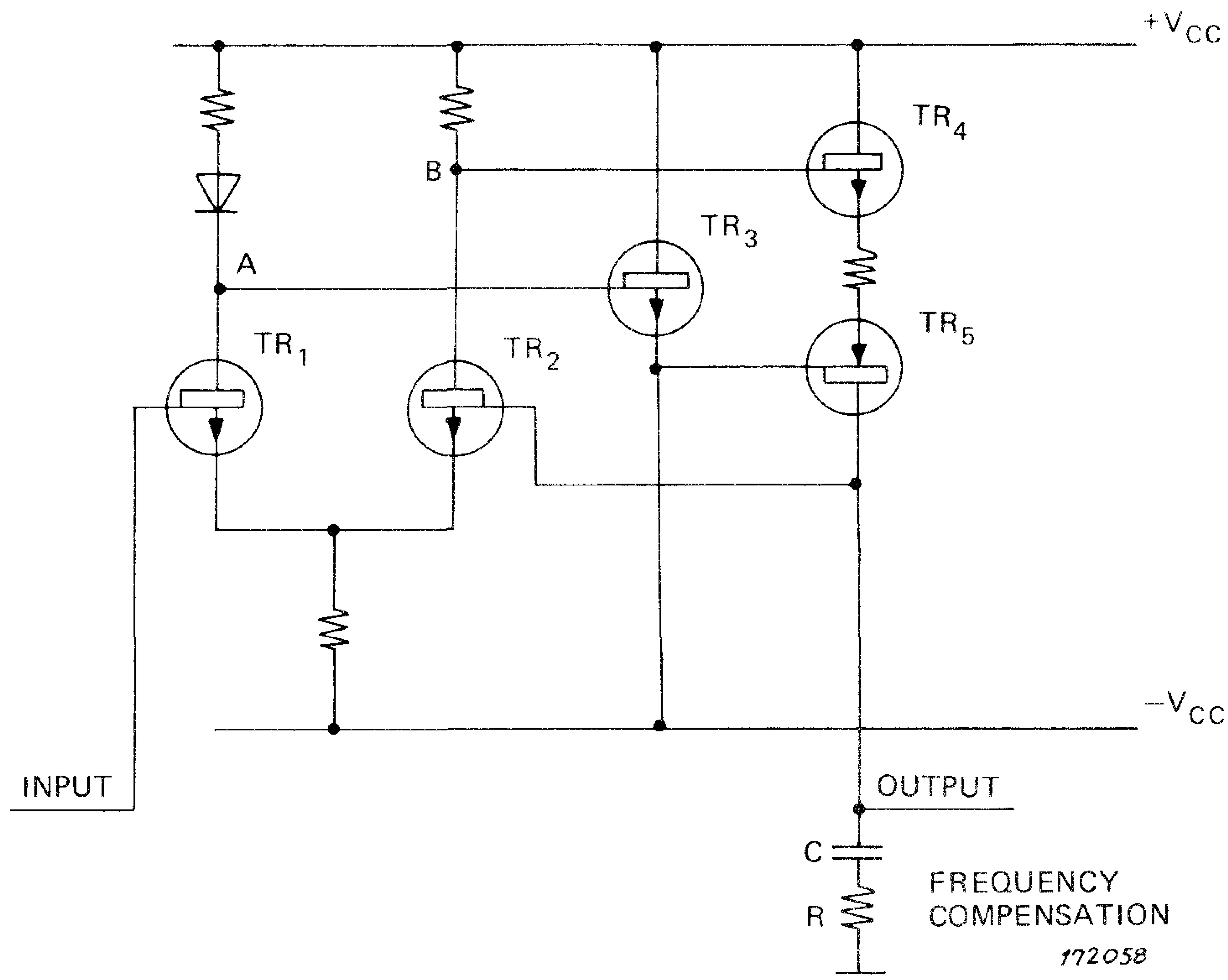

Fig.9. Idealized diode with external frequency compensation

some subsequent DC amplifier. However, the circuit of Fig.7 can be realized in another way which removes its limitations. A fundamental diagram is shown in Fig.9.

The idea is to place the frequency compensating components after the rectifying junction so that no high frequency phase shift can limit the frequency range. The B-E diode of TR_5 takes care of the rectification. When $V_i > V_o$, TR_5 will conduct and the circuit is a compensated unity gain amplifier. When $V_i < V_o$, TR_5 is cut-off and the gain is open loop and uncompensated. To increase the common mode rejection and the gain without sacrificing maximum output swing, TR_5 is connected as a single transistor differential amplifier.

It is obvious, from the position of the frequency compensation circuit that if the frequency is increased or if a bigger capacitance is used, the diode will start to act as a peak rectifier which causes the converter output to increase. Although C is small an inconvenient error is present at about 0.1 MHz. The slope of the frequency curve makes it difficult to compensate by the aid of normal R-C combinations. However, it is now possible to convert the frequency limiting effect of the "normal" compensation between A and B in Fig.9 to an advantage. By connecting a suitable RC circuit between these two points the frequency range can be extended to 0.5 MHz with errors less than 3%.

The problem of reverse current which was mentioned in connection with Fig.7 can be overcome by using a matched pair of FET transistors connected as source followers to feed TR_1 and TR_2. This gives the further advantage that the B-E breakdown in the input transistors is no longer destructive since the current is limited by the large FET source resistor. When breakdown occurs the FET is biased off. Consequently the peak voltage can be increased to at least 75% of $+V_{cc}$.

Automatic ranging system
A situation can arise where the best available diode is not able to give the required dynamic range. This is most likely to happen if the frequency range must also be wide since the two demands are often contradictory. In this case the problem may be solved by using automatic ranging.

When the signal falls below some pre-determined value an extra amplifier is coupled in (or an attenuator is coupled out) before the circuit, and at the output the opposite operation is performed leaving the overall gain unchanged.

At the instant of level change it is necessary to change the capacitor voltage to make it suitable to the new condition. This can be done either by changing the potential at the "grounded side" or by actually charging or discharging the signal side of the capacitor. The first method requires that the shift takes place at exactly the same level for both increasing and decreasing signals. However, this may cause the attenuators to operate continuously which is undesirable.

If the second method is used, a hysteresis of any size can be chosen, with a sacrifice of dynamic range equal to the hysteresis. If a 30 dB circuit is used with 5 dB hysteresis a total range of 55 dB is obtained.

The described diode and automatic ranging system is used in the Measuring Amplifier, Type 2607 which has a crest factor capability of 5 and a dynamic range of 50 dB with an accuracy better than ±0.5 dB. (case 2).

The output DC amplifier can handle a signal 10 dB larger than the level corresponding to full scale deflection and consequently a dynamic range of 60 dB is available for crest factor of 1.6 (sine: 1.4).

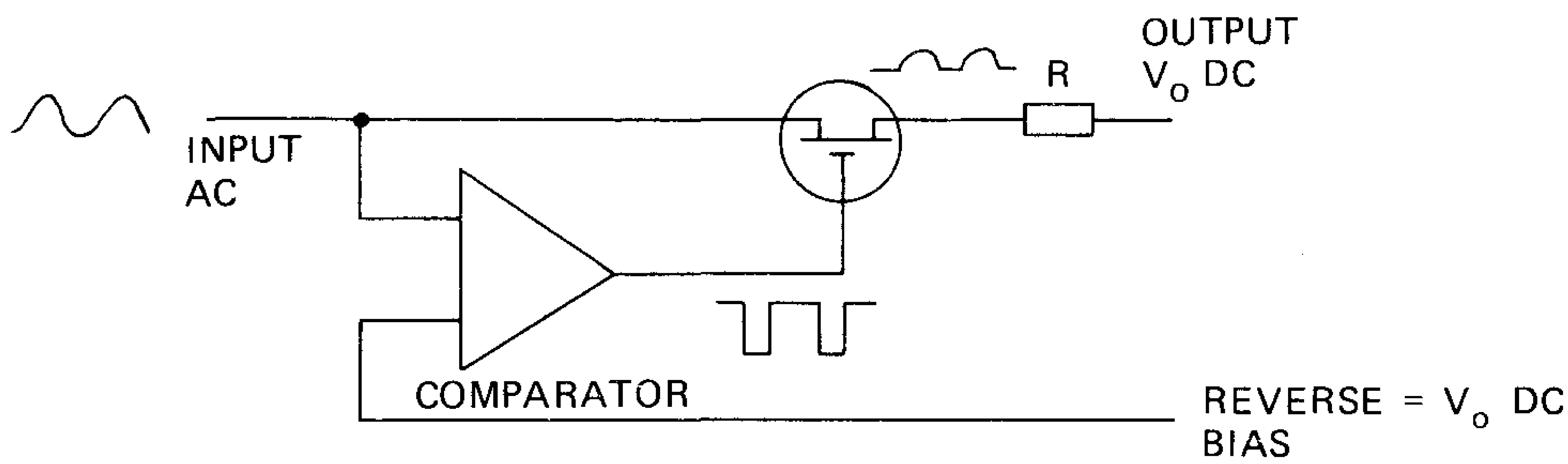

Fig. 10. Idealized diode without PN junction in the signal path

In the last diode to be treated the use of PN junctions in the signal path is avoided. A simplified diagram is shown in Fig. 10. The function of the circuit is self-explanatory. The dynamic range is limited by the resolution of the comparator and the error signal, which is coupled from the gate to the output terminal via the capacity between these points. This error is of course frequency dependent which may cause a frequency limit. To some extent the error can be compensated for by a signal of opposite polarity with respect to the output connected to the resistor side of the MOS-FET transistor, via a small capacitor. This diode is used in the Frequency Analyzer Type 2130 and the Heterodyne Slave Filter Type 2021. In the analyzer Type 2130 some assumptions can be made regarding the shape of the input signal since it must pass a narrow band filter; consequently, it is sufficient to use one diode in the rectifier (i.e. only to measure one of the halfwaves) because the signal will always be symmetrically distributed around the zero line. Further, a three straight line approximation to the parabola (i.e. two diodes) is sufficient. The dynamic range is 50 dB (case 1). The tolerance is specified in a number of different situations and ranges which makes it difficult to compare.

The circuit for the Slave Filter, Type 2021 is a full wave quasi detector making it suitable for measurements on noise and sinewaves. The dynamic range is 50 dB with a tolerance of ±0.2 dB on the linearity. The error introduced from the parabola approximation is examined in the reference. Neither in the 2130 nor in the 2021 are automatic attenuators used.

Extension of averaging time range

It is important that long averaging times are available for certain applications of RMS detectors (e.g. measurement of low frequency narrow band noise). To obtain this, some special measures must be taken. Since the performance of most of the idealized diodes discussed in the text depend severely upon the load resistor, there are serious limitations in its value.

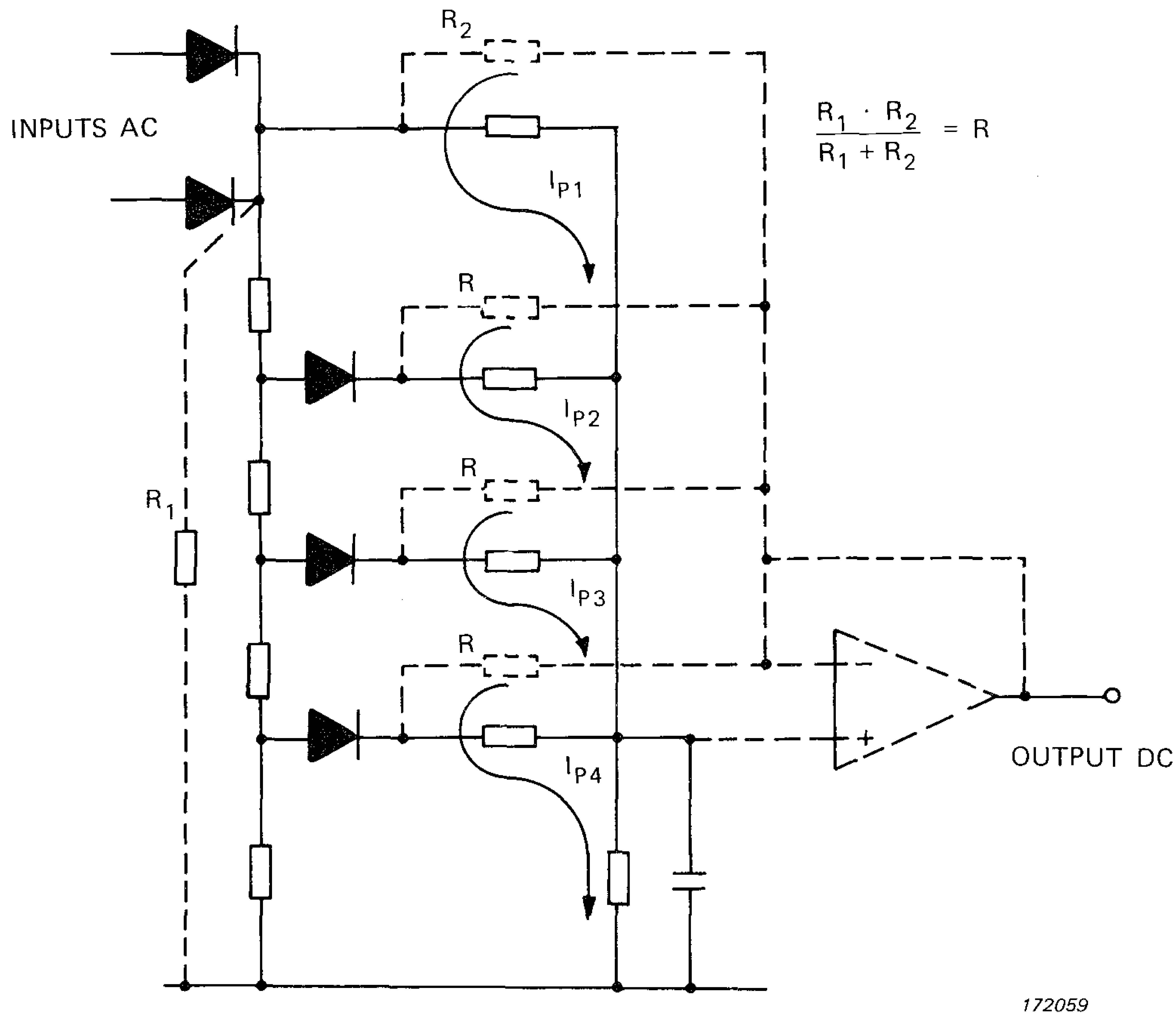

Fig. 11. Modified basic detector circuit

Fig. 11 is a modified version of the fundamental diagram of Fig. 1. The changes are indicated by dotted lines. As it is seen from the diagram the resistors R provide for the low impedance loading of the diodes, but since the capacitor voltage is present on both sides the resistors take no part in the charge or discharge. The parabola shaping resistors can be made very large and are only limited by second order effects such as surface leakage in the printed circuit board, amplifier input current, etc.

One serious disadvantage is evident from the diagram. The unity gain amplifier has positive feedback via the paths marked I_p. This causes the temperature drift to increase by a factor determined by the highest crest factor since the voltage division approaches unity as the crest factor is increased. All of the instruments mentioned which have long averaging times, use the outlined technique and none of them have crest factor capabilities greater than 5.

Conclusion

It has been shown that it is difficult to obtain wide frequency and dynamic range simultaneously and that the same problems are associated with high crest factors and long averaging times combined in the same circuit. Therefore it is believed that it is still necessary to develop RMS-converters individually in order to equip each new instrument with a detector most suitable to solve the type of problem it is intended for.

Reference:

C.G. Wahrman: A true RMS Instrument, B & K Tech. Rev. No. 3–1958.

Scandinavian Efforts To Standardize Acoustic Response in Theaters and Dubbing Rooms

by

Erik Rasmussen

ABSTRACT
After some years of work on the subject, a draft proposal for sound reproduction in cinemas was released by the Scandinavian Film- and T.V. Society in 1969. Since that time 100 public cinemas have been measured in accordance with the specifications in the proposal. This paper describes the technique used and presents the results, to date.

SOMMAIRE
Après quelques années de travail sur le sujet, une proposition de norme pour la reproduction du son dans les cinémas a été publiée en 1969 par la Société Scandinave du Film et de la Télévision. Depuis lors, on a effectué des mesures conformes aux spécifications de cette proposition dans 100 cinémas publics. Cet article décrit les techniques utilisées et donne les résultats obtenus à ce jour.

ZUSAMMENFASSUNG
Nach mehrjährigen Vorarbeiten wirde 1969 von der Skandinavischen Film- und Fernsehgesellschaft ein Vorschlag für die Schallwiedergabe in Kinos im Entwurf freigegeben. Seither wurden 100 öffentliche Kinos in Übereinstimmung mit den Spezifikationen dieses Vorschlags ausgemessen. Dieser Beitrag beschreibt die Technik und präsentiert die bisherigen Resultate.

As reported in 1969[1], work on acoustic response measurements of cinemas and dubbing rooms began in Denmark in April 1968 under the auspices of the Danish Filmfoundation. At the end of 1969 we had measured the acoustic response of 25 cinemas. Since that time we have expanded our measurements to 100 cinemas, improved our measurement techniques and generated Document NFTU-1069, "Recommended Practice for Measuring Overall Frequency Response in Cinemas for Optical Sound, 35 mm.

Very early in our work a close collaboration was initiated with interests in Sweden, where Lennart Ljungberg has done a considerable amount of work on acoustic response.[2] He splits the amplifier chain into two parts: "A" and

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"B". The "A" part includes the optical (or magnetic) sound head, equalizers and preamplifier and is purely electrical. Its frequency characteristic can be measured with conventional multifrequency test films. The "B" part of the chain is represented by the power amplifier (normally flat), the loudspeakers, screen losses and room acoustics.

The reason for splitting the amplifying chain into two parts is evident when one considers that the "B" part is constant and does not change whether the "A" part consists of an optical reproducer, magnetic soundtracks, disc records, tape deck, etc. Because of the various equalization practices employed in different countries and in different studios, it was believed that any attempt at standardization should begin with the "B" part. The overall system response is obtained by combining parts "A" and "B".

In Denmark we wanted to measure a large number of theaters, and since the majority of feature films are still released with optical soundtracks on 35 mm film, we elected to measure the overall acoustic frequency response in one operation, i.e., from the soundtrack to the patron's ear.

For this purpose we simply replaced the normally used APFA multifrequency test film with a new one containing octave filtered "pink-noise" with center frequencies from 31 Hz to 8 kHz.[3] The conventional voltmeter was replaced with a precision sound-level meter in the auditorium.

To avoid uncertainties in printing, the pink-noise test film is a direct positive recording, individually calibrated on an electrically flat playback head with

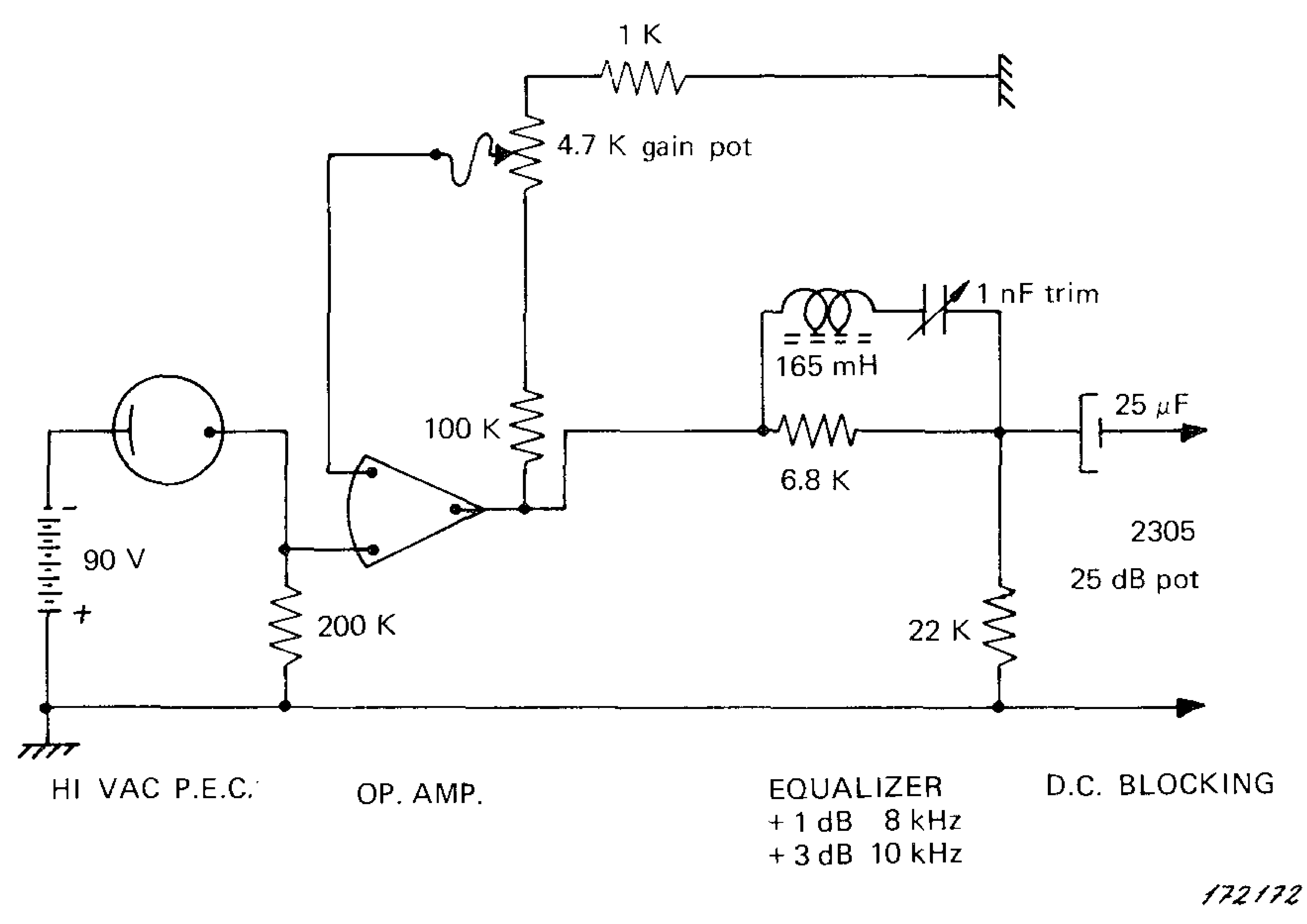

Fig. 1. Calibrated playback amplifier

1/4-mil slit (Fig.1). At this writing some 40 test films have been made and a third of these delivered to the USA. The calibrating reproducer is checked with the SMPTE laboratory type APFA test film.

There has been some academic discussion about the 8-kHz octave band which nominally extends from about 5.6 to 11.2 kHz. Since this band is limited to the roll-off characteristics of the specific recording galvanometer above resonance frequency (RCA MI 10-717), it was believed that the upper third of this octave (9.0 to 11.2 kHz) would not be filled in. Accordingly, the 8-kHz octave was raised about 2 dB in recording so as to play back flat on the calibrated reproducer.

An arbitrarily chosen specimen of our test films has recently been analyzed using one-third octave filters and with the following results: 6.3-kHz band + 2 dB, 8-kHz band + 1 dB and the 10-kHz band −4 1/2 dB, all referred to the 1-kHz output, again with some uncertainty as our reference test film APFA ends at 10 kHz, the playback response from 10 to 11.2 kHz being unknown. It was suggested that we record the test film at some lower speed, say 60 or 45 ft/min to overcome these difficulties in the 8-kHz band, but it has not yet been put into practice. A wide response recording galvanometer is another proposed solution.

In December 1969, the Scandinavian Film- and Television Society (NFTU) distributed Document NFTU.1069 as a Recommended Practice. This Document specifies the test film, the sound-level meter, measurement positions, averaging method, overall acoustic response curve and acceptable tolerances.

The first real problem was to establish acoustic response tolerance limits for each of the nine octave noise bands (Fig.2). The question was one of determining what could be accepted as "good". The first 25 theaters measured did not provide much guidance since deviations of as much as 10 dB from the ideal curve were observed in some important octave bands. Based on a moderate amount of measurement data and many years of listening experience, we settled for these limits:

63 Hz,	±6 dB
125 Hz–4 kHz,	±3 dB
8 kHz,	±4 dB

The 31-Hz band is not significant and is included only to see that it does not rise, especially compared to the 63-Hz response, since this region is sensitive to specific low-frequency components inherent in the operation of the noise reduction shutter and microphone boom. The solid line in Fig.2 is the desired response, while the shaded area defines the acceptable tolerances.

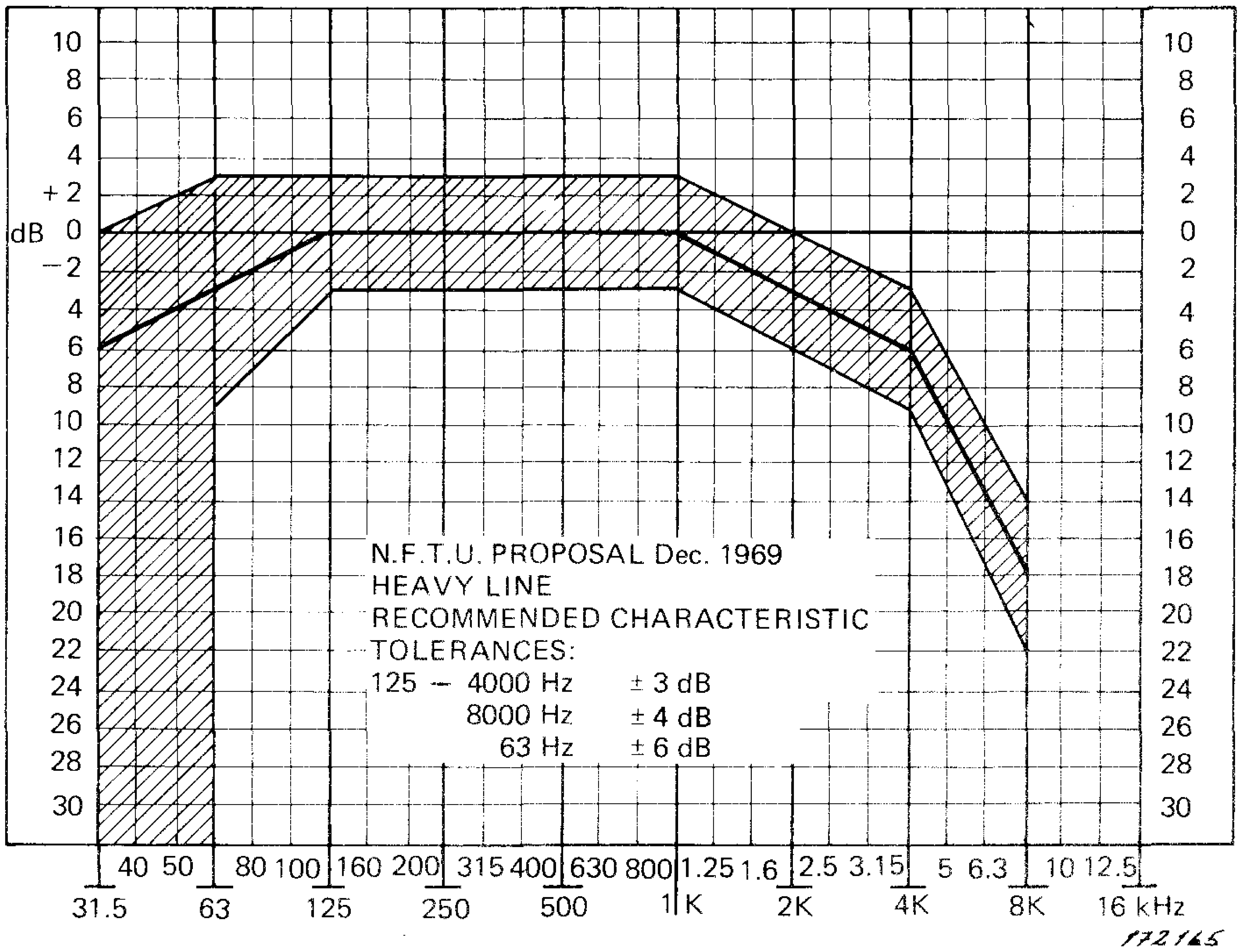

Fig.2. NFTU Proposal-1069

The placement of the microphone was given much consideration. Obviously, a one-point measurement alone would not give complete information. As is the case in the stereo reproduction, where theoretically an infinite number of channels should be used, practical limitations restrict the number to from two to five. We settled for five different microphone positions for each seating area. A cinema may have only one seating area, the main floor; on the other hand, it may have as many as three or more seating areas: main floor in front of balcony, floor area under balcony and one or more balconies. Since the frequency response is usually different for each of these areas, the five readings for each area must be averaged separately, since an average taken over all areas may hide useful information.

Microphone placement is shown in Fig.3. It is adjusted to be at ear height with the microphone pointed toward the screen. In our case we use a one-inch free field type of microphone fitted with a random incidence corrector. A cautionary note is included in the Recommended Practice about acoustic defects. Any attempt to use the acoustic response data to adjust electrical equalization must be preceded by a determination that the theater does not have serious acoustical faults. The measurements deal only with the steady-state properties of the auditorium, and such defects as back stage hang-over, harmful echoes and the like, do not show up. Methods for finding or eliminating such faults are not covered in the Recommended Practice.

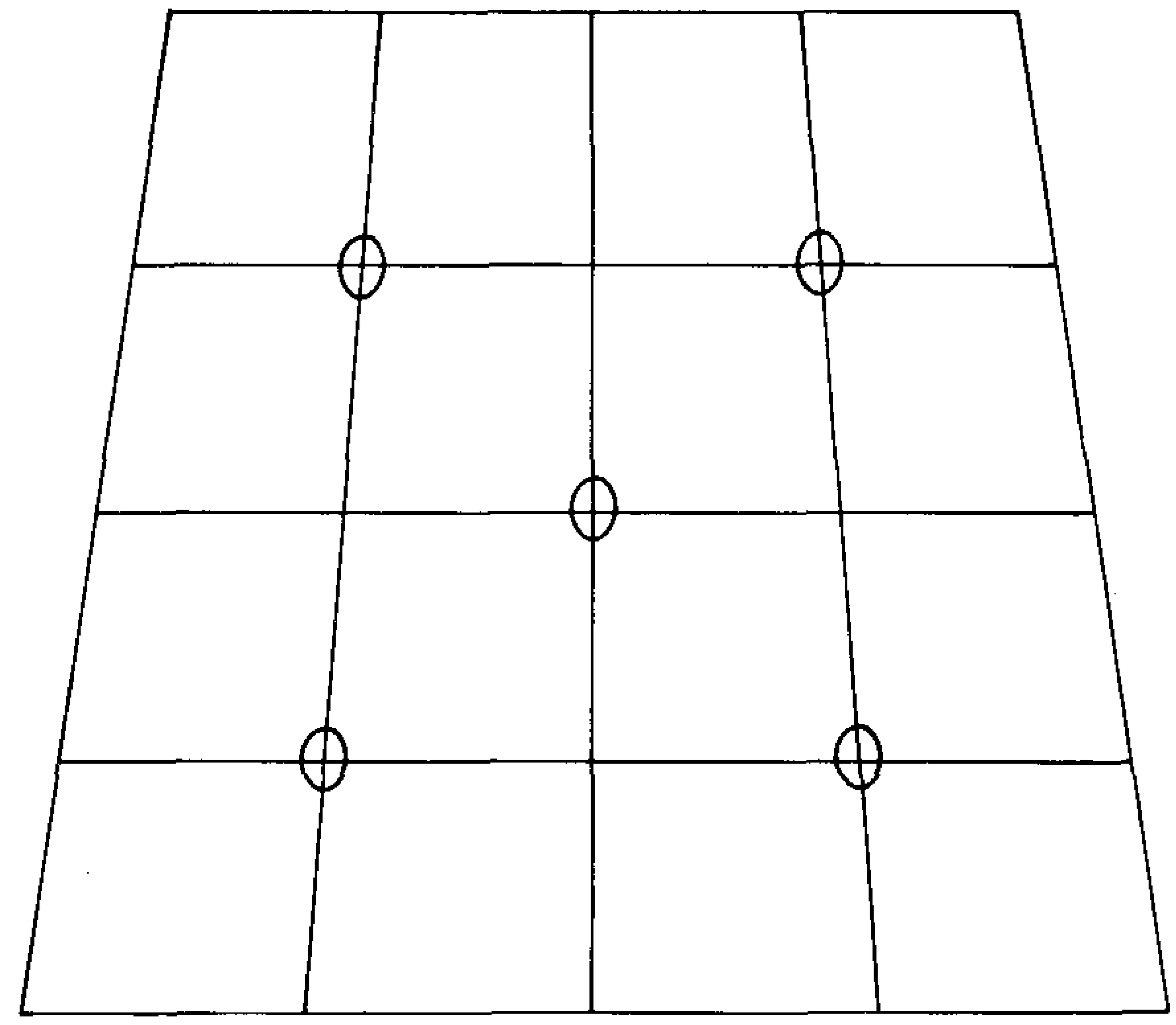

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Fig.3. Microphone placement

As mentioned earlier, the five readings from each seating area should be averaged. To do this correctly, the technique described in ISO R-140 must be used to get the true sound pressure level in the room. If the five readings for a given band differ by only a few dB then a simple average is adequate. If the readings differ by 10 dB or more, then simple averaging will introduce an error.

The method for averaging described in ISO R-140 with its conversion to microbars, squaring and root extraction would never be popular in the service field. Since we were not interested in the absolute microbar values, we found it more relevant to work out a Sound Pressure Level (S.P.L.) averaging table. All that is required is to add up the five figures and divide by five in accordance with the conversion table (see Table 1).

In most cases the spread in response among the five microphone positions will be under 5 dB, and simple averaging is adequate. Fig.4 is a histogram of 10 theaters. The first bar at 63 Hz, for example, indicates that for this octave band, one microphone position gave a reading 8 dB higher than the lowest reading in the same position. If the spread approaches 10 dB, then one might be suspicious of the room acoustics or the loudspeaker distribution.

One of the recommendations in the NFTU-1069 Document is the use of an octave filter connected to the precision sound-level meter. As each octave noise band on the test film is reproduced, the filter is switched manually to that octave. This practice is very effective in eliminating traffic rumble and

dB	"n"	dB	"n"	dB	"n"
40	.01	60	1.0	80	100
½	.01122	½	1.122	½	112.2
41	.01259	61	1.259	81	125.9
½	.01413	½	1.413	½	141.3
42	.01585	62	1.585	82	158.5
½	.01778	½	1.778	½	177.8
43	.01995	63	1.995	83	199.5
½	.02239	½	2.239	½	223.9
44	.02512	64	2.512	84	251.2
½	.02818	½	2.818	½	281.8
45	.03162	65	3.162	85	316.2
½	.03548	½	3.548	½	354.8
46	.03981	66	3.981	86	398.1
½	.04467	½	4.467	½	446.7
47	.05012	67	5.012	87	501.2
½	.05623	½	5.623	½	562.3
48	.06310	68	6.310	88	631.0
½	.07079	½	7.079	½	707.9
49	.07943	69	7.943	89	794.3
½	.08912	½	8.912	½	891.2
50	.1000	70	10.00	90	1000.0
½	.1122	½	11.22	½	1122
51	.1259	71	12.59	91	1259
½	.1413	½	14.13	½	1413
52	.1585	72	15.85	92	1585
½	.1778	½	17.78	½	1778
53	.1995	73	19.95	93	1995
½	.2239	½	22.39	½	2239
54	.2512	74	25.12	94	2512
½	.2818	½	28.18	½	2818
55	.3162	75	31.62	95	3162
½	.3548	½	35.48	½	3548
56	.3981	76	39.81	96	3981
½	.4467	½	44.67	½	4467
57	.5012	77	50.12	97	5012
½	.5623	½	56.23	½	5623
58	.6310	78	63.10	98	6310
½	.7079	½	70.79	½	7079
59	.7943	79	79.43	99	7943
½	.8912	½	89.12	½	8912
60	1.0000	80	100.00	100	10000

Table 1. *S.P.L. Averaging Table. Find the figure in the "n" column corresponding to the measuring point's dB value. Add the chosen "n" values together and divide the sum by the number of measuring points. The quotient will now give the correct S.P.L. in nearest 1/2-dB step, reading from the "n" column to the "dB" column.*

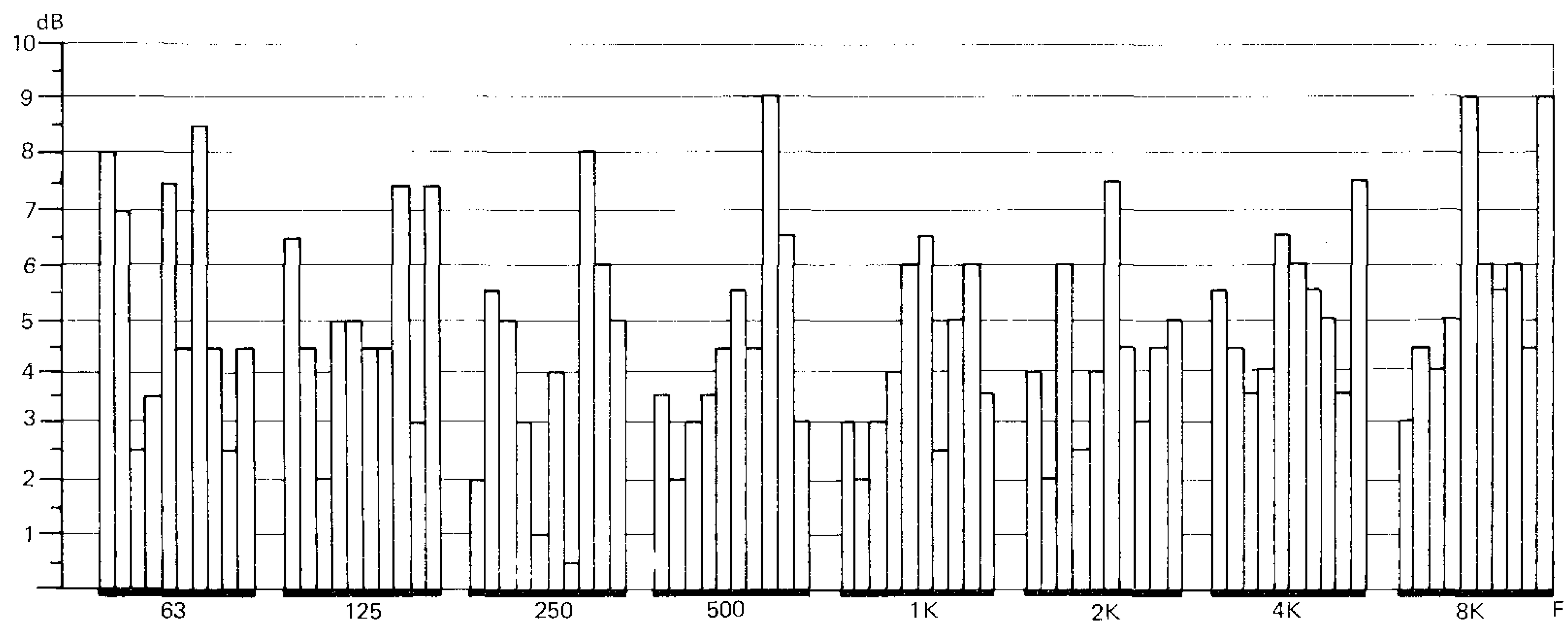


Fig.4. Histogram of ten theaters

other noise which can be as high as 60 dB when the sound-level meter is in the linear position. It is especially important when reading the 8-kHz band since it is at such a low level. Only an abnormally high volume level could override such extraneous noises but it risks the possibility of overloading the amplifier.

It is possible to read the sound-level meter visually, but in our work we have used a high-speed level recorder in every case. We wanted a reliable readout, partly for the statistical reasons and partly because we wanted to deliver a "fool-proof" report to the theater. The strip-chart recording is not subject to human errors of reading a meter and writing down numbers. At low frequencies, the needle of the sound-level meter swings over a wide range and a visual average is not very accurate unless one uses a separate meter with a very long time constant. The strip-chart recording makes possible an accuracy of 1/2 dB without difficulty.

The test film bands have a duration of 10 s, separated by a quiet interval of 5 s. The quiet space between bands is very useful in determining that we are recording the desired signal and not some parasitic noise.

A final remark in the Document 1069 is a comment about not having people cluster around the microphone. Of course this advice is in accordance with general instructions provided with sound-level meters.

When our first two-year trial period expired in March 1970, the evidence was so convincing that our executive committee granted another two-year period to continue this work. In the meantime, the background for our work had been published in professional periodicals circulating in the indus-

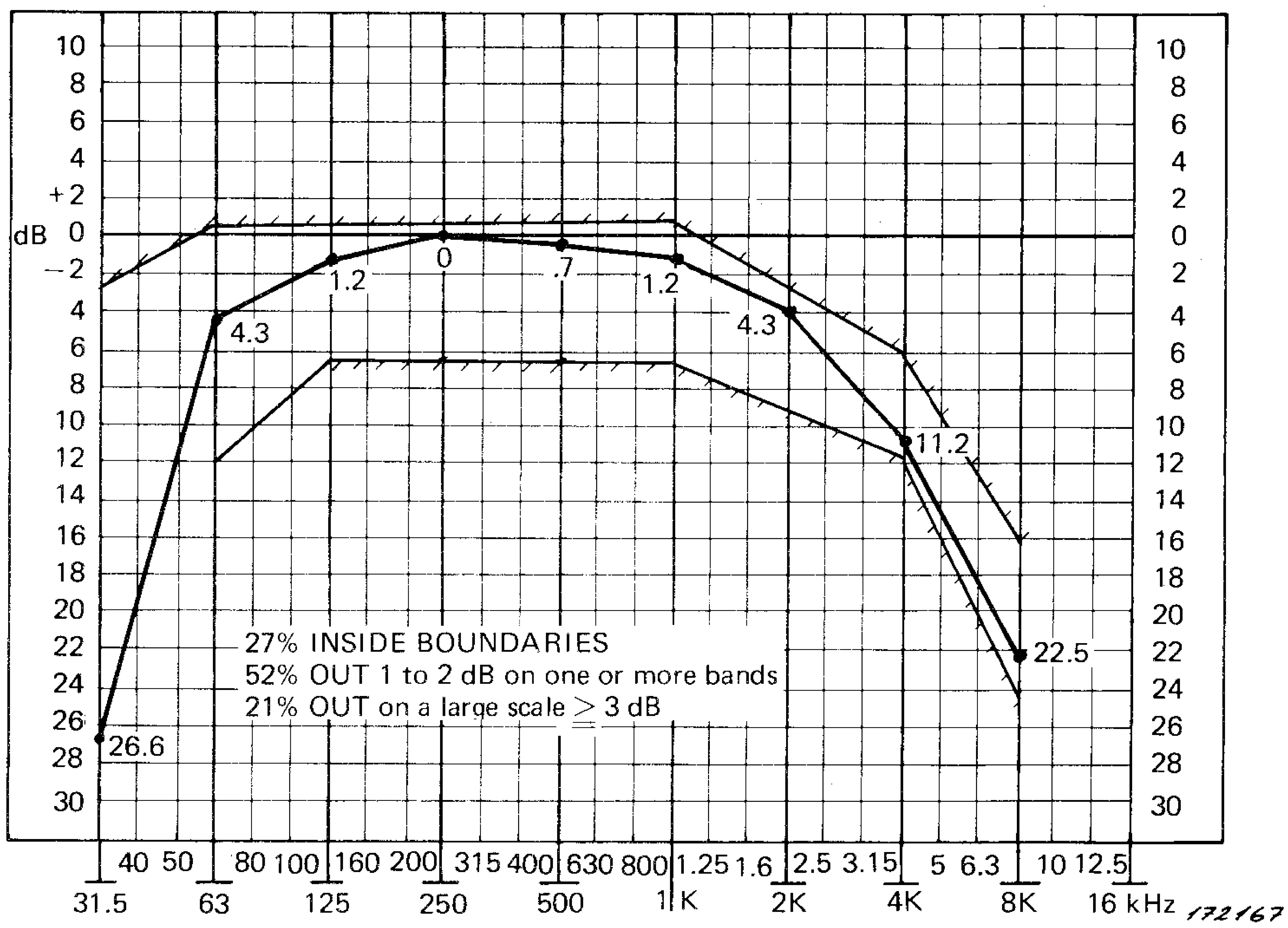


Fig.5. Arithmetic average of 100 theaters

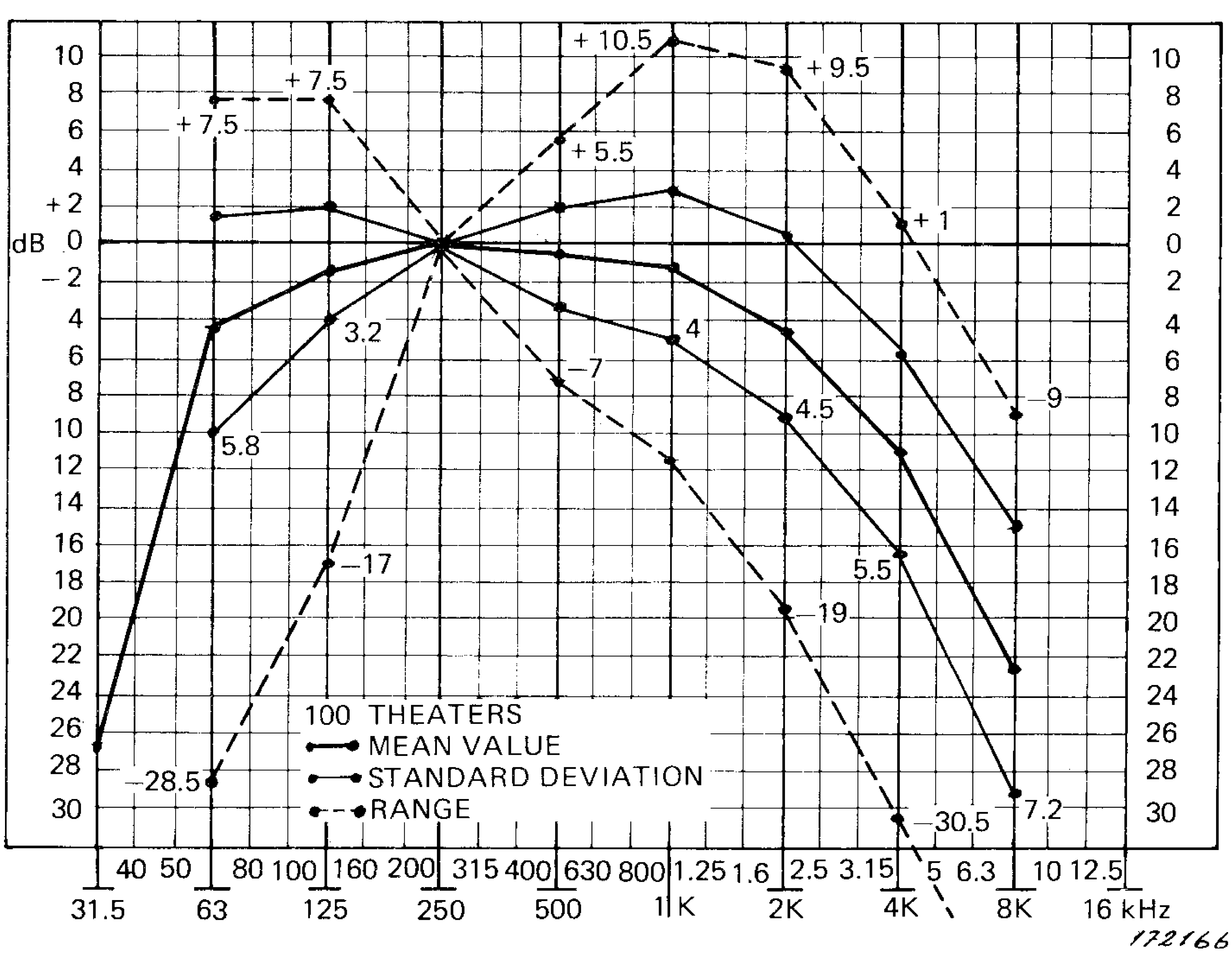


Fig.6. Mean standard deviation

try. Since our work had become known we decided to forward an application form to every cinema owner in Denmark, offering a sound test in his theater free of charge. This free offer increased the interest enormously, and we now have measured more than 100 public theaters.

Fig.5 shows the mean curve of the final listening characteristic of our 100 theaters. We have found that when a large number of theaters is averaged, the maximum response is always the 250-Hz band. Using a template to indicate the upper and lower tolerance limits, our 100 theaters ranged as follows:

27% were within tolerance;
52% exceeded the tolerance limit by as much as 1 to 2 dB; and
21% exceeded the tolerance limits by as much as 3 dB or more.

The average is just inside the tolerance, being a little heavy at the low end and somewhat deficient at the treble end. It should be emphasized strongly that these figures give the recorded situation and do not include any modifications for improving the frequency response given from our side.

The range of variation between theaters is shown in Fig.6. The heavy line is the average response. The lighter solid lines are the standard deviations, while the dashed lines indicate the extreme variations. It should be noted that the curves are all referred to zero at 250 Hz (rather than the usual 1000 Hz) as a result of a suggestion by Ljungberg. His choice of 250 Hz is based on the argument that below 250 Hz loudspeaker radiation is essentially omnidirectional, and that above 250 Hz loudspeaker directional patterns become a significant influence. We chose 250 Hz for another good reason. We do not adjust our subjective listening level according to the 1000-Hz output but rather to the 250-Hz octave level. It is not important which frequency is chosen as a zero reference, but the choice of 250-Hz provides a much more relevant picture of the enormous spread we experience in the treble bands.

Two illustrations will show how bad the situation can get. Neither of these was included in our statistics.

Fig.7 is a graph pertaining to a small preview room with 29 seats. The broken line indicates an electrical response that looks quite normal, and considered alone it would not be cause for suspicion. The acoustic response, however, shows an extremely high peak at 500 Hz and a steep drop-off at both the high and low ends of the spectrum.

Fig.8 shows both the electrical and the averaged acoustical response of a

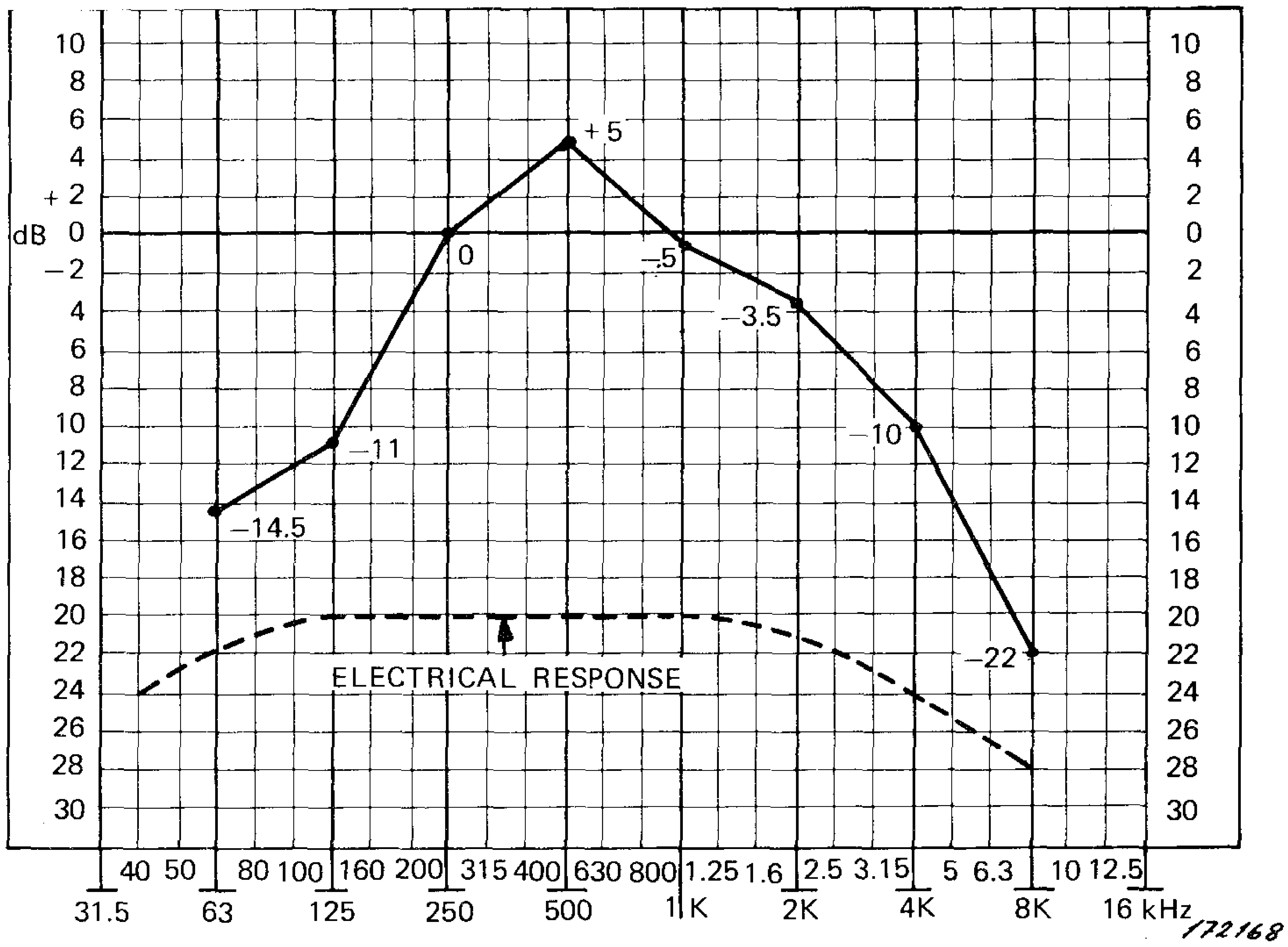


Fig.7. Studio preview room with 29 seats

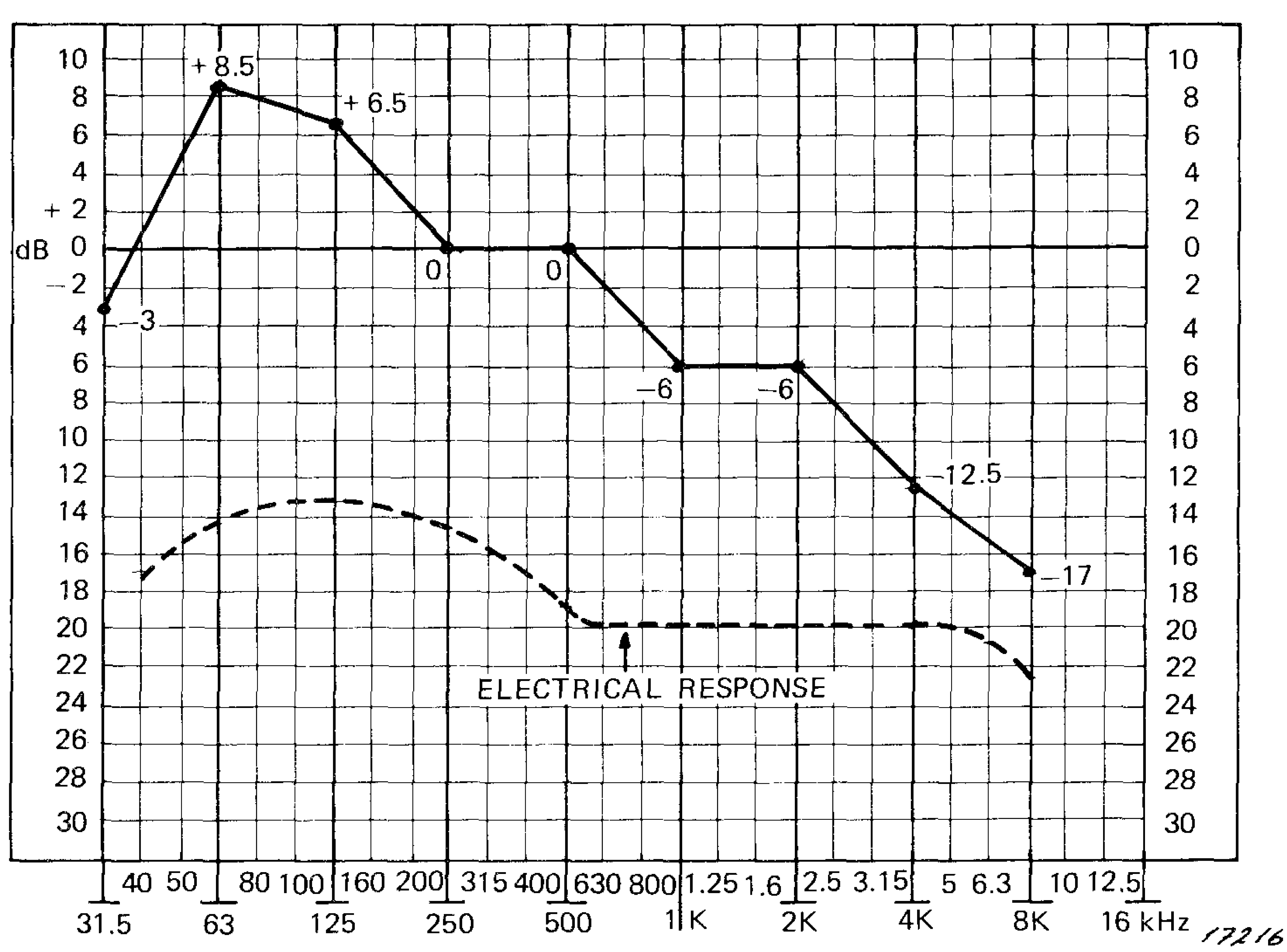


Fig.8. Theater with 750 seats

750-seat cinema. The loudspeakers appear to be unusually efficient in the 63- and 125-Hz bands, when compared with the 1000-Hz level. The electrical response had been raised about 7 dB at 125 Hz for no apparent reason aggravating an already bad situation.

In our cinema measurement program, we wanted to minimize the time and labor for making measurements and therefore used nine bands of octave noise. We also considered the use of one-third octaves. However, as Ljungberg pointed out: "Too much detailed information can turn out to be a refinement in the wrong direction". Since the number of microphone positions needed for third-octave measurements should be three times as many as for octave measurements and since there are three times as many noise bands, the time consumed increases by a factor of nine.

Fig.9 shows a one-position acoustic response of a mixing room obtained by using one-third octave noise. Keeping all measurement conditions unchanged, the response was measured again with octave bands noise which produced the curve shown in Fig.10. The octave measurement agrees well with subjective listening impressions, and any desired equalization could be readily achieved with simple R-C networks. The third-octave measurements are confusing and equalization would be incalculable.

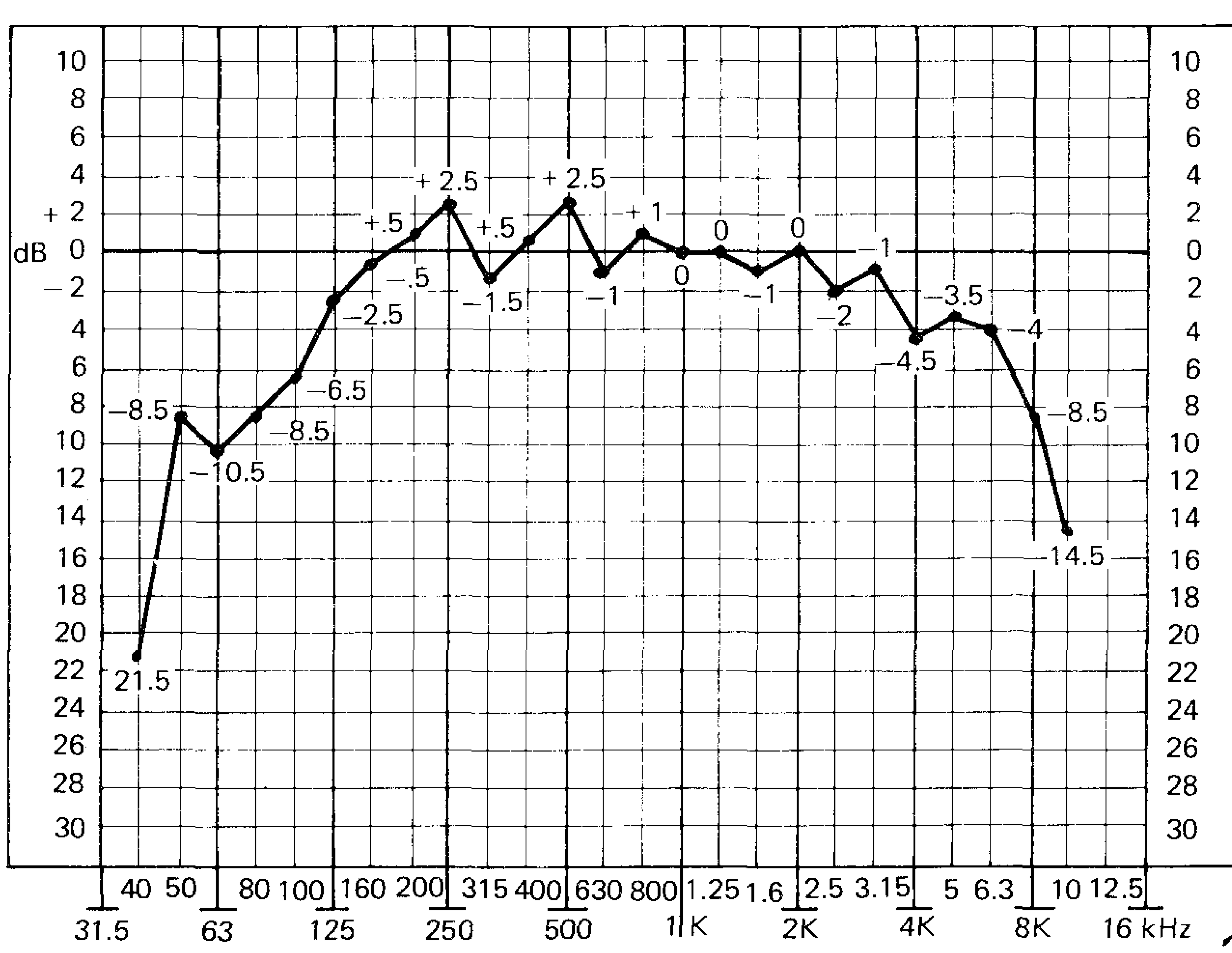


Fig.9. One-third octave pink noise

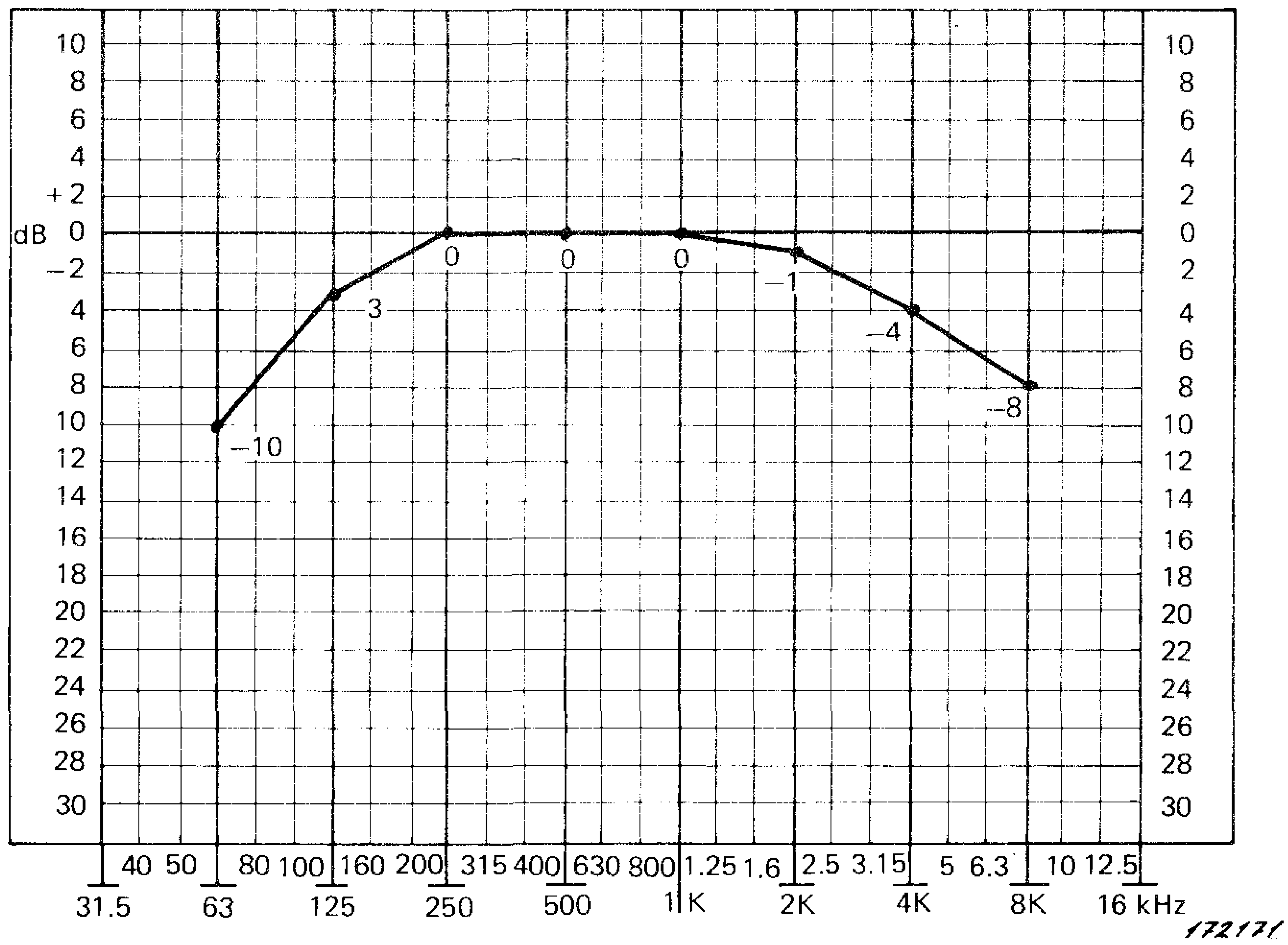


Fig. 10. Octave pink noise

In certain difficult cases where a more thorough investigation might be needed, octave noise shifted in one-third octave steps may be a solution. Petro Vlahos of the USA has obtained special test films to investigate this technique.

The work we have accomplished during the last three years has shown us that we cannot close our eyes (or ears) to what happens after the signal is delivered to the loudspeaker terminals. The old prevalance of stopping all measurements at that point is insufficient. The conversion of the signal to sound pressure waves by the loudspeaker and the room should not be left to psychological factors alone. At a recent symposium of professional engineers from all Scandinavian countries it was decided that measurements and listening tests must go hand in hand; the one cannot do without the other. We must not overlook the evolution in acoustic measurement instrumentation, which has made available very dependable equipment. A great deal of uncertainty will be avoided if an agreement is reached on a standard acoustic frequency response.

References

1. Erik Rasmussen — "A report on listening characteristics in 25 Danish cinemas," Jour. SMPTE, 78: 1054–1057, Dec. 1969.

2. Lennert Ljungberg "Standardized sound reproduction in cinemas and control rooms," Jour. SMPTE, 78: 1046–1053, Dec. 1969.

3. Erik Rasmussen "Production of an optical 35 mm test film with narrow-band filtered random noise," presented at the 8th Congress of UNIATEC, Brussels, September 23–27, 1968.

Acoustic Response Standardization is at present being discussed by ISO TC-36, WG-3.

Erik Rasmussen

Noise Dose Measurements

by

Leif S. Christensen

ABSTRACT

An outline is given of the relevant parameters in the single-number description of acoustic noise climate contained in several recent standards and recommendations. A special instrument, Type 4423 Noise Dose Meter has been developed for this kind of measurement, its principle of operation is explained, and examples of practical measurements using it are given, along with examples of more conventional equipment.

SOMMAIRE

L'article souligne les différentes composantes des descriptions de l'environnement acoustique contenues dans plusieurs normes et recommandations récentes. Un instrument spécial, le Dosimètre de Bruit Type 4423, a été construit pour ce genre de mesures; on décrit son fonctionnement, et des exemples de mesures pratiques l'utilisant sont donnés, en même temps que des exemples faisant appel à un équipement plus conventionnel.

ZUSAMMENFASSUNG

Es wird ein Überblick über die Parameter gegeben, die für die Beschreibung des akustischen Lärmklimas durch eine einzige Größe von Bedeutung sind, wie es in verschiedenen neuen Normen und Empfehlungen festgelegt ist. Für diese Art von Messungen wurde ein spezielles Gerät entwickelt, das Lärmdosimeter 4423, dessen Arbeitsprinzip erklärt und dessen praktischer Einsatz von Beispielen dargestellt und mit konventioneller Ausrüstung verglichen wird.

Introduction

Problems caused by noise pollution are more prevalent in modern industrial society because noise levels are increasing and because of the focus on environmental problems in general. Therefore, a need is felt for simple principles and equipment for evaluation of noise and its effects.

Noise may cause hearing loss or may be harmful through its less direct effects such as interference with sleep, conversation and work or leisure activities. In both cases the effect is a question mainly of noise intensity and duration. Further, the frequency content of noise is of great importance; a variety of weighting principles have been suggested, but none have been able to cover all applications. However, the "A" filter has gained extensive use since it is simple and yet provides a fair approximation to human response, from the point of view of annoyance as well as hearing loss.

As a preliminary illustration of the components of a noise dose measurement, consider Fig. 1. The simplest measuring problem is one involving

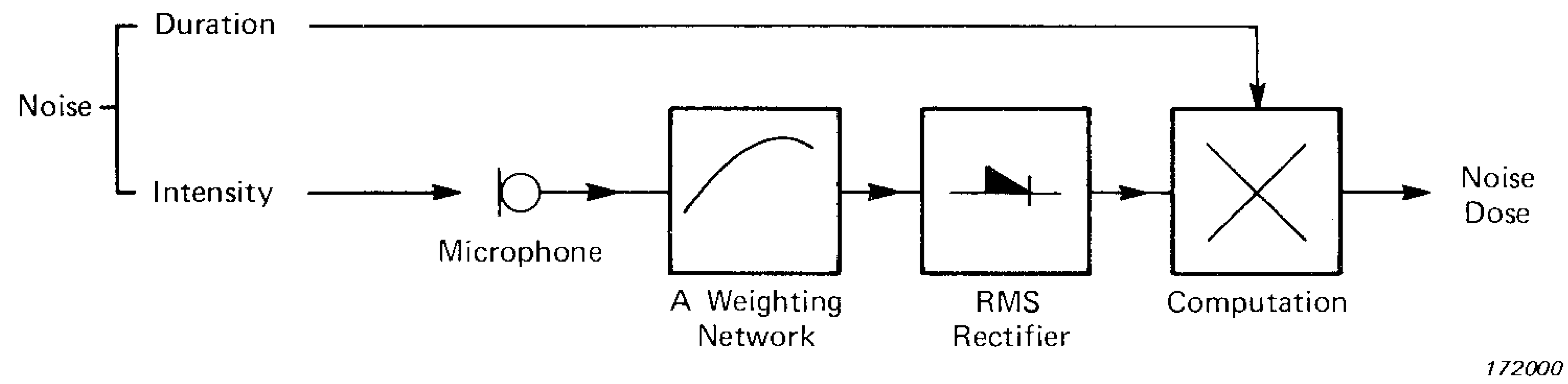


Fig. 1. Components of a noise dose measurement

steady, continuous noise such as may be measured by a constant deflection on a sound level meter. In this case the noise dose may be specified as being the constant sound level for the time duration. If the sound level varies with time — which it does as a rule — some sort of "trade — off" of intensity and duration must be applied, as indicated by the block "computation" in Fig. 1. Such a trading relationship would tell us, for example, how much longer 85 dBA can be tolerated than 88 dBA to give the same annoyance or damage.

Unfortunately no final relationship of this kind has emerged, and probably never will as both noise and people come in a very wide variety. The economic aspects of noise criteria also tend to complicate the issue.

Standards for noise dose measurements

Despite the above mentioned difficulties the need for noise dose measurements has led to several recent standards and recommendations. Among these are:

1. ISO Recommendation R 1999: Assesment of Occupational Noise Exposure for Hearing Conservation Purposes.
2. ISO Recommendation R 1996: Assessment of Noise with Respect to Community Response.
3. Walsh-Healey Act (USA federal law concerning hearing conservation in government supplying industry).
4. DIN 45 641 (BRD proposal concerning noise dose measurements).

The trading relationships recommended in these standards are shown in Figs.2 and 3. To allow comparison the criterion of a maximum allowable level of 90 dBA for an 8 hour working day has been taken. It can be seen that the trading relationships in this representation are in the form of straight lines. The slope of these straight lines is given the symbol "q".

ISO and DIN operate with the parameter q = 3, i.e. 3 dBA increase of sound level corresponds to half the duration. (DIN does allow for alternative q's).

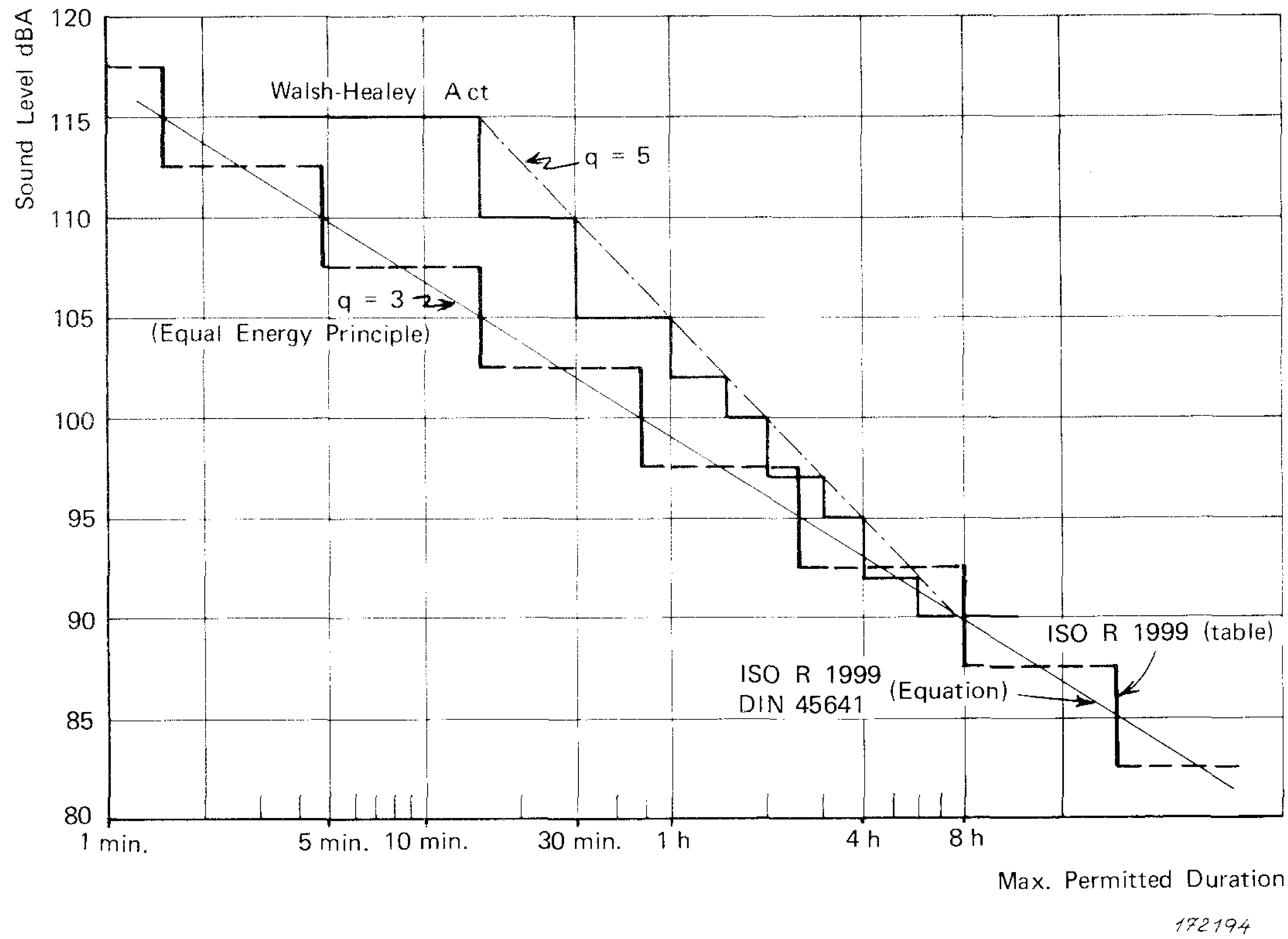


Fig.2. The relationship between sound level and duration for the Walsh-Healey Act, ISO R 1999 and the DIN 45 641 proposal, (For ISO and DIN applying the criterion 90 dBA x 8 h per workday and no corrections)

This corresponds to constant sound energy for a given dose (the product of intensity and duration being constant), hence the designation Equal Energy Principle. On the other hand, the Walsh-Healey Act has q = 5 (5 dBA increase for halving of duration). For the criterion of 90 dBA for an 8 hour working day, the q = 5 curve is seen to be more permissive at higher levels than the q = 3 curve. Simple averaging would correspond to q = 6. Once the numerical value of the parameter q is specified it is possible to convert the varying sound level p(t) within the considered time period T to an equivalent continuous sound level L_{eq} according to the equation.

$$L_{eq} = \frac{q}{\log_{10} 2} \log_{10} \left(\frac{1}{T} \int_0^T \left(\frac{p(t)}{p_o} \right)^{\frac{20 \log_{10} 2}{q}} dt \right) \qquad (1)$$

where p_o represents the reference sound pressure 2×10^{-5} N/m^2. The conversion may be performed on the basis of level classification — as shown

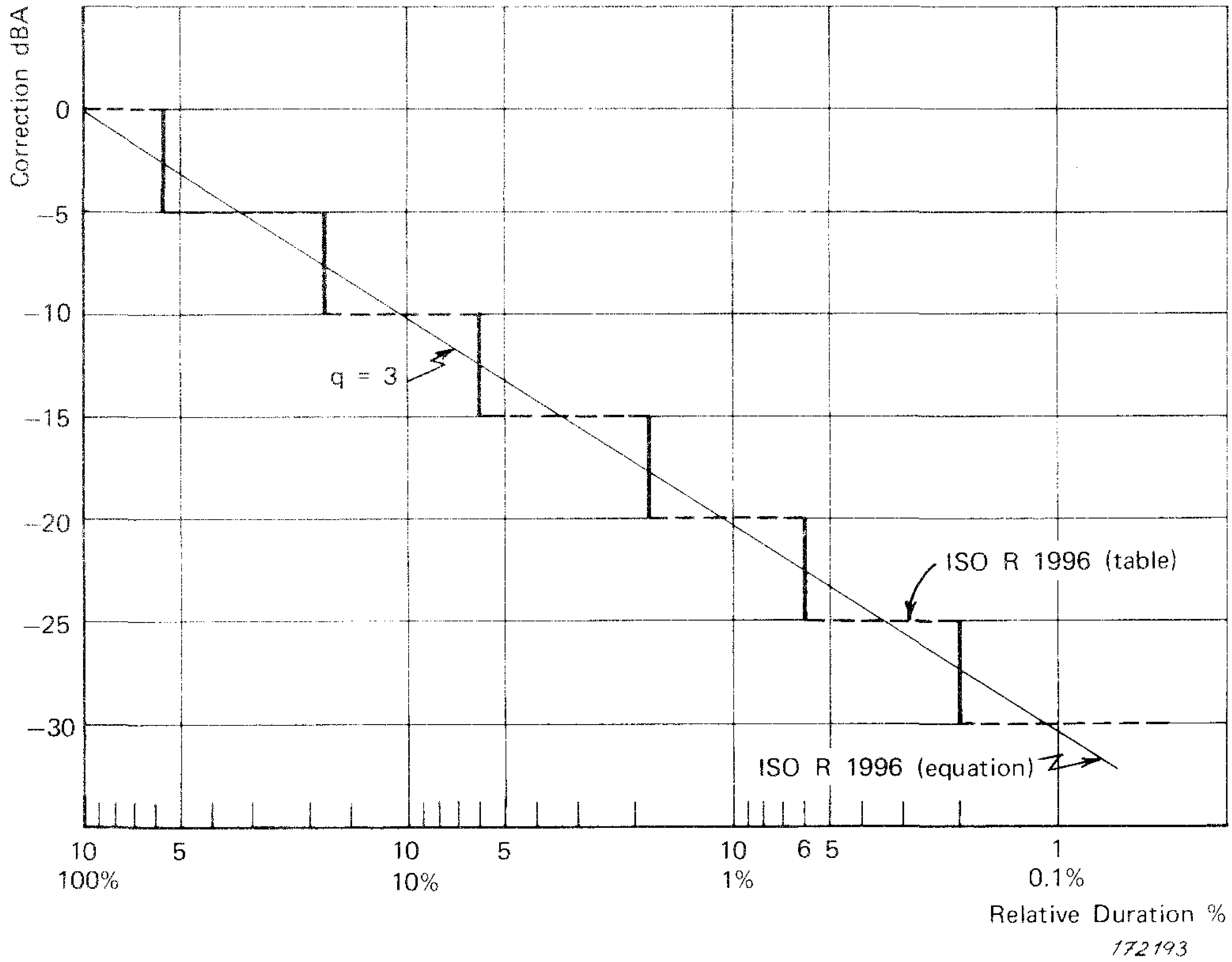


Fig.3. The relationship between sound level and relative duration for ISO R 1996

by the staircase curves of Figs.2 and 3 or by integration according to the above equation (which is the way the Noise Dose Meter Type 4423 works).

Tones and impulsive noise

While the technique for measurement and evaluation of continuous and intermittent noise is fairly well established, tones and noises of an impulsive nature still present a major problem. ISO R 1999 recommends the addition of 10 dBA for noise such as that from hammering and riveting, while ISO R 1996 prescribes the addition of 5 dBA for such noise and also for audible tone components. The DIN proposal attempts to provide for a more accurate assessment of "impulsiveness" by specifying the use of equipment with "Impulse" time constant circuitry.

Type 4423 Noise Dose Meter

Standards such as those mentioned above along with to-day's semiconductor technology permit the design of compact instrumentation for measurement of noise dose.

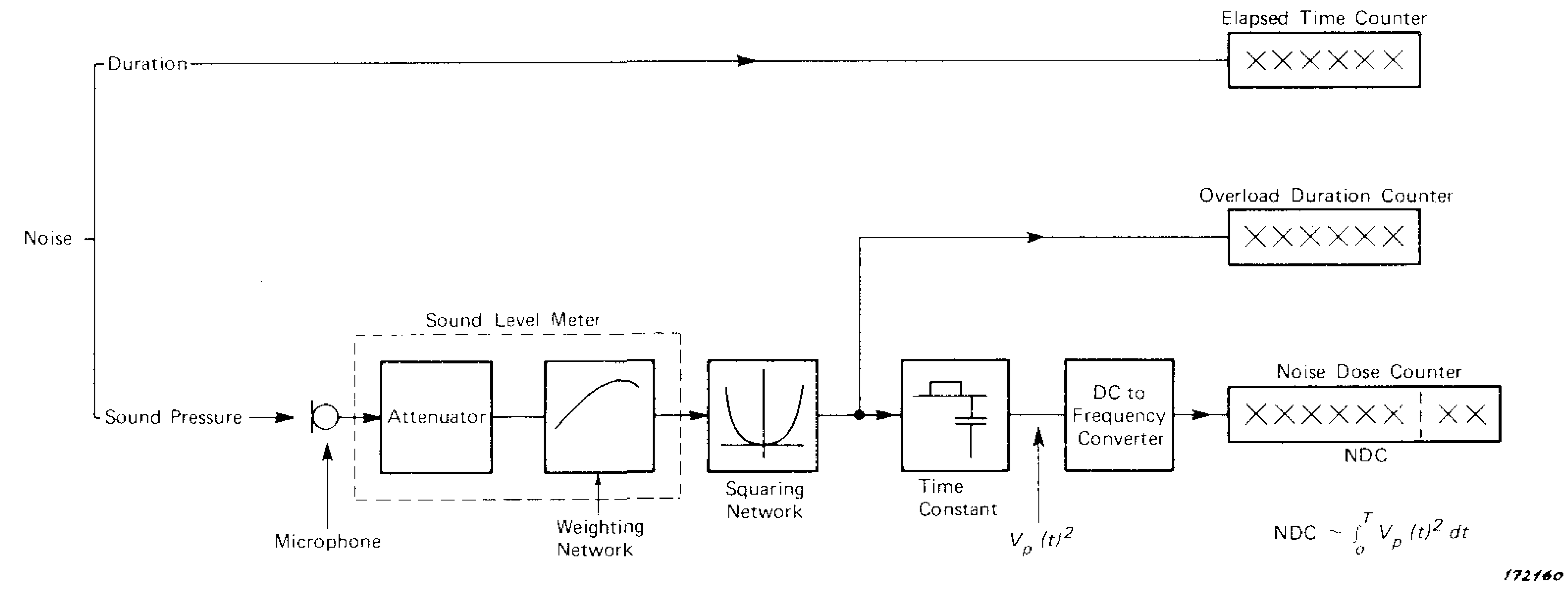


Fig.4. Functional diagram of the Noise Dose Meter, Type 4423, (Showing measurement set-up including SLM)

The Type 4423 Noise Dose Meter works according to the equal energy principle, i.e. using q = 3 in equation (1):

$$L_{eq} = 10 \log \left(\frac{1}{T} \int_0^T \left(\frac{p(t)}{p_o}\right)^2 dt\right) \text{ or}$$

$$L_{eq} = K + 10 \log \left(\frac{1}{T} \int_0^T V_p\,(t)^2\, dt\right) \qquad (2)$$

where $V_p(t)$ is the voltage representation of p(t) within the instrument, and K includes p_o along with the instrument constants. The block diagram Fig.1 may thus be modified as shown in Fig.4.

The RMS rectifier — which will normally contain a squaring network, time constant circuit and square root extractor — has been combined with the squaring function required for dose measurement, so that only one squaring network is required. The integration is performed through a voltage to frequency converter and a counter which indicates the Noise Dose Count (NDC) while the measurement duration T is recorded on a second counter. In the connected sound level meter another variable enters, the attenuator setting (Att.), and separating it from the constant K in (2) we obtain:

$$L_{eq} = \text{Att.} + K' + 10 \log \frac{NDC}{T} = \text{Att.} + X \qquad (3)$$

X may be obtained from the special slide rule (Fig.5), where X appears above the arrow when T is set above NDC on the lower scale.

Electrical overload may have an adverse effect on the measurement accuracy, so a third counter (Fig.4) is provided to record the duration of any overload occurring at the squaring circuit output, this being the most critical point in the system.

For analysis of the instantaneous sound level a DC output from the time constant circuit is provided, time constants corresponding to "Fast", "Slow", "1 ms" and "Impulse" being available. Whichever of the first three is selected will make no difference to the noise dose count (see appendix for proof). Selection of "Impulse" may, however, affect the noise dose count, a more "impulsive" type of noise giving a higher count.

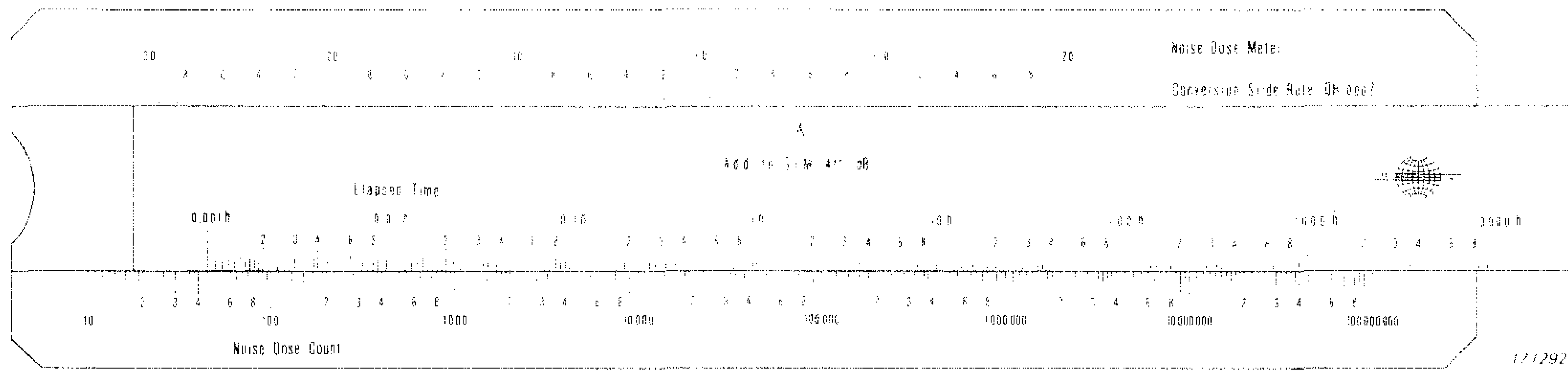


Fig.5. Conversion slide rule for the Noise Dose Meter

Examples of practical noise dose measurements
The energy equivalent continuous level (q = 3) was obtained for two tape recorded samples of noise. One recording (duration 10 minutes) was from a 4 lane road carrying about 3000 vehicles per hour at approx. 80 km/h average speed, the microphone being placed 10 m from the edge of the nearest lane. Fig.6 is a level recording of part of the sample.

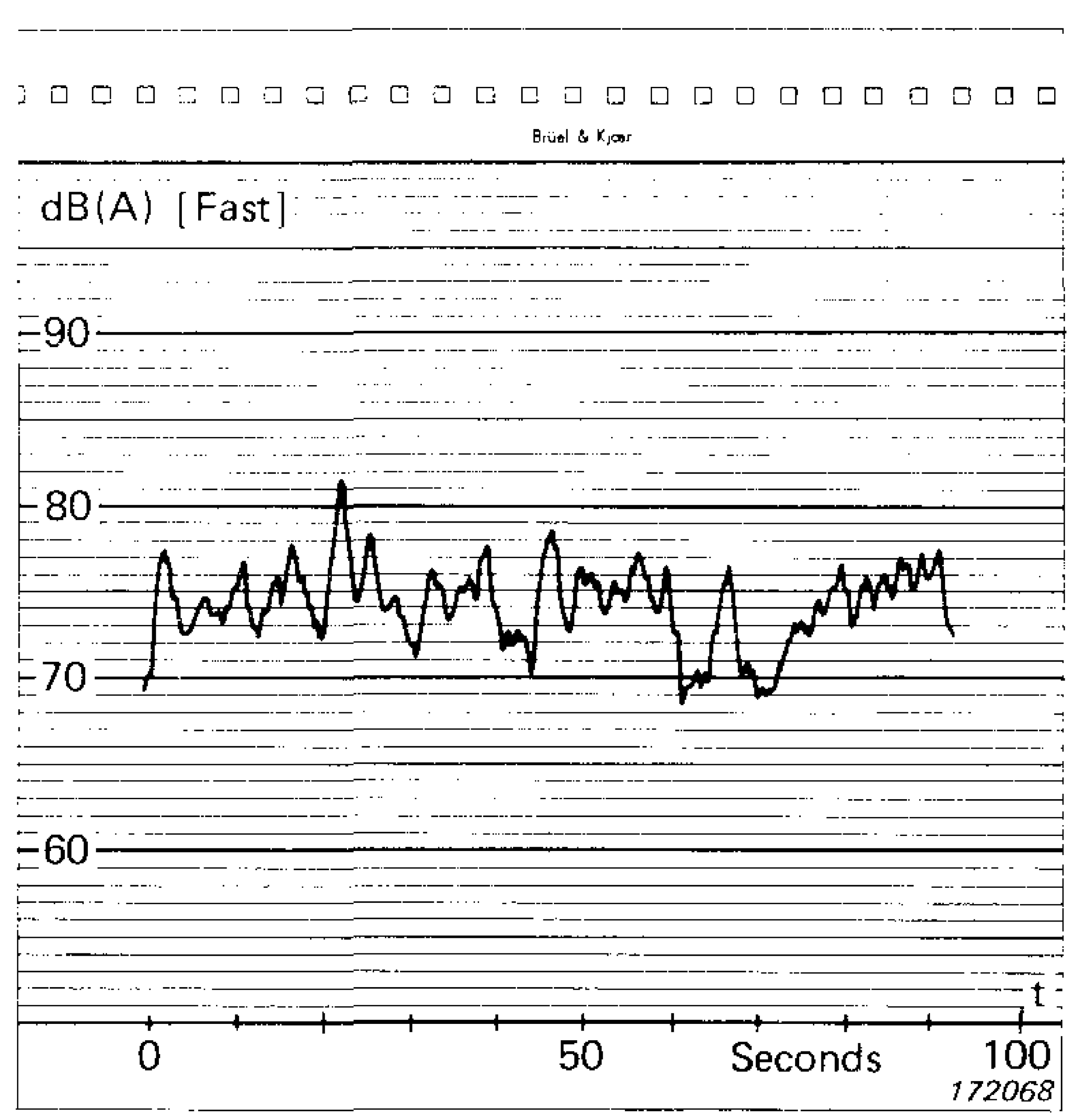


Fig.6. Traffic noise, sample of level recording

This is a rather docile type of noise as far as RMS detection goes: the opposite is true for the other recording, 12 minutes of noise from an automatic punching machine (sample shown on Fig.7) being impact noise occuring at varying time intervals (1 sec. or more) with peak values 10 – 20 times, the RMS level.

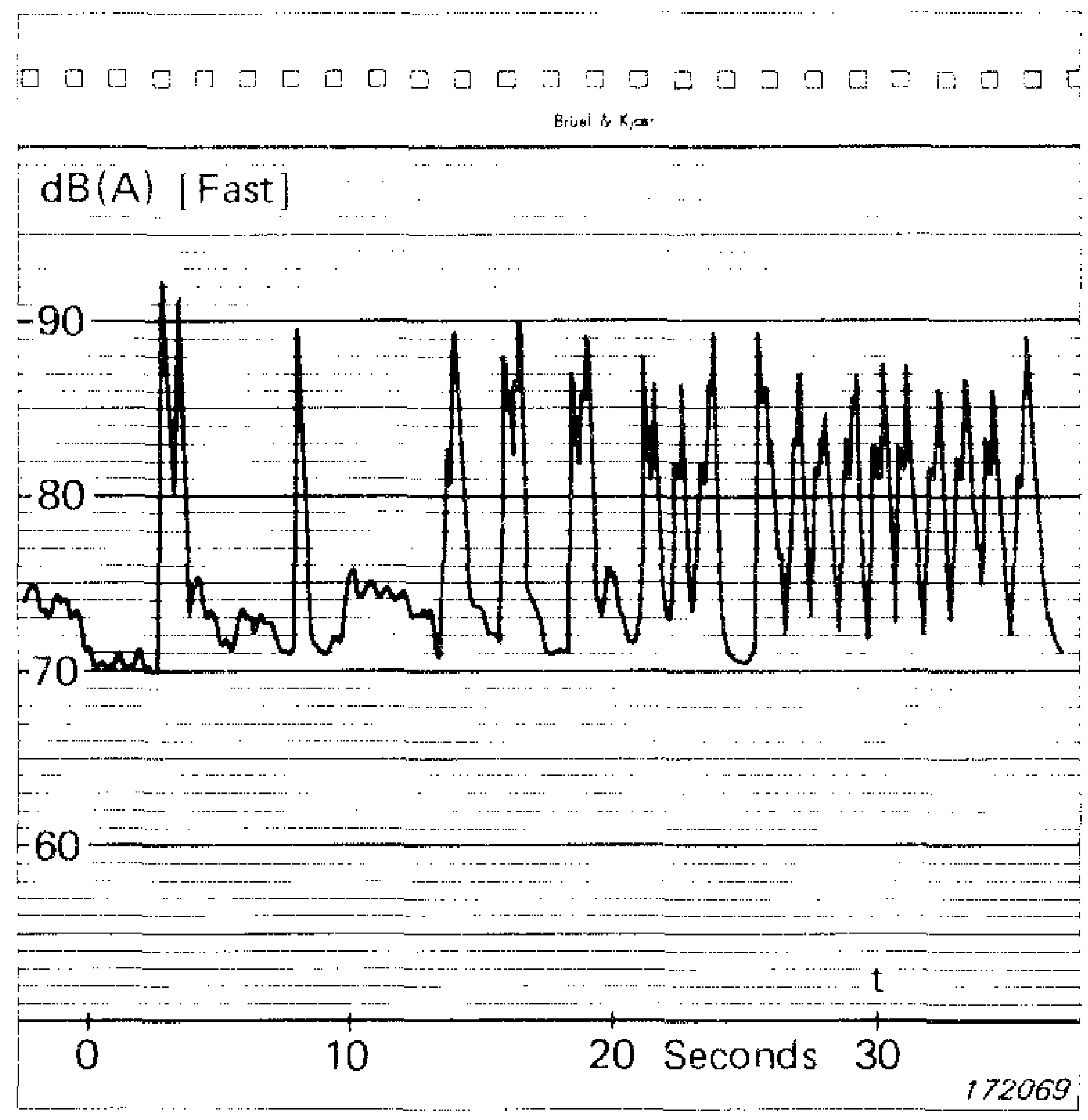


Fig.7. Impact noise, sample of level recording

Three different measuring arrangements were used. Fig.8 shows a conventional set-up using level recorder and statistical distribution analyzer.

From the temporal distribution within the analyzer's level classes L_{eq} may be calculated (ISO R 1999 has directions for such calculations). In this case the RMS rectifier of the level recorder is used, and the 2305 controls were set according to the recommended approximations to "Fast" and "Slow" in the manufacturer's instruction manual. By adding a Measuring Amplifier Type 2606 a similar arrangement may be used also for impulsive noise (Fig.9).

Here the 2606 rectifier is used, which is designed for measuring impulsive signals applying the time constants "Fast", "Slow" and "Impulse". From 2606 the DC output signal is taken for analysis in 2305/4420, providing good linearity over an app. 20 dB range.

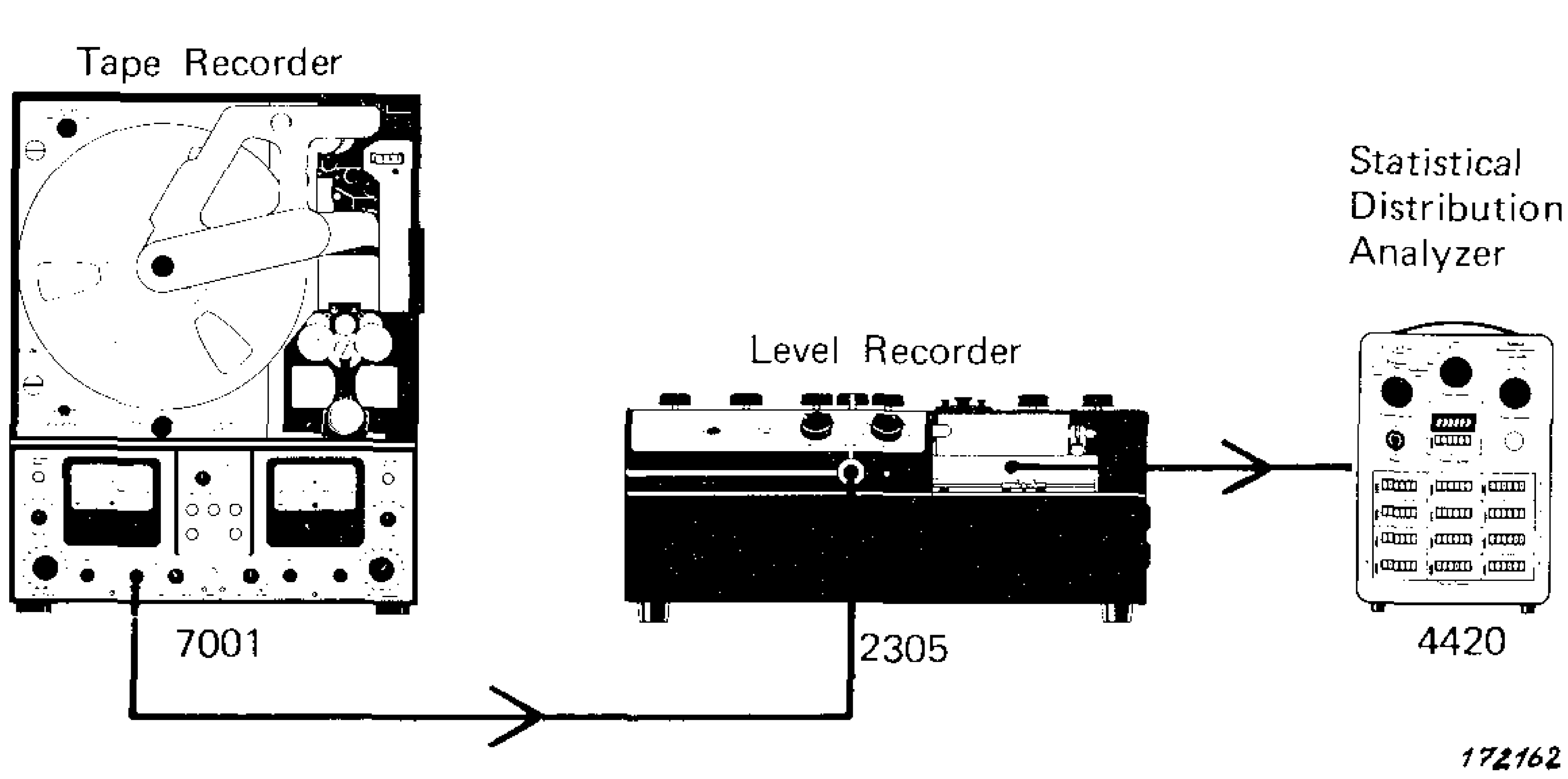


Fig.8. Conventional equipment used for traffic noise measurement

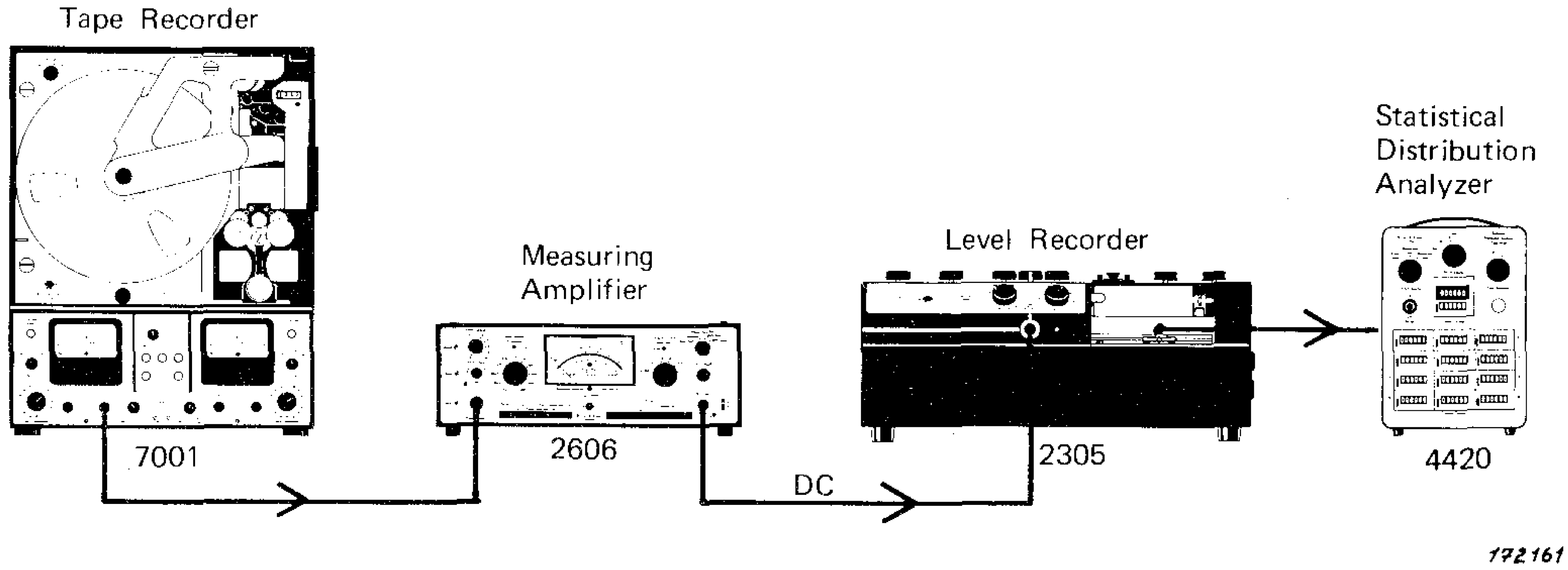


Fig.9. Conventional arrangement used for traffic noise and impact noise measurements

Finally, measurements were made using a prototype Noise Dose Meter as shown in Fig.10. With 4423, "crest factor" considerations are irrelevant since there is no piecemeal linear approximation to squaring (integrated circuitry with inherent squaring characteristics being used).

In the table below, the measurement results are summarized.

Within the calibration – and resolution accuracy (estimated at ±0.3 dB) the agreement is good between the results for the three set-ups. As expected, no appreciable difference occurs between "Fast" and "Slow" results, but "Impulse" yields an extra 1 and 7 dB resp. for the two types of noise.

As a further check on the results, gain was changed in 1 dB steps. Fig.11 and Fig.12 show the resulting change in L_{eq}. A 50 dB potentiometer was

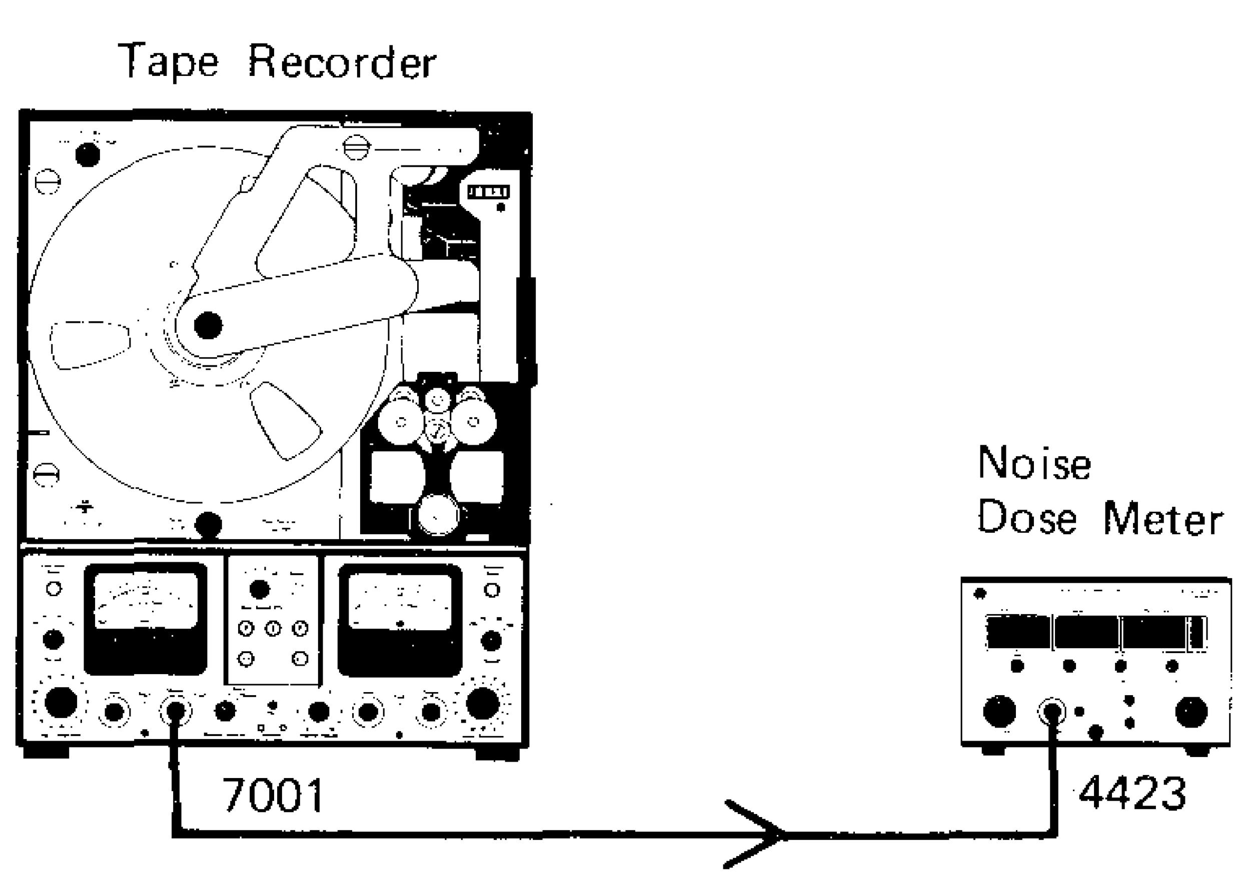


Fig.10. Noise Dose Meter, Type 4423 used for traffic noise and impact noise measurements

Measuring Equipment	2305 + 4420		2606 + 2305 + 4420			4423		
"Time Constant"	Fast	Slow	Fast	Slow	Impulse	Fast	Slow	Impulse
Traffic Noise	75,5	75,5	75,3	75,3	76,4	75,2	75,3	76,3
Impact Noise	Not applicable		82,4	82,7	89,8	82,4	82,5	89,6

Table 2. Measurement results L_{eq} dBA (non-corrected) for two types of noise

used in 2305, corresponding to 5 dB classes in 4420: A too coarse class resolution is likely to show up as periodic deviation (5 dB period) from the 1 dB/1 dB line. A minor effect of this sort may be traced in the two curves (max. excursion ±0,1 – 0,2 dB). For comparison, the 4423 results are also shown.

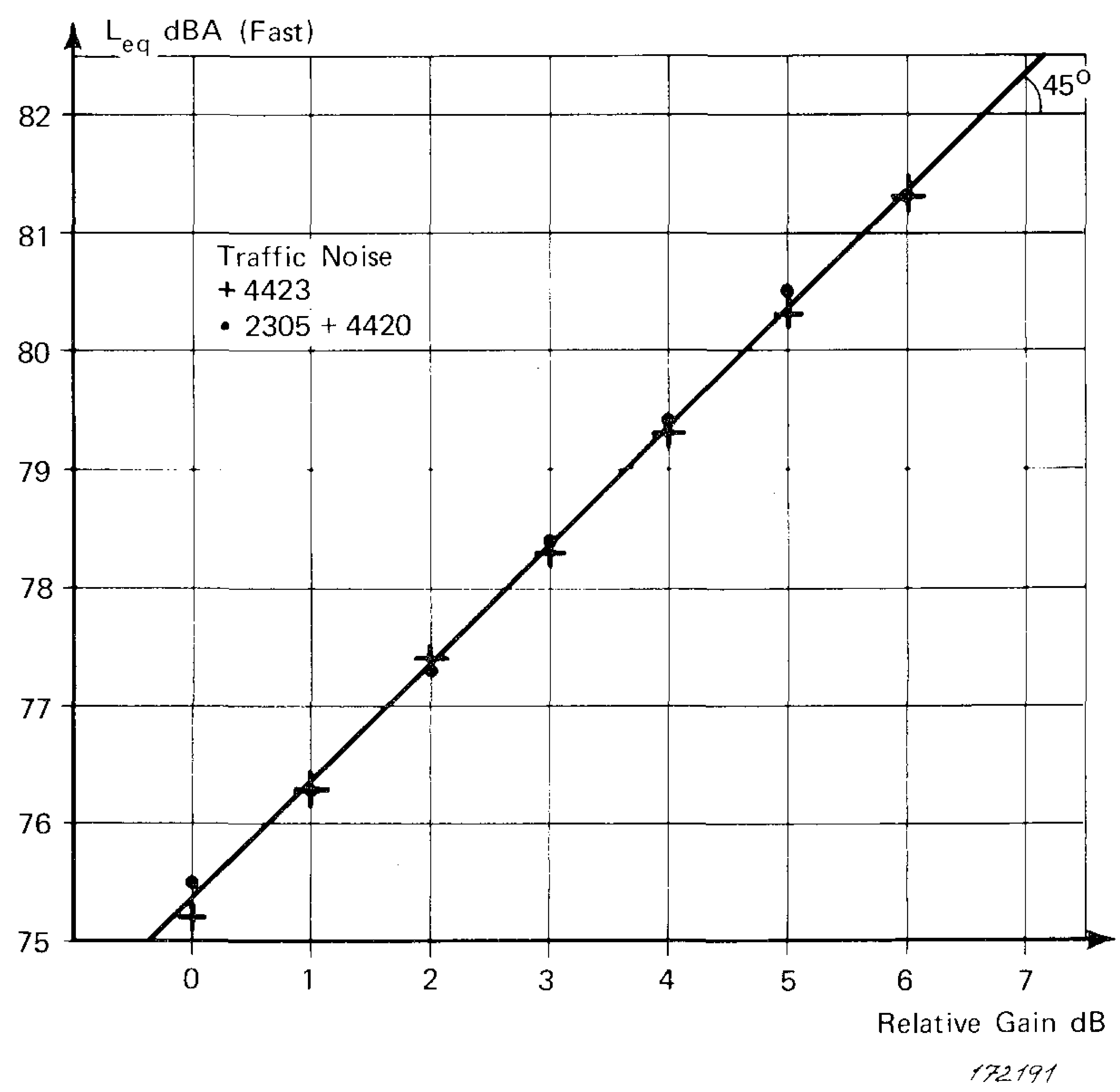


Fig. 11. Effect of change of gain, traffic noise measurements

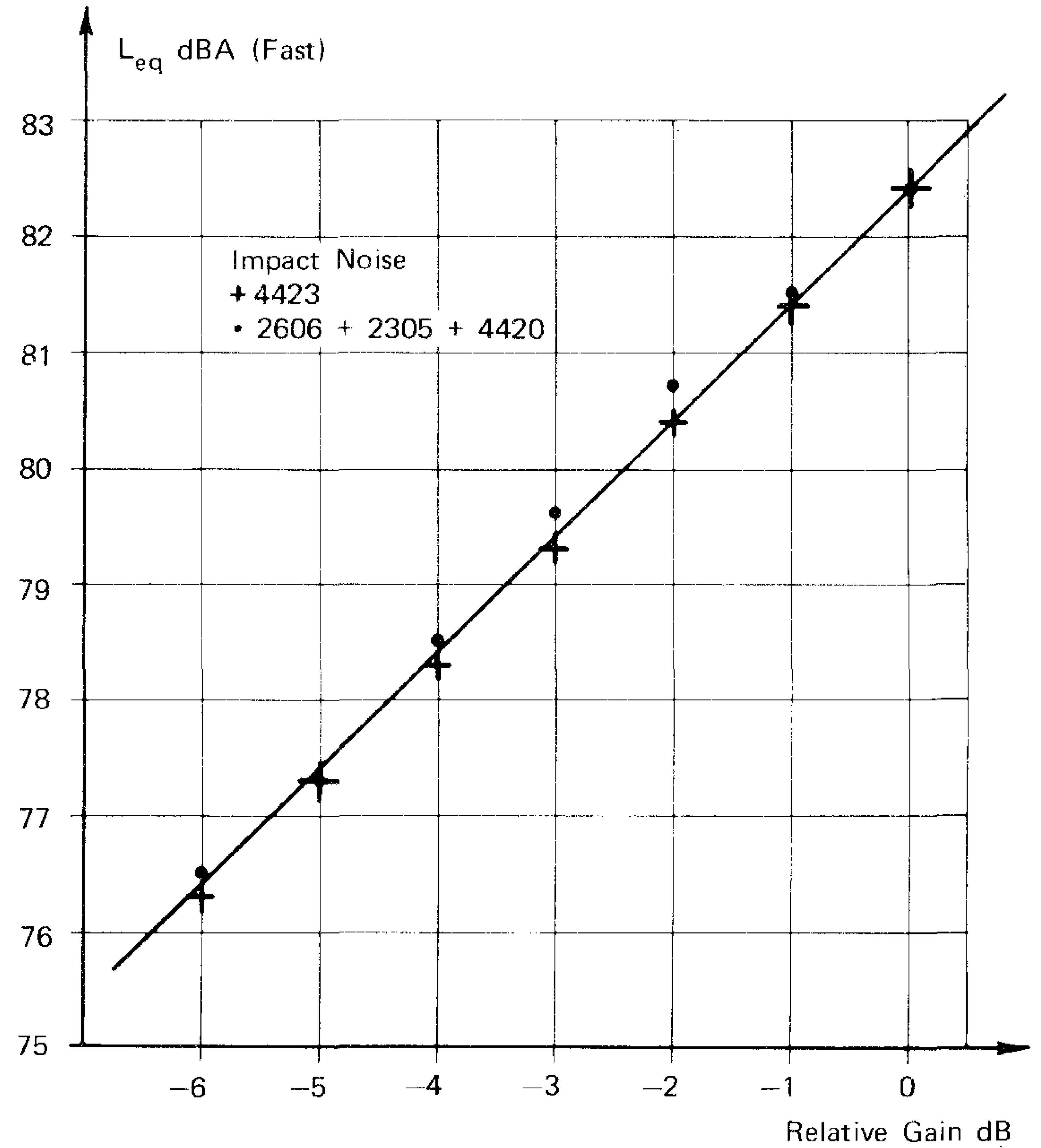


Fig. 12. Effect of change of gain, impact noise measurements

Overload indication

As mentioned previously in connection with Fig.4, system overload is recorded on the 4423 "Overload Duration" counter. Unfortunately this duration — if any — does not permit directly an estimate of the L_{eq} error result-

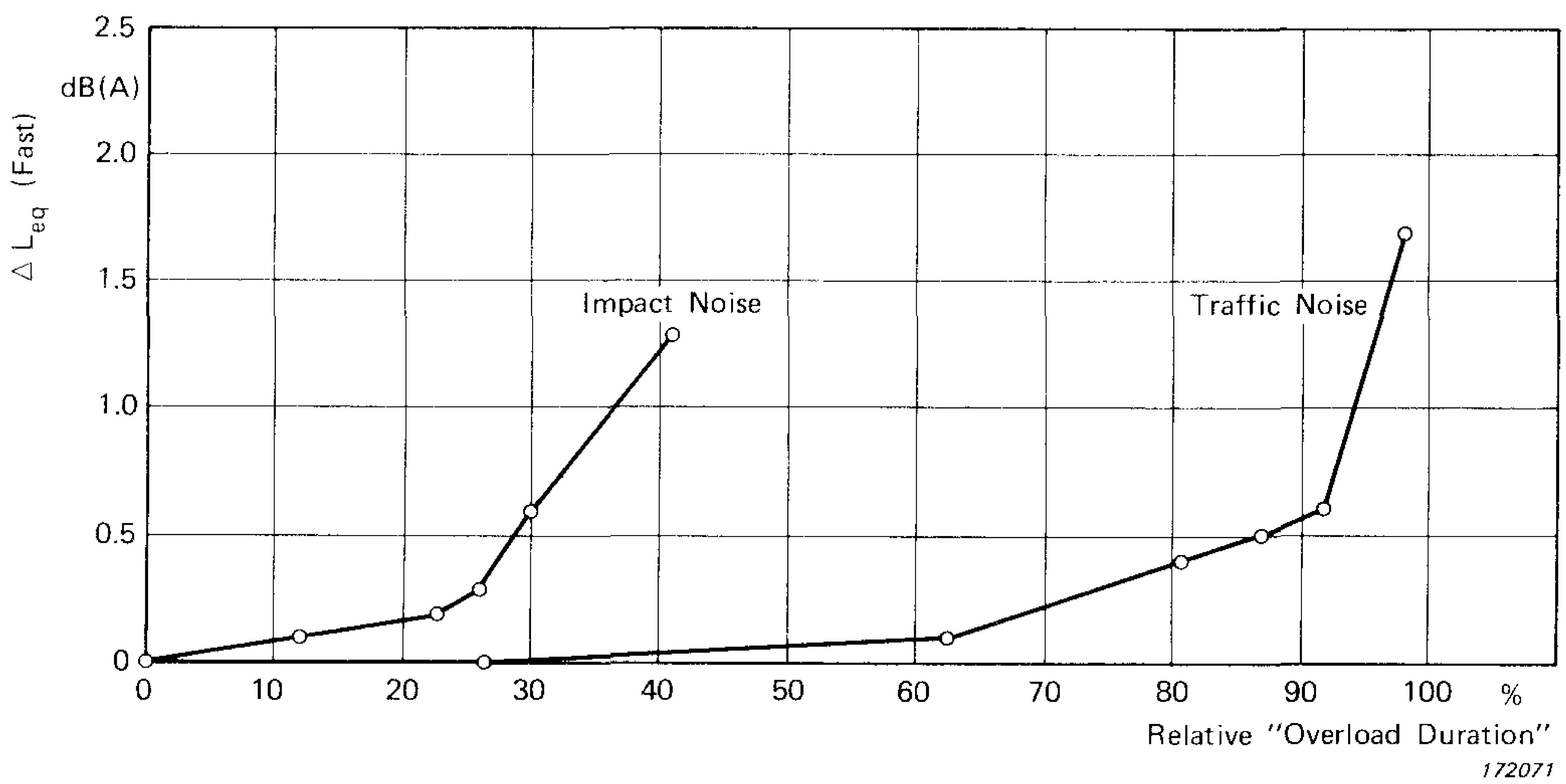


Fig. 13. L_{eq} error as a function of Type 4423 overload duration

ing from overload. This is illustrated in Fig. 13, which shows the error as a function of "Overload Duration" for the two investigated recordings of noise. For an error of about 0,5 dB, "Overload Duration" is as different as 30% and 90% in the two cases.

Appendix

Consider Fig. 14 concerning the (lack of) effect of the simple time constant on energy equivalent continuous level.

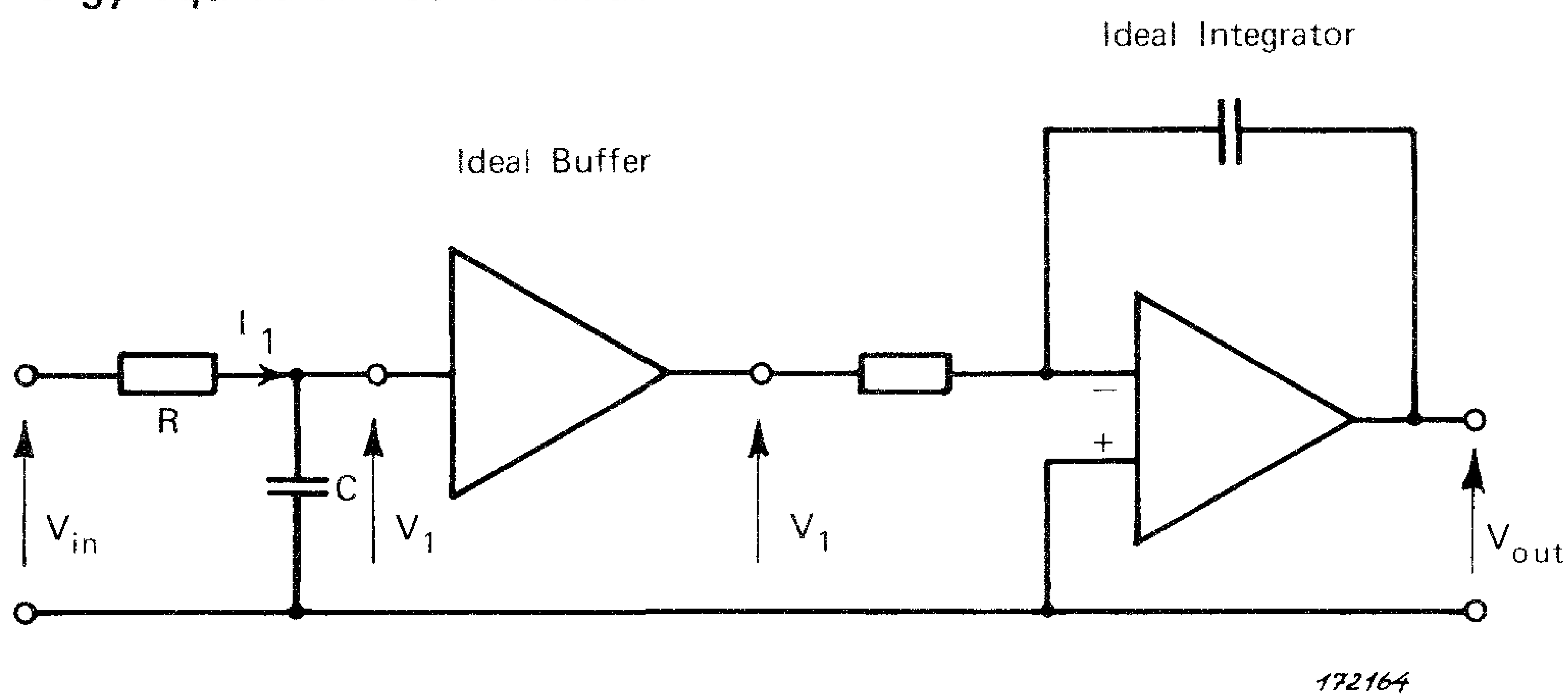


Fig. 14. Schematic Diagram of Time Constant and Integrator Circuit

Generally

$$i_1 = \frac{V_{in} - V_1}{R} = C\frac{dV_1}{dt}$$

$$V_{in} - V_1 = RC\frac{dV_1}{dt}$$

$$V_{in} = V_1 + RC\frac{dV_1}{dt}$$

Integrating $\quad \int_o^T V_{in}dt = \int_o^T V_1 dt + RC \int_o^T dV_1$

Assuming V_{in} (t = 0) = V_1 (t = 0) = V_{in} (t = T) = 0, we get

$$\int_o^T V_{in}dt = \int_o^T V_1 dt + [RCV_1]_o^T$$

$$\int_o^T V_{in}dt = \int_o^T V_1 dt$$

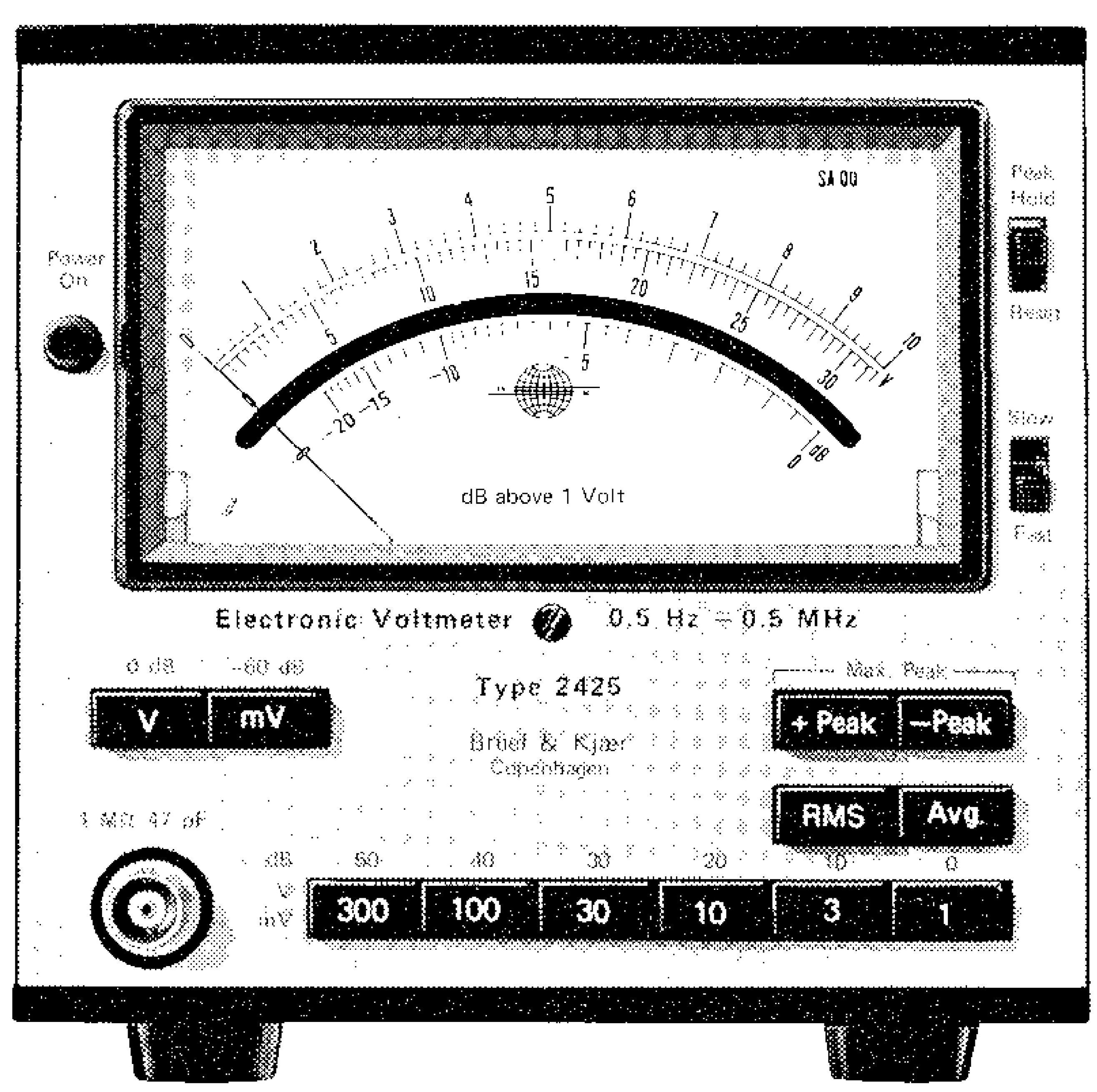
which is the desired result.

News from the Factory

Electronic Voltmeter Type 2425

The Electronic Voltmeter Type 2425 is a general purpose push button compact voltmeter well suited for measurements on signals of complex waveforms in the frequency range 0,5 Hz to 500 kHz. Besides indicating + and —Peak, true RMS and average value of signals, it contains a peak hold and a maximum peak function with manual reset for measurements on impulse signals. The calibrated amplification of the voltmeter is 60 dB and adjustable in 10 dB steps while the meter has sensitivities from 1 mV to 300 V full scale deflection. Linear AC and DC outputs are also available.

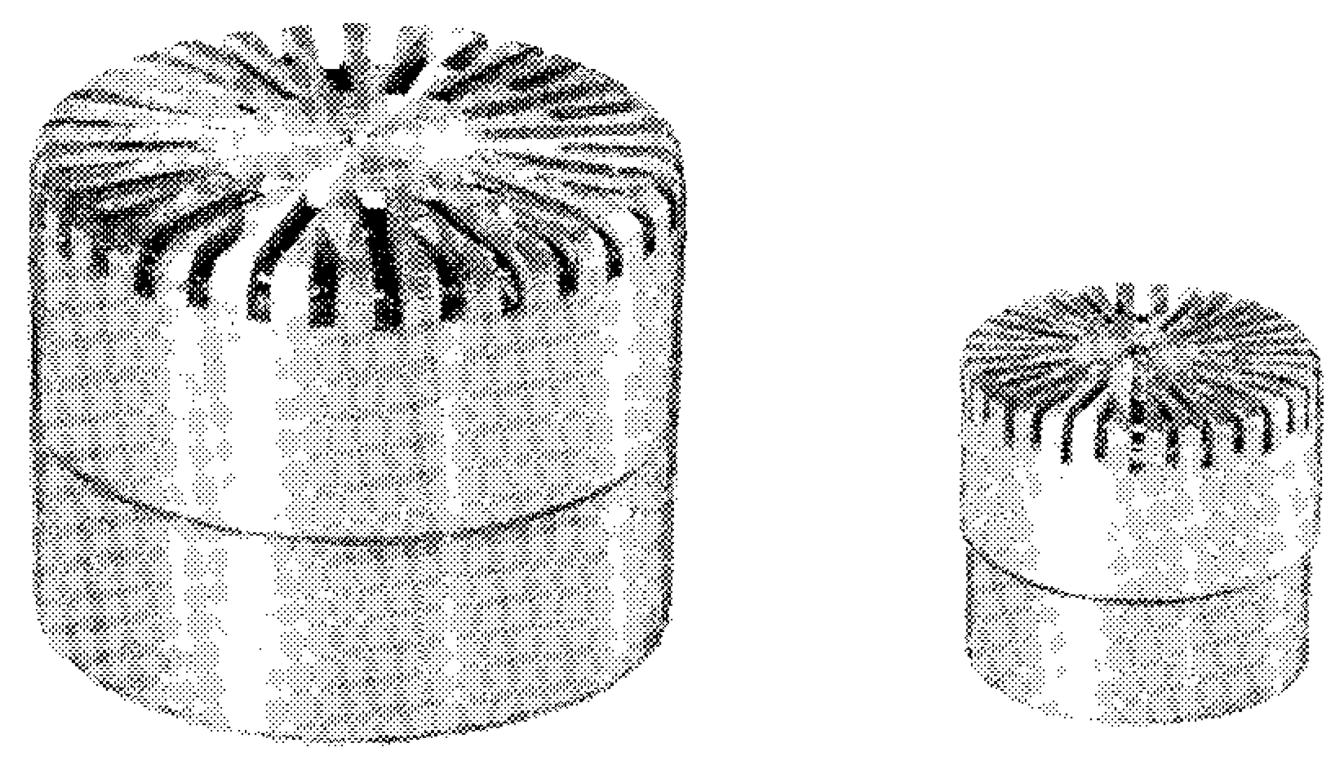
The "Fast" mode is in accordance with the standards for VU measurements, while the "Slow" mode is included for measurements on narrow band random noise. Facilities are also provided for varying the time constant by connection of external capacitors.



Microphone Cartridge Type 4161 and 4163

The one inch Microphone Type 4161 and half inch Microphone Type 4163 are essentially identical to Types 4145 and 4133 respectively, except that the capillary tube, venting the microphone has been led to the back of the microphone instead of the side. This facility makes it possible to use these two microphones with the 1" Dehumidifier UA 0310 and 1/2" Dehumidifier UA 0308 respectively.

In applications where there is a risk of water condensating inside the microphone, the Dehumidifiers containing silicagel (which effectively removes humidity from the air in the microphone) should be mounted between the microphone and the preamplifier. A small window in the Dehumidifier casing thas been incorporated to control the humidity content of the silicagel which changes colour from blue in the dry state to red when humid. The silicagel can easily be dried out again by heating for some hours at 100°C or longer at lower temperatures. When used in environment of 100% relative humidity they would require drying approximately once a month.



Quartz Coated 1/2" Condenser Microphone Type 4149

The quartz coated 1/2" Condenser Microphone Type 4149 is a high precision measuring micophone especially developed for use in severe environments. The microphone has the same excellent performance as other B & K condenser microphones, but has its diaphragm and backplate covered with a thin film of quartz.

The quartz layers, 0.7 μm thick, effectively seal pinholes and ensure that humidity does not penetrate through, thus protecting the microphone against corrosion of the diaphragm surface. On the back plate where the surface is liable to corrode on account of humidity and possible condensation (as well as very high electrical field between the diaphragm and back plate 100 kV/cm), the quartz layer provides adequate protection thus ensuring a much longer life time.

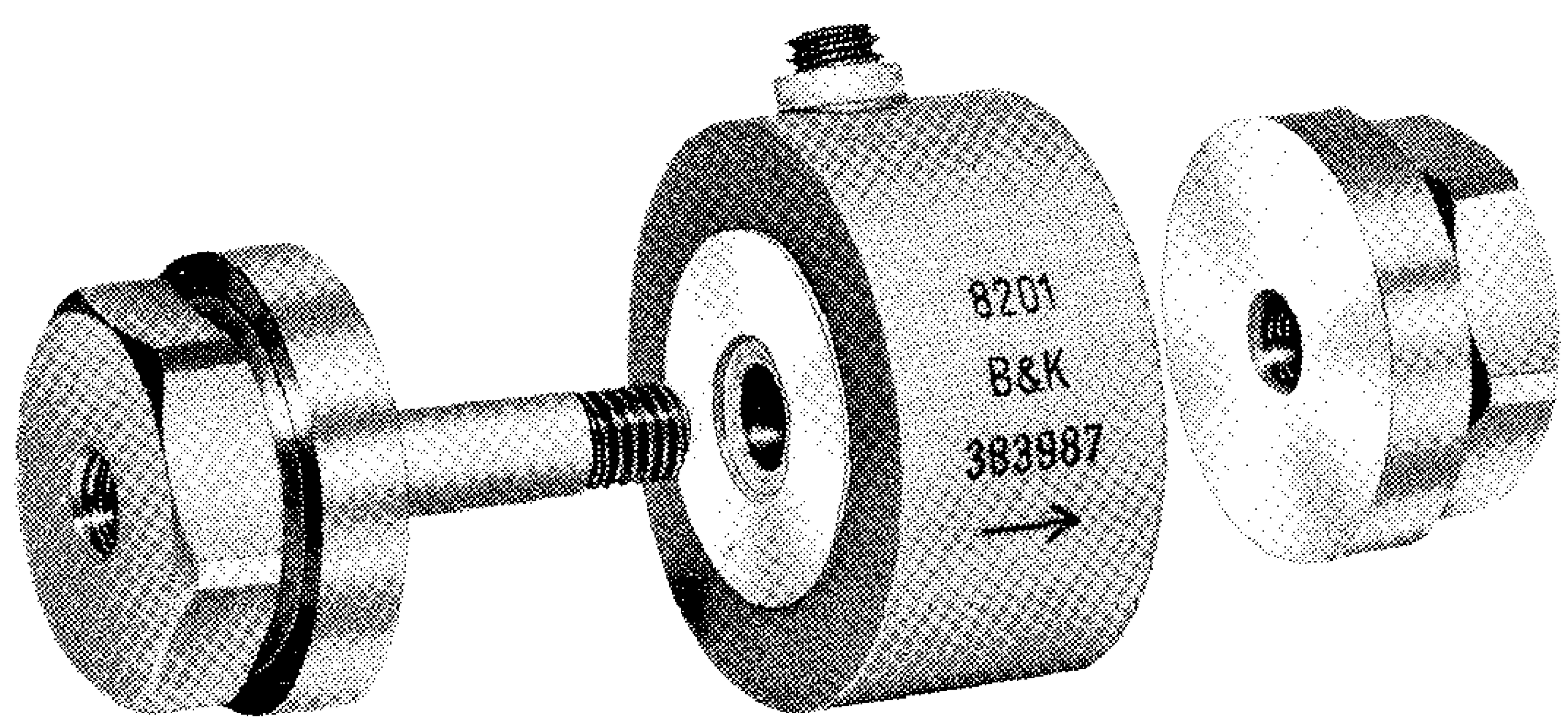
Although the mass of the diaphragm is increased by 4% its influence on the electro-acoustic performance is negligible. Each microphone is artificially aged to ensure long term stability and is supplied with an individual calibration chart giving full frequency response curve, sensitivity and all other relevant data. Since long time testing has shown a very great reduction in failure (increased noise level) compared to conventional condenser microphones, it would be ideal in applications of outdoor monitoring systems. It is possible to utilize the Dehumidifier UA 0308 also for this microphone since the capillary tube venting the microphone has been led to the back of the microphone.

Force Transducer Type 8201

The force transducer Type 8201 is a larger version of the Type 8200 with the added features of a larger force range and low sensitivity to bending moment and transverse forces. It is designed to measure dynamic, short duration static and impact, tensile and compressive forces in machinery and other constructions. When used together with an accelerometer and vibration exciter it is well suited for controlling the applied force in measurements of mechanical impedance.

The principal feature of this transducer is that the upper and lower preloading nuts can be removed enabling the transducer to be used as a load washer in compressive force measurements. After remounting the preloading nuts, the calibration is, however, still valid. Tensile forces of 4000 N (900 lbf) and compressive forces of 16000 N (3600 lbf) can be measured when the preloading nuts are mounted, while as a load washer alone compressive forces up to 20,000 N (4500 lbf) are measurable.

The transducer can be used under severe environmental conditions on account of its rugged, all welded, hermetically sealed construction and because of its ceramic insulated micro-plug connector sealed with moulded glass. The allowable temperature range is $-196^{o}C$ to $150^{o}C$. The quartz



piezoelectric element having a very low temperature sensitivity makes the transducer rather insensitive to temperature transients and tempearature changes. On the other hand the transducer has to be used together with charge preamplifiers such as Type 2624, Type 2626 or Type 2628 to secure lower limiting frequencies. The transducer when used with these preamplifiers permits measurements down to extremely low frequencies on account of its very high leakage resistance.

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(Continued from cover page 2)

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Power Spectral Density Measurements and Frequency Analysis (English)
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Instruction manuals (English, some available in German, French, Russian)
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