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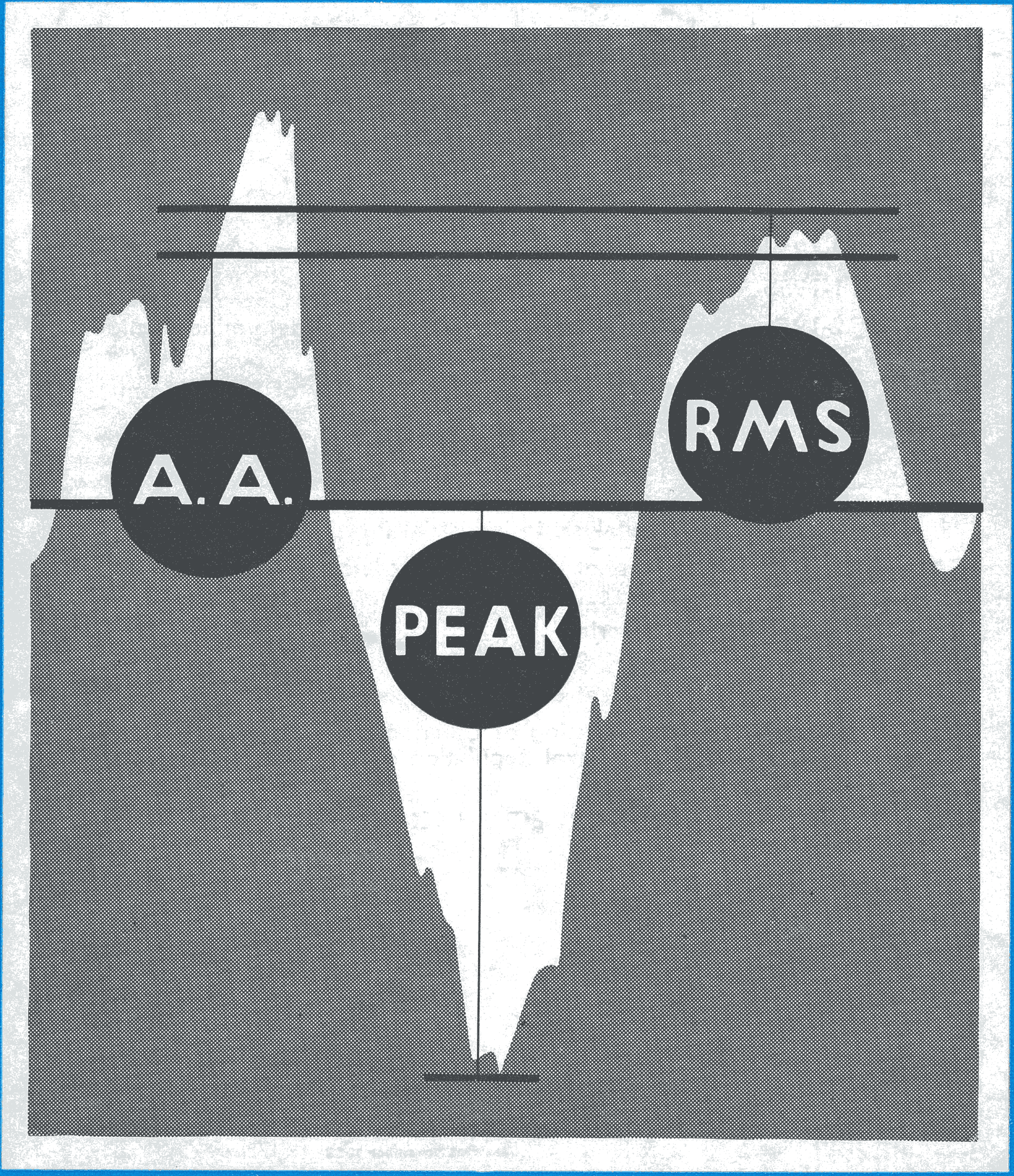

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Design features in

Microphone Amplifier Type 2603.

and

A. F. Spectrometer Type 2110.

by

Hans Michaelsen and *Henry Petersen.*

Summary.
The Audio Frequency Spectrometer Type 2110 and the Microphone Amplifier Type 2603 which are new versions of the previous models Type 2109 and Type 2602, respectively, are described. The most important improvement is that these instruments now correctly measure the true RMS value of acoustical, vibrational, and electrical signals. Besides the instruments can be switched for peak and normal average indication.

ZUSAMMENFASSUNG
Der folgende Artikel behandelt die neuen Geräte Terzfilter-Analysator 2110 und Mikrophonverstärker 2603, welche die bisherigen Modelle Typ 2109 und 2602 ablösen. Die wesentliche Verbesserung die in den neuen Geräten eingeführt ist, ist die umschaltbare Gleichrichterschaltung, mit der sowohl der Effektivwert als auch der Spitzen- und Mittelwert gemessen wird.

SOMMAIRE
On décrit le spectromètre basse fréquence type 2110 et l'amplificateur de microphone type 2603, versions nouvelles des modèles précédents type 2109 et type 2602.
L'amélioration la plus importante apportée à ces appareils est qu'ils mesurent à présent la valeur efficace vraie des signaux acoustiques, vibratoires et électriques. En outre par commutation ils permettent une lecture soit de la valeur de crête, soit de la valeur moyenne.

As stated in the standards for the measurement of sound and vibration a true RMS indicating instrument is required, and because instruments indicating the true RMS value have not previously been commercially available for this purpose, the need for the instruments Type 2110 and 2603 is very great.
The true RMS indication in connection with a frequency response characteristic which is linear from 2 to 35000 c/s now enable the full advantage to

be drawn from the Accelerometers Type 4308 and 4309. An additional frequency characteristic from 20 to 20000 c/s together with the internationally standardized weighting networks for sound level measurements make, furthermore, the instruments suitable for very accurate acoustical measurements. As mentioned in the succeeding article, the instruments incorporate a switch for selection between true RMS, peak, (half peak to peak) and average indication on the built-in meter, the damping of which is changeable between the

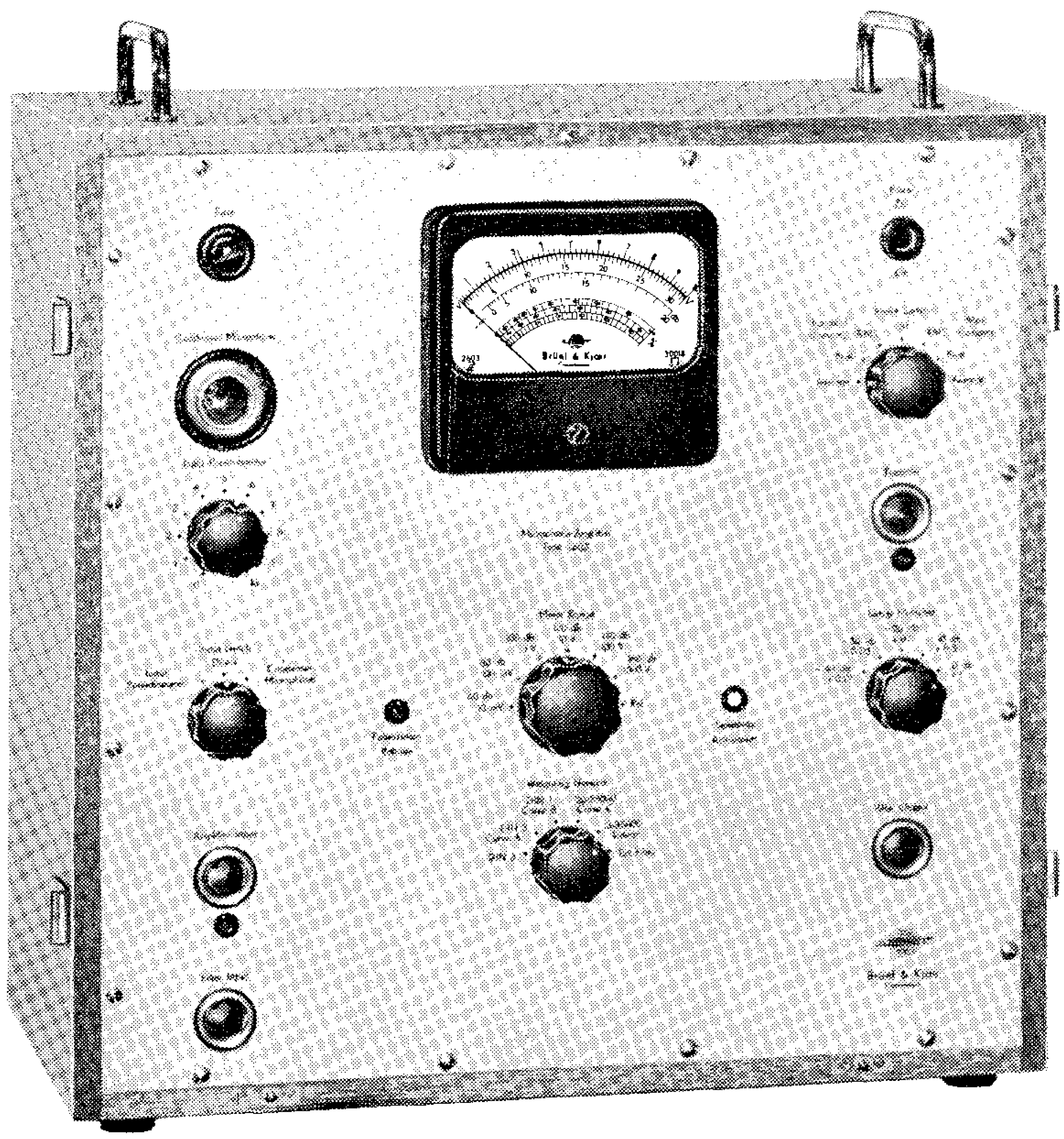


Fig. 1. Photo of the Microphone Amplifier Type 2603.

one standardized for sound level measurements and an increased damping for measurements of very low frequency signals. The new RMS rectifier is described in the succeeding article.

Like the previous models, the new instruments have a total hum-level which is less than 2 μV on the grid of the first tube in the amplifier. This is obtained by using DC heating for the filaments in the first tubes. The effect of the hum signal is clearly seen from fig. 3. Underneath the system of co-ordinates is shown a simplified block-diagram of the Microphone Amplifier Type 2603. The diagram for the Spectrometer Type 2110 is the same except for the thirty $^1/_3$-octave filters which are inserted between the Amplifier I and the "Range Multiplier". In the system of co-ordinates the signal levels are depicted so that the abscissae correspond to the place in the instruments indicated by the block-diagram. From the figure is seen that the "Meter

Range" attenuator ensures a maximum voltage on the input of Amplifier I of 10 mV when correct setting is used. The amplification in Amplifier I is constant 40 db. The maximum voltage which can be supplied to the filters ("Filter Input" terminal) is consequently 1 V presuming a correct setting of the attenuators. The "Range Multiplier" attenuator which preceeds the Ampli-

Fig. 2. Photo of the A.F. Spectrometer Type 2110.

fier II ensures a maximum voltage of 10 mV on the input grid of Amplifier II. This amplifier has a constant gain of 60 db corresponding to an output voltage of 10 V full meter scale deflection. The dotted line which starts at 2 μV indicates the hum-level. The hum-signal from Amplifier I is amplified and added to a hum level of approx. 10 μV on the Amplifier II input. It is seen that the hum level at the output terminal depending on the setting of the "Range Multiplier" is smaller than 210 mV in the worst case, it is with zero attenuation in the "Range Multiplier", and less than 12 mV at its minimum. The hatched area in the Meter Circuit section indicates the usable meter range, that is the range covered by the meter scale, and the hatched area in the "Recorder" section indicates the usable voltage range

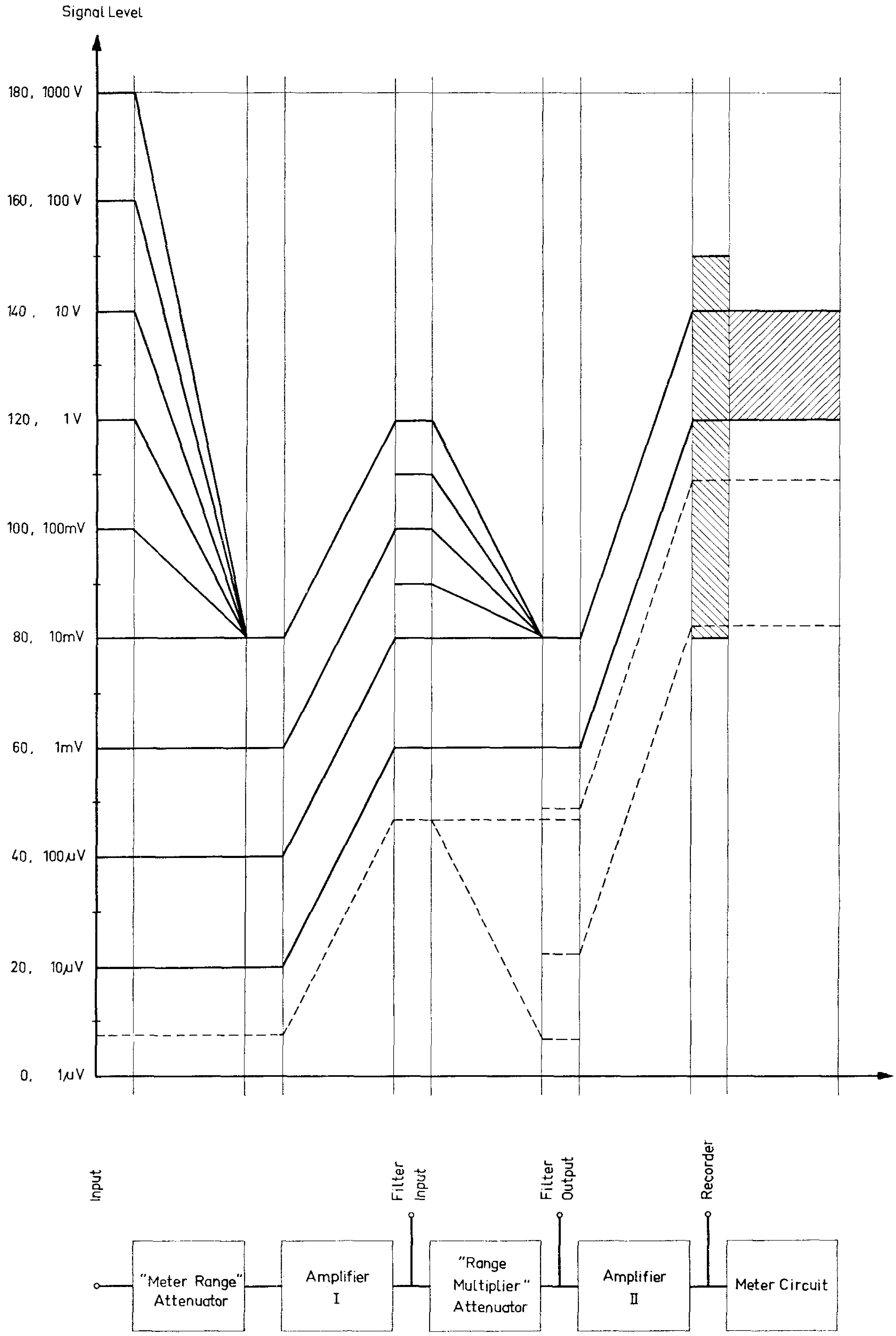


Fig. 3. The figure shows the signal level in the Microphone Amplifier and the A.F. Spectrometer as a function of the place in the two instruments.

on the "Recorder" output bushing using the Level Recorder Type 2304. The lower limit is set by the Level Recorder which does not respond to voltages smaller than 10 mV, and the upper limit is set by the maximum output voltage available on the "Recorder" output bushing. It is seen that this limit lies 10 db higher than full meter deflection. This voltage can be obtained without overloading *"Amplifier II"*, and the corresponding total gain of the

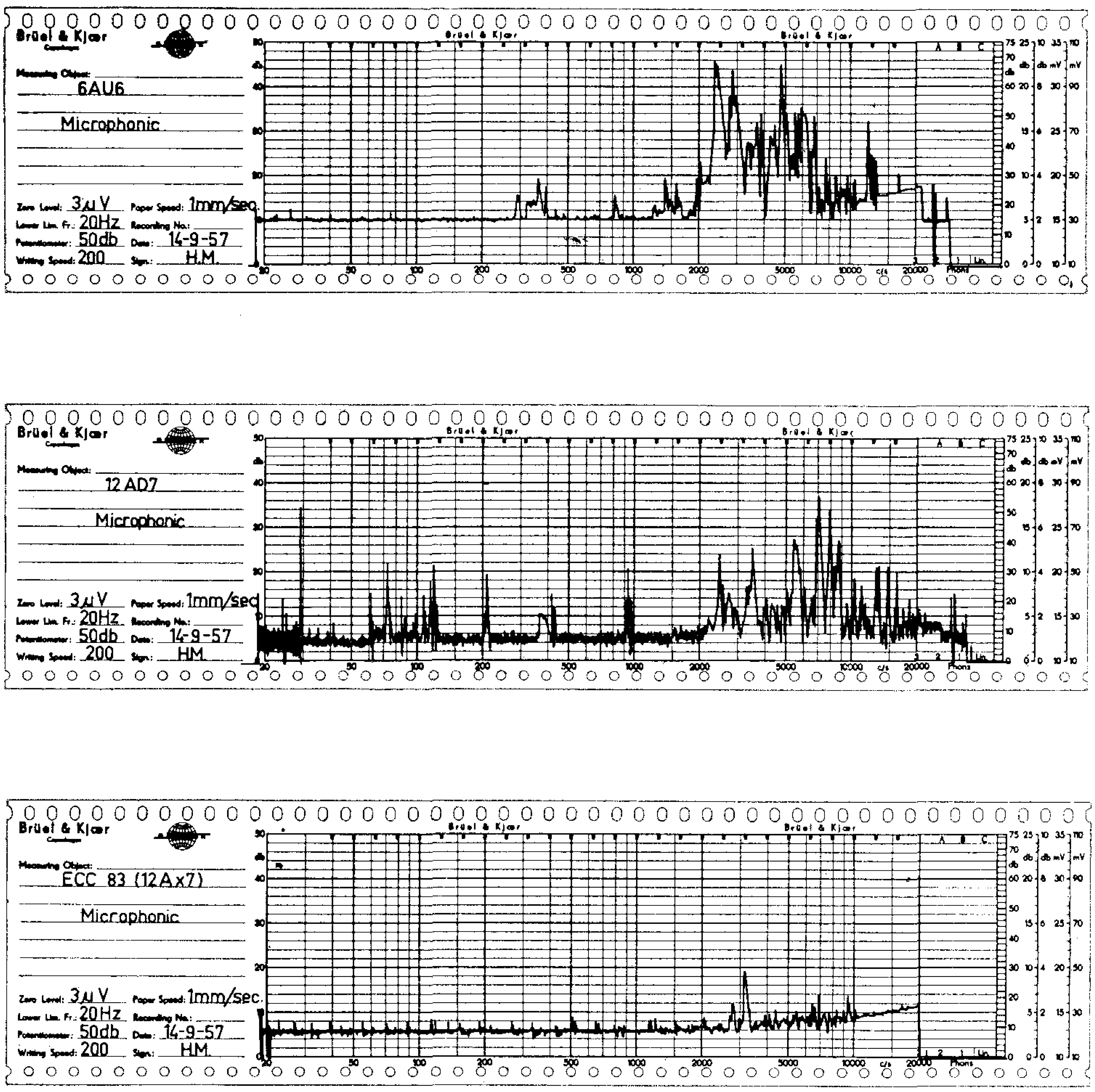


Fig. 4. Recording of microphonic signals in the Microphone Amplifier and the A.F. Spectrometer, when subjected to a sound pressure level of 100 db re. 2 × 10^{-4} μbar for three different amplifier tubes.

instruments must, consequently, be used only by decreasing the attenuation of the "Range Multiplier" by an amount of 10 db.

To ensure a very low microphonic level of the Amplifier, every amplifier tube socket is mounted on a rubber plate acting as shock-absorber, and experiments have been carried out to find the type of tubes which showed the smallest microphonic effect. (The type had to be found among the types available all over the world). In the experiments the amplifier under test was placed in a sound field which, at a 100 db sound pressure level was frequency scanned over most of the frequency range of the amplifier. The output signal of the amplifier was then recorded by means of the Level Recorder Type 2304. In fig. 4 is shown some recordings obtained for different tubes. On the basis of these measurements, the twin-triode ECC83 (12AX7) was chosen.

Another feature of the instruments is the low output impedances on the bushing "Filter Input" and "Recorder". The output impedance at "Filter Input" is as low as 10 Ω in order to decrease the variation of the input voltage with a variation in load impedance. Used in connection with the $^1/_3$-Octave Filter Set Type 1610 (built-into the Spectrometer Type 2110), the input impedance of which varies from approx. 500 to 2000 Ω, the corresponding voltage variations are smaller than 1 %. This low impedance is obtained by a paralleling the two sections of a high-G twin-triode coupled as a cathode follower, the output impedance of which is approx. $\frac{2}{2\,G}$, where G is the transconductance of the tube. The output impedance at the "Recorder" bushing is 50 Ω. Due to the wide frequency range and the high gain, an output transformer is not suitable for reduction of the output impedance. The low value is also here obtained by means of a high-G triode coupled as a cathode-follower.

In order to make the corrections for the change in frequency response as a

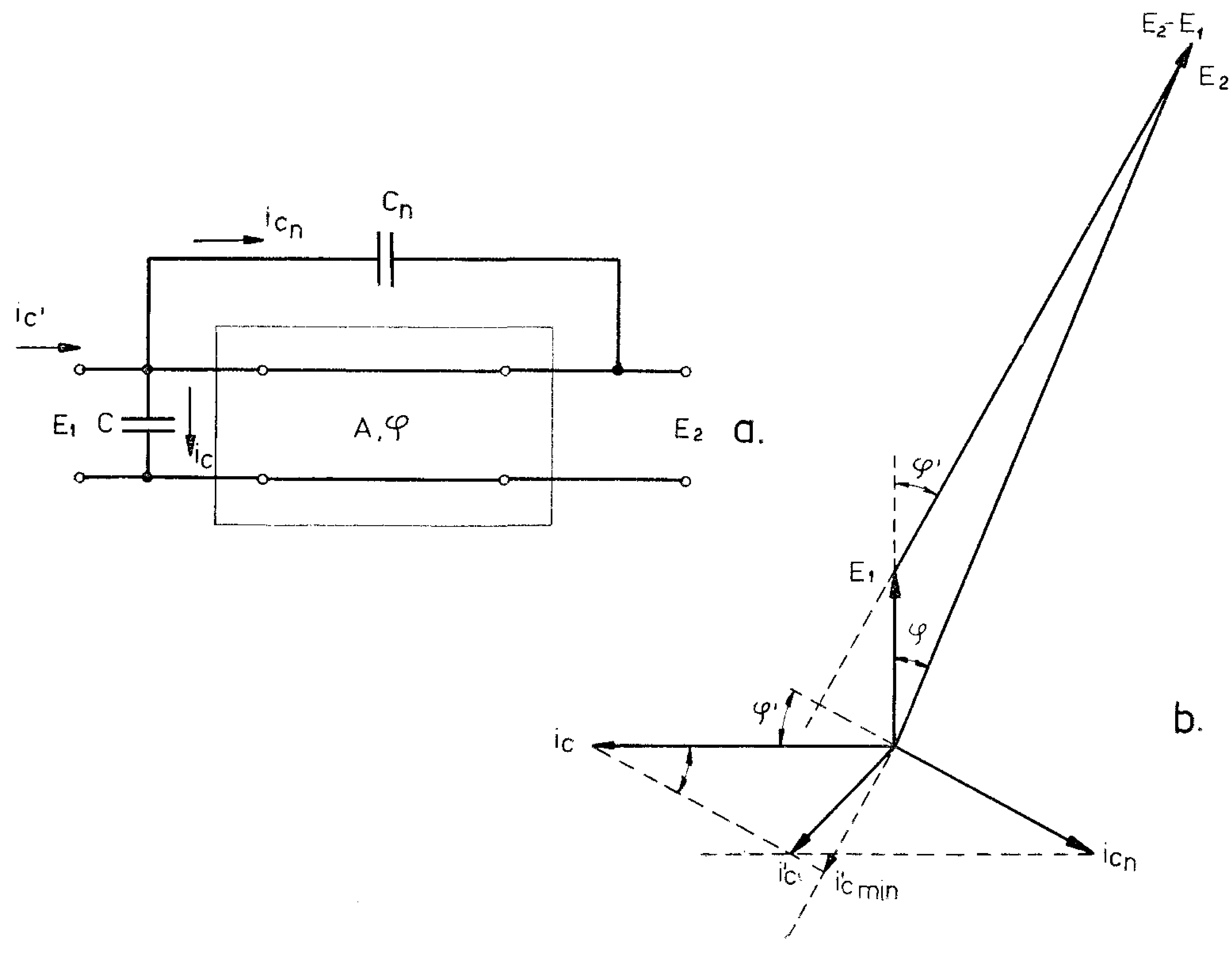


Fig. 5.
a. The coupling used to reduce the input capacity of Amplifier II.
b. Vector diagram showing the voltage-current situation in the circuit.

function of the "Meter Range" attenuator setting as simple as possible, the resultant input capacity of the Amplifier II ought to be zero. This can be obtained approximately by means of a very simple coupling as shown in fig. 5. The amplifier shown has a gain of A and a phase angle shift of φ, where φ is presumed to be smaller than $90° + (n \times 360°)$. In the vector diagram is shown the input voltage E_1, the capacitive current $i_c = E_1 \times jC\omega$ through C, the output voltage E_2 ($|E_2| = A|E_1|$), the voltage $E_2 - E_1$ across C_n and the capacitive current $i_{cn} = (E_1 - E_2)\, jC_n\omega$. The total capacitive current i_c', is consequently $i_c' = i_c + i_{cn}$. From the figure is seen that the minimum value $i_c'{}_{min}$ of i_c' for a certain phase angle φ is $|i_c'{}_{min}| = |i_c| \sin \varphi' \sim |i_c| \sin \varphi$ for $A >> 1$. $i_{c\,min} = 0$ is consequently obtained only for $\varphi = 0\ (+ n \times 360°)$.

The corresponding value of C_n is found to be $C_n = \dfrac{C}{A} \cos \varphi$.

For Type 2603 and 2110 $\varphi < 7.5°$, and consequently $C_n \sim \dfrac{C}{A}$ whereby $i_c' = i_c \sin \varphi \sim 0$, so the input capacity is zero.

Finally should be mentioned how a reference voltage which is perfectly independent of a variation of $\pm$ 10 % in the power supply voltage is obtained. In fig. 6a is shown a normal circuit for stabilizing the reference voltage. By

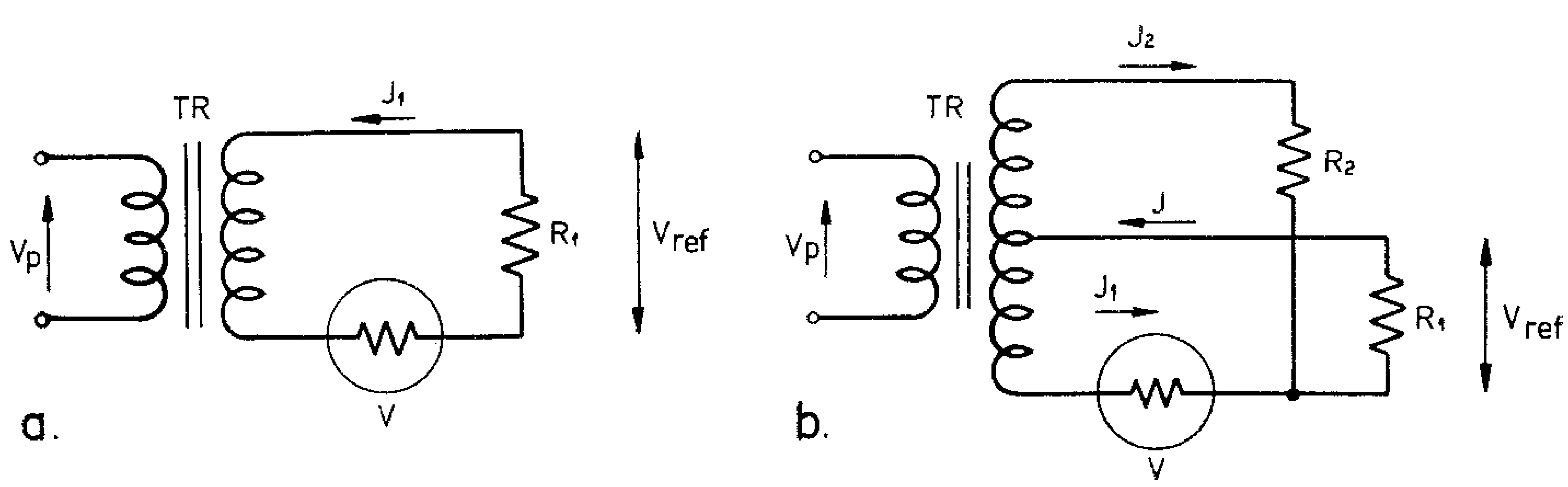


Fig. 6.
a. Normal way of obtaining a stabilized reference voltage.
b. Extra compensation introduced to reduce voltage variations to zero.

means of a transformer TR the supply voltage V_p is reduced to a suitable value. The current I_1 in the secondary circuit of the transformer is stabilized to vary only 1.4 % for a 10 % variation in V_p by means of the current regulator V. The ratio between these variation is consequently

$$\frac{\Delta I_1}{\Delta V_p} = \frac{1.4}{10}$$

= 0.14. From this is found $I_1 = 0.14 \times V_p + K$, where K is a constant. By subtracting a current $I_2 = 0.14 \times V_p$, the total current $I = I_1 - I_2$ would then be $I = K$ independent of the voltage V_p. This is only correct for variations in V_p which do not exceed the regulating range of the current regulator.

Because the subtraction of two inphase currents is the same as adding the corresponding in-antiphase currents, the problem is solved by designing the transformer TR with two secondary windings as shown in fig. 6b. The currents I_1 and I_2 are 180° out of phase. The correct numerical value of the current I_2 is then obtained by proper designing of the Resistor R_2 and the transformer ratio. The reference voltage is obtained as the voltage drop of the current I across a stable resistor R_1.

A true RMS Instrument.

by

C. G. Wahrman, M.Sc.

Summary.

It is shown that by a simple change in the design of common average type instruments an approx. RMS indication can be obtained. Based on this design a more complicated circuit is developed which is accurate to within o.5 db for the measurement of signals with crest factors up to 5. This circuit is used in the Spectrometer Type 2110 and the Microphone Amplifier Type 2603.

ZUSAMMENFASSUNG

Es wird gezeigt, wie ein gewöhnlicher Mittelwertgleichrichter in einfacher Weise so geändert werden kann, daß eine Art von Effektivwertgleichrichtung entsteht.

Ausgehend von diesen Betrachtungen wird ein etwas komplizierterer Gleichrichter entwickelt, welcher eine Genauigkeit von o.5 dB, bei der Messung von Signalen mit Scheitelwerten bis zu 5 aufweist.

Dieser Gleichrichter findet sowohl in dem Terzfilter-Analysator Typ 2110 als auch in dem Mikrophonverstärker Typ 2603 Anwendung.

SOMMAIRE

On montre comment, par un simple changement dans la conception des appareils du type moyen courant, on peut obtenir une indication approchée de la valeur efficace. On établit ensuite sur cette base un circuit plus élaboré, d'une précision de o,5 dB, pour la mesure de signaux présentant un facteur de crête allant jusqu'à 5. Appel est fait à ce circuit pour le spectromètre type 2110 et pour l'amplificateur de microphone type 2603.

When it is desired to characterize the magnitude of an AC signal by means of a single figure the problem arises: Which property of the signal is the most useful one to know. That is, which one of the three commonly used values, the peak, the average or the RMS value should be chosen. (See Appendix page 20).*

In the case of a square-wave signal this is no problem because the peak value is the only amplitude existing in the signal, and this value is also equal to the average and the RMS value. However for all other wave-forms the above mentioned problem always exists. If the signal is displayed on the screen of an oscilloscope the most obvious property to determine would seem to be the peak or the peak-to-peak value. This value is also in some cases the most valuable to know, for example when considering whether a given amplifier is overdriven or not by the signal in question.

For many other purposes the peak or the peak-to-peak value would give little or no information unless also the wave-form is known, and should therefore not be used as "characteristic quantity". As an illustrative example can be

*) Other values than the three mentioned here might be used, but will not be considered in the following.

mentioned the measurement of random noise which theoretically has an infinitely high peak value.
Another quantity which would seem far more valuable is the arithmetic average value of the signal, or rather the rectified average value (the average value of a pure AC signal is zero). This quantity depends on the whole wave-form, and not, as the peak value, of only one or two points of the wave-form. However, because it is relatively seldom in the mathematical treatment of AC voltages an expression is obtained in which the arithmetic average value is involved, even this quantity seems to be of limited importance.
In most cases the third quantity, the RMS value of the signal, will be the most convenient one to know. One of the reasons is that in linear circuits the dissipated power depends directly upon this value. In fact, the RMS value is so commonly used to characterize the magnitude of AC signals that this is normally the quantity given unless otherwise specially stated.
Because the RMS value of the signal is such an important quantity it has become common practice to calibrate AC instruments to indicate the RMS value of a sinusoidal signal, even if the instrument itself measures the average or peak value. The use of a sinusoidal signal for calibration has been found convenient because this wave-shape is very often met in practice. This means, however, that the reading on the instrument meter will be incorrect whenever other than sinusoidal signals are being measured. If it is desired to measure a non-sinusoidal voltage or current it is therefore important to employ an instrument which really *measures* the RMS value.
In the case of relatively low-frequency signals (power line, lower part of audible range) it is possible to use moving-iron or electro-dynamic meters to obtain a true RMS indication. At higher frequencies thermo-couples or hot-wire meters can be used. These instruments, however, are slow and will normally not stand much overload. Basically also electrostatic type meters

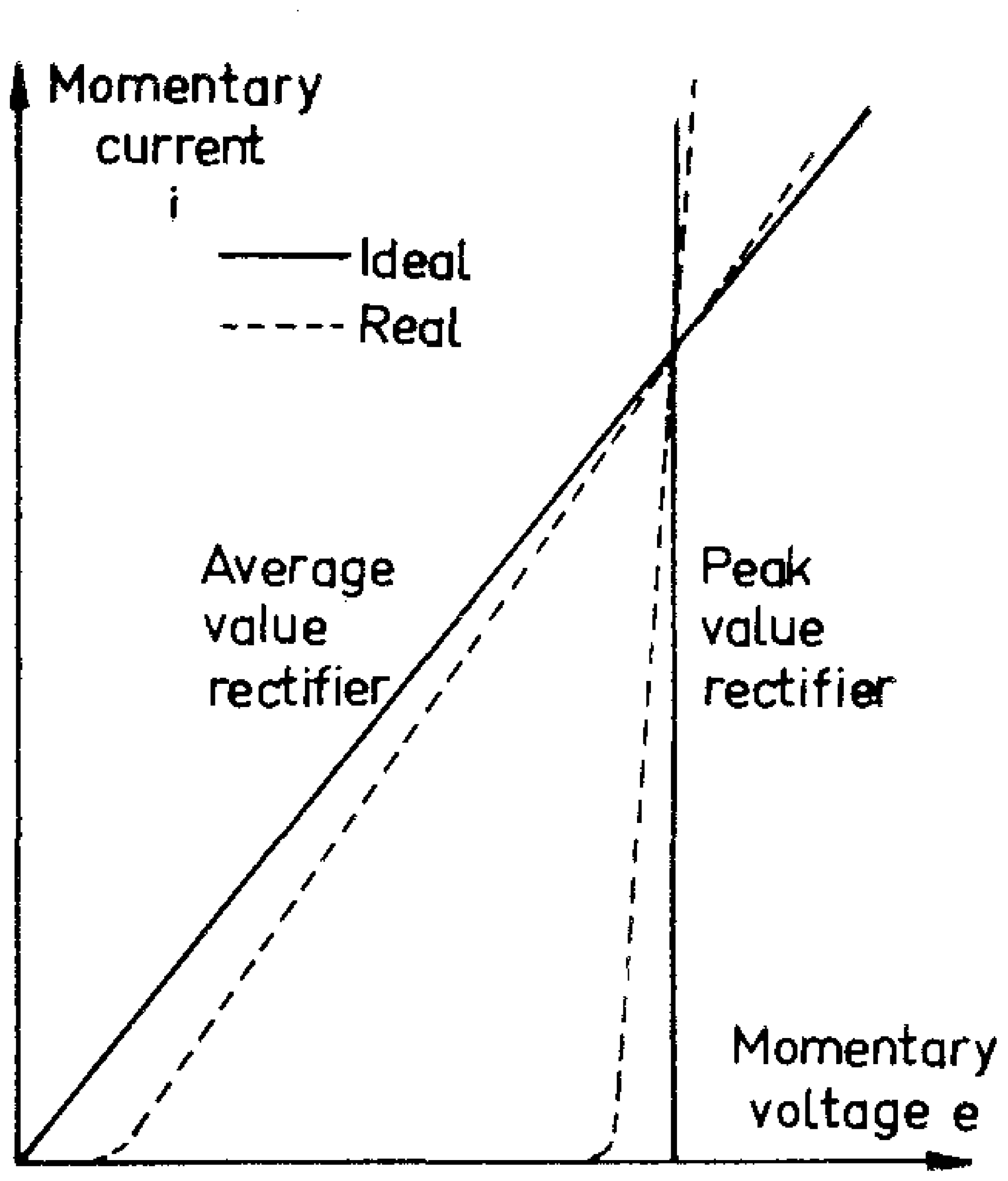


Fig. 1. Current voltage characteristics of an average and a peak value rectifier circuit.

could be used but, unfortunately, these require a relatively high voltage to drive a normal indicating mechanism (meter pointer). It has therefore been common practice for the measurement of signals with frequencies up to the lower part of the broadcasting bands to use moving coil instruments with associated rectifiers the result being an arithmetic average type instrument. In the following a method will be outlined by means of which the rectifier circuit used on these types of instruments can be modified to measure the true RMS value of the input signal.

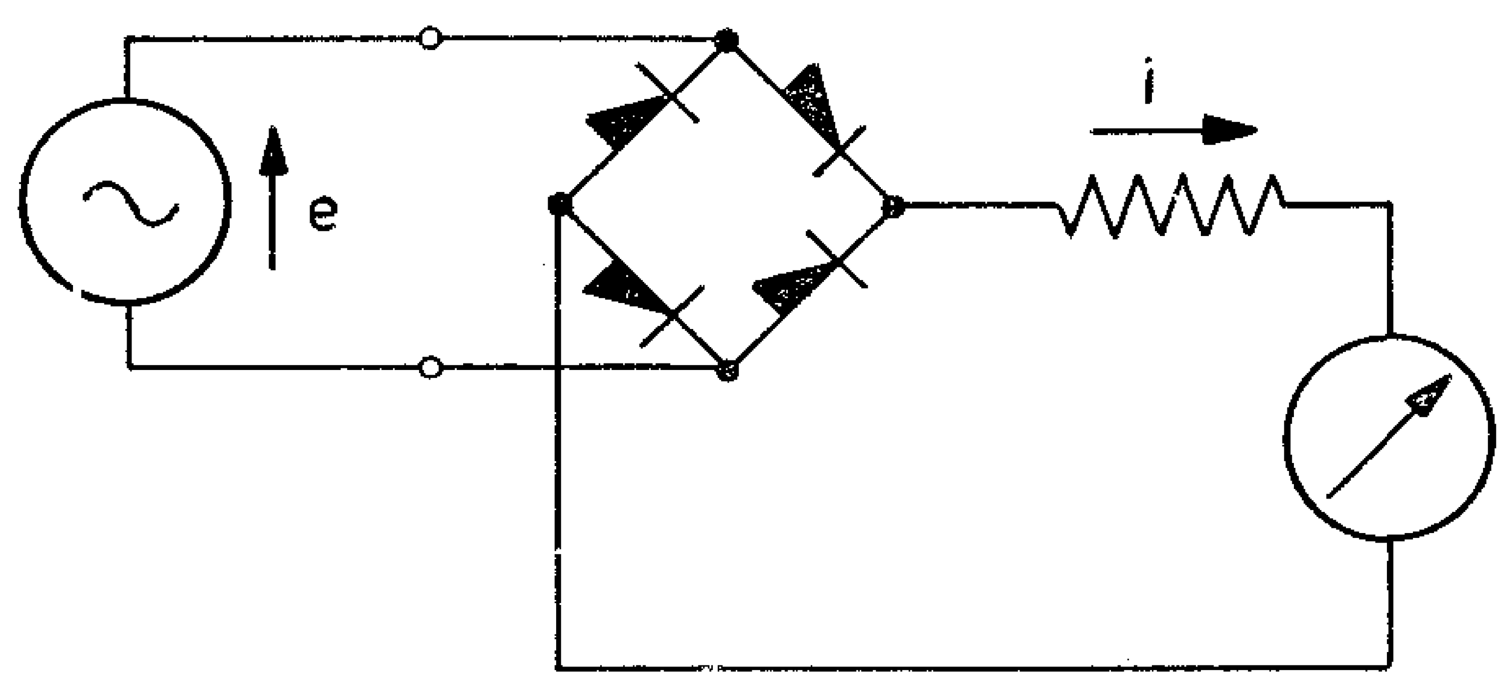


Fig. 2. Schematic diagram of an average value rectifier circuit.

The current-voltage characteristics of an ideal average and an ideal peak rectifier are shown in fig. 1. On fig. 2 and 3 are shown the schematic diagrams of these rectifiers. In the case of the arithmetic average type rectifier the ideal current-voltage characteristic is a straight line through zero, i.e. the momentary current is proportional to the momentary input voltage. In the case of the peak rectifier the current is zero until the momentary voltage exceeds the voltage on the capacitor. The current will then be relatively great and charge the capacitor to the new peak value. In practice ideal rectifier characteristics do not exist and the characteristics will be more like those shown with dotted lines in fig. 1.

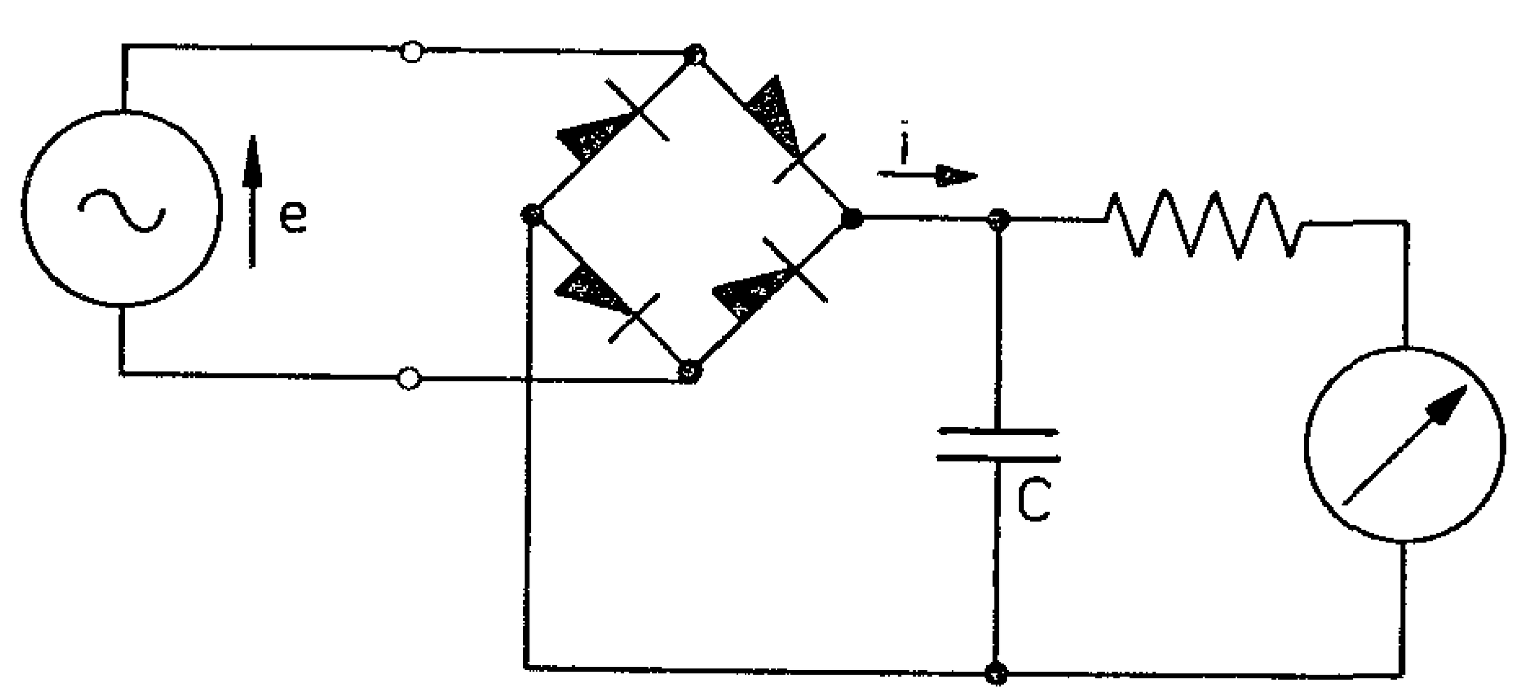


Fig. 3. Schematic diagram of a peak value rectifier circuit.

In fig. 4 is shown a current voltage characteristic of an ideal RMS rectifier as well as the characteristic of a rectifier which is neither of the peak nor of the average type. The ideal RMS characteristic is a parabola. It can, furthermore, be seen from fig. 4 that if the rectifier characteristic is given a suitable slope a much better approximation of the parabola is obtained than by means of the arithmetic average or peak type rectifier circuit. Fig. 5 shows the schematic diagram of a rectifier circuit which will give the approximate RMS characteristic shown in fig. 4. From fig. 6 can be seen how the characterstics are "displaced" as a function of the DC voltage across the

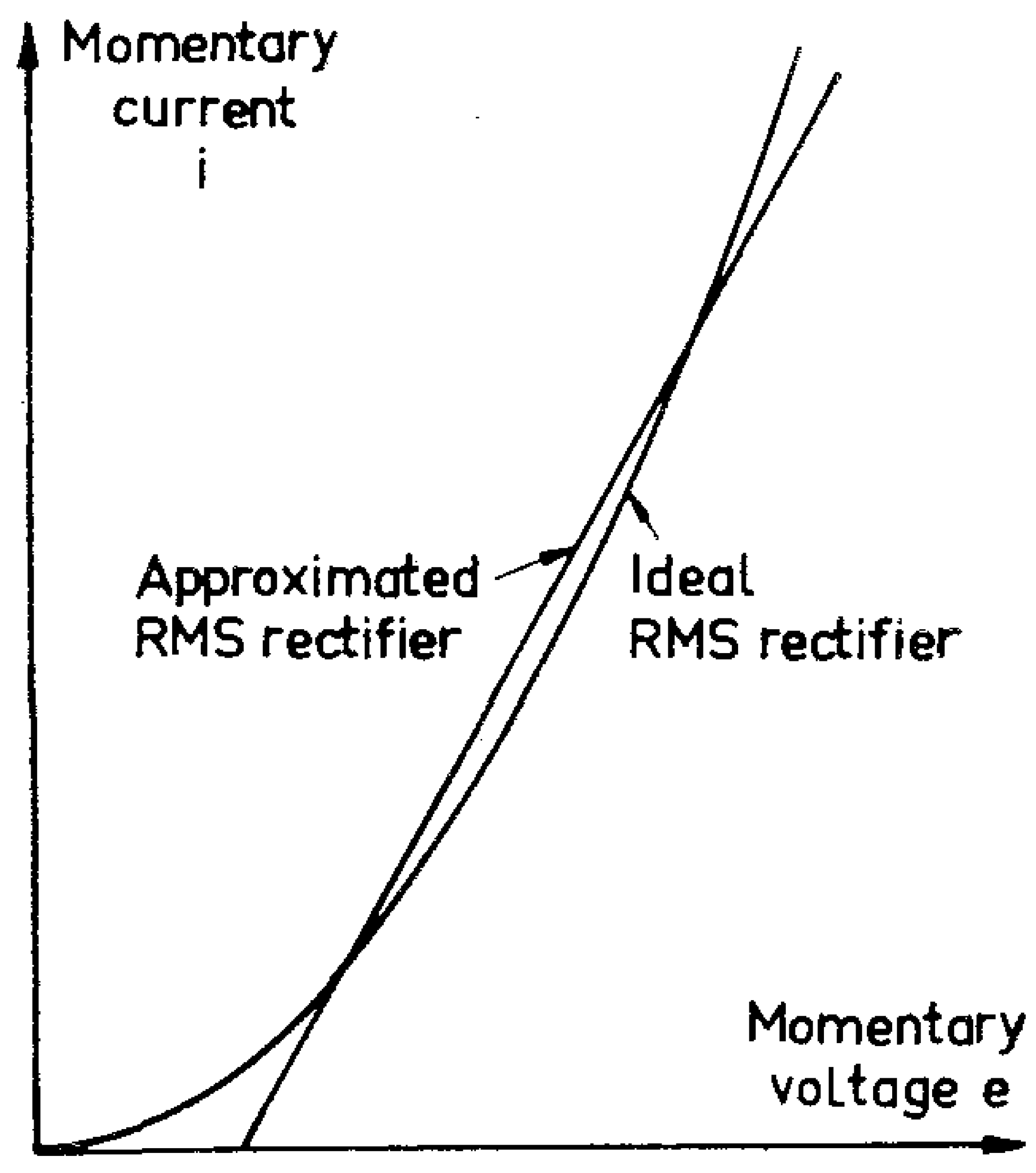


Fig. 4. Current voltage characteristic of an ideal and an approximated RMS value rectifier circuit.

capacitor. The "displacement" of the characteristics is equivalent to a multiplication of all the distances from the curve to the origin or which is the same, a multiplication of all abscissae and ordinates by a constant. This means that if, for instance, full scale deflection is obtained for a certain input voltage, half scale deflection is obtained for half the voltage as long as the

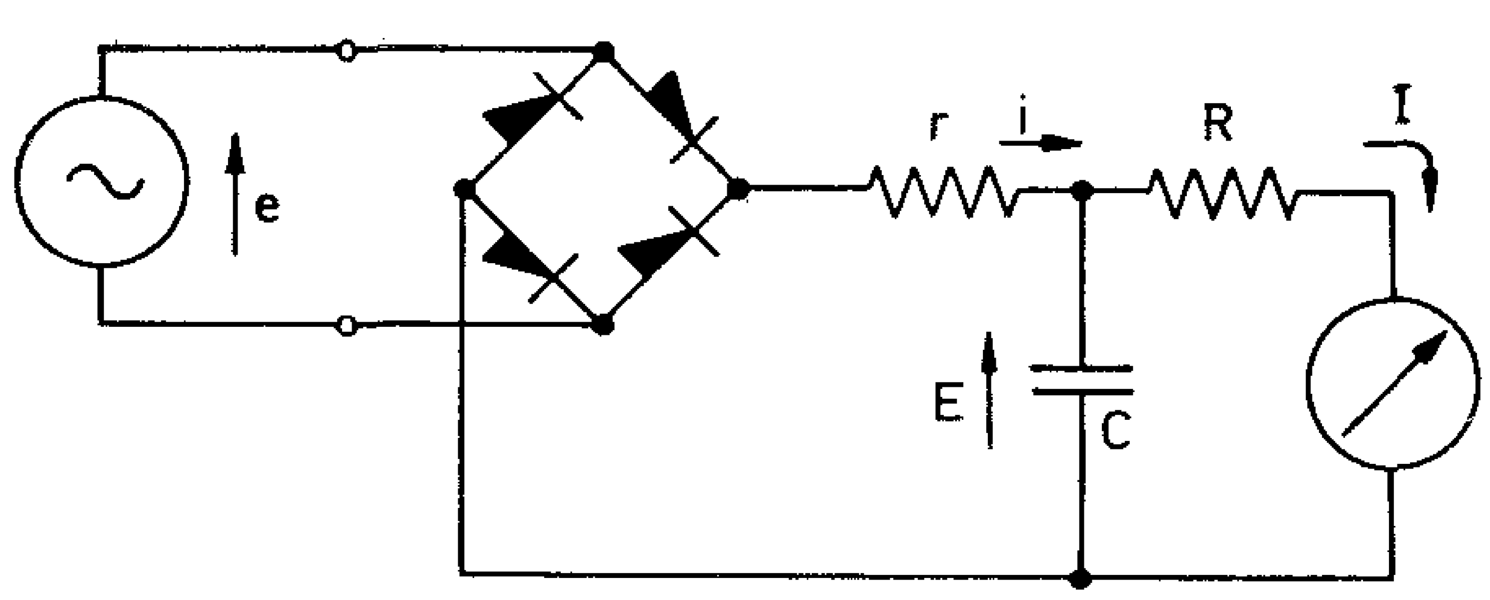


Fig. 5. Schematic diagram of that RMS rectifier circuit which gives the approximated characteristic of fig. 4.

wave-form of the signal is the same. The meter scale will thus be linear even though the rectifier characteristic is (approximately) parabolic. In practice the scale will not be exactly linear due to the non-linear characteristic of the diodes in their forward direction.

From fig. 6 can also be seen that the relative deviation between the approximated and ideal parabola will be the same for different meter deflections, or in other words that a certain wave-form is measured with the same percentage error at any point on the scale.

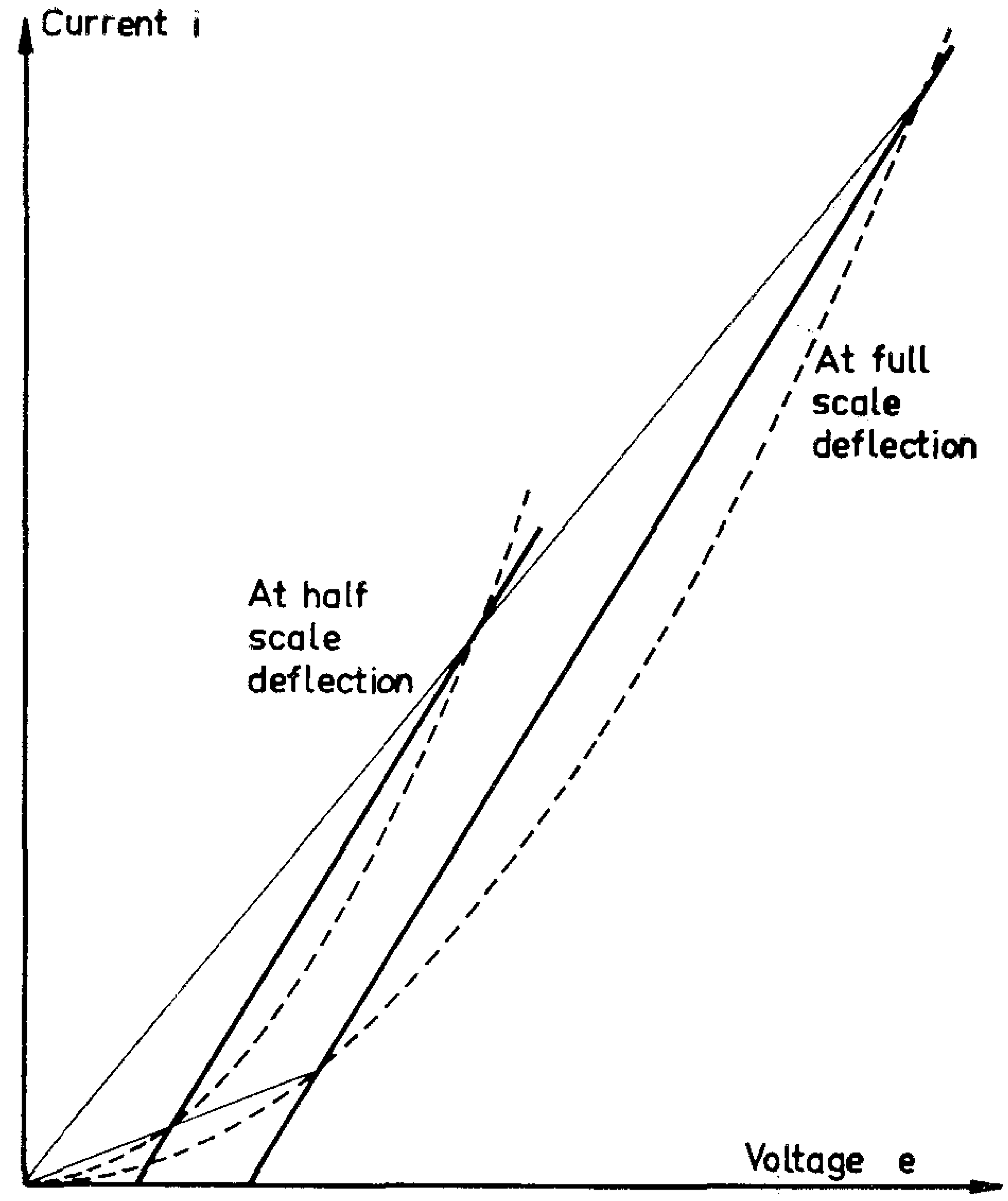


Fig. 6. Illustration of the "displacement" of the characteristics.

The problem is now to find that ratio $\frac{R}{r}$ between the two resistors shown in fig. 5, which gives the best approximation to the parabola. In fig. 7 a

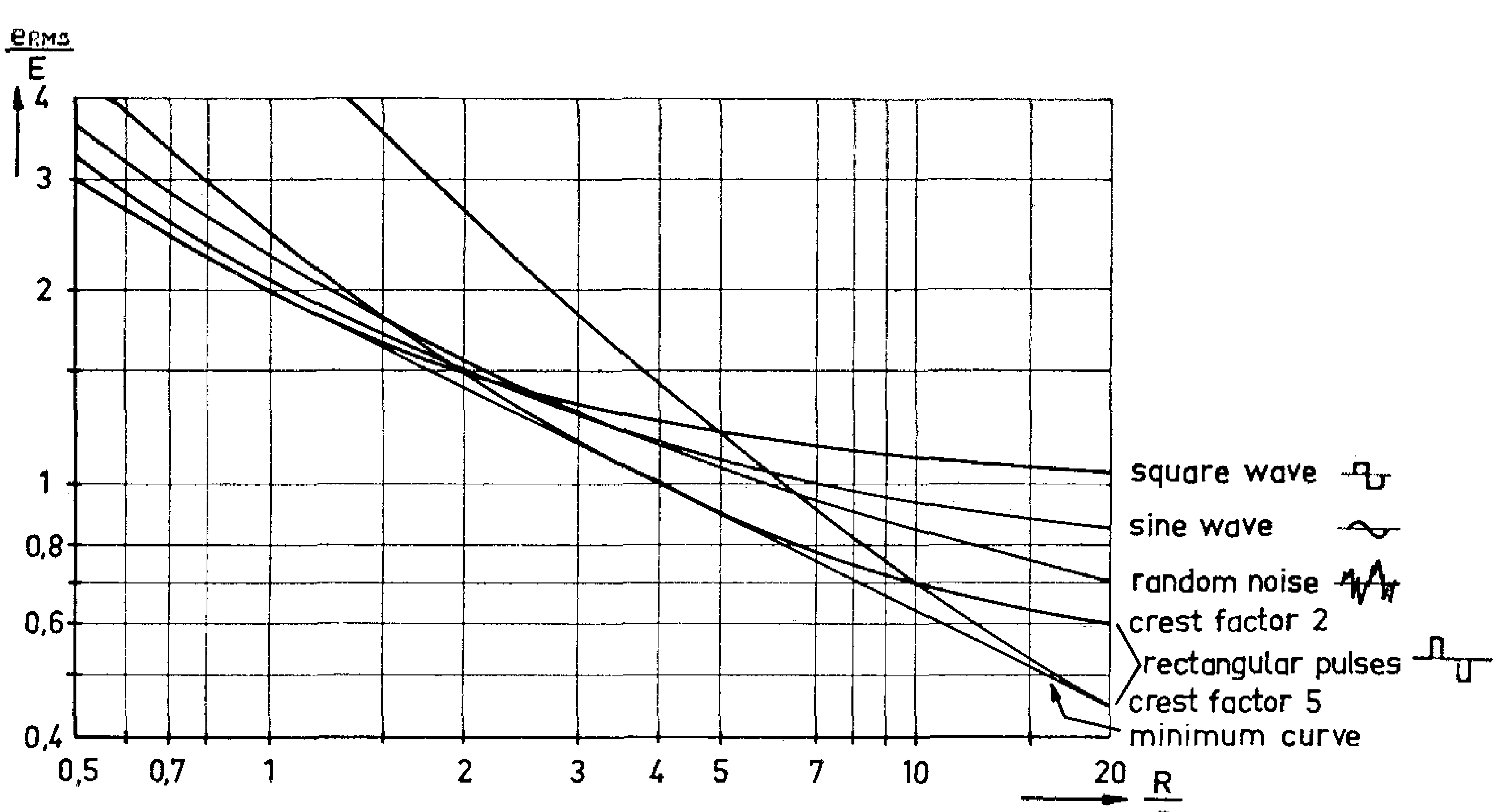


Fig. 7. Curves showing the ratio $\frac{e_{RMS}}{E}$ *as a function of the ratio* $\frac{R}{r}$ *for differently shaped signals. See also fig. 5.*

number of curves are plotted which show the ratio of the RMS input voltage, e_{RMS}, and the voltage across the capacitor, E, as a function of the ratio $\frac{R}{r}$ for different signal wave-forms. These curves are calculated theoretically as follows:

1. Periodic Signals.

In case of a periodic signal (for example a sine-wave, see fig. 8) a certain charge is always stored in the capacitor and only the voltage-time integral of

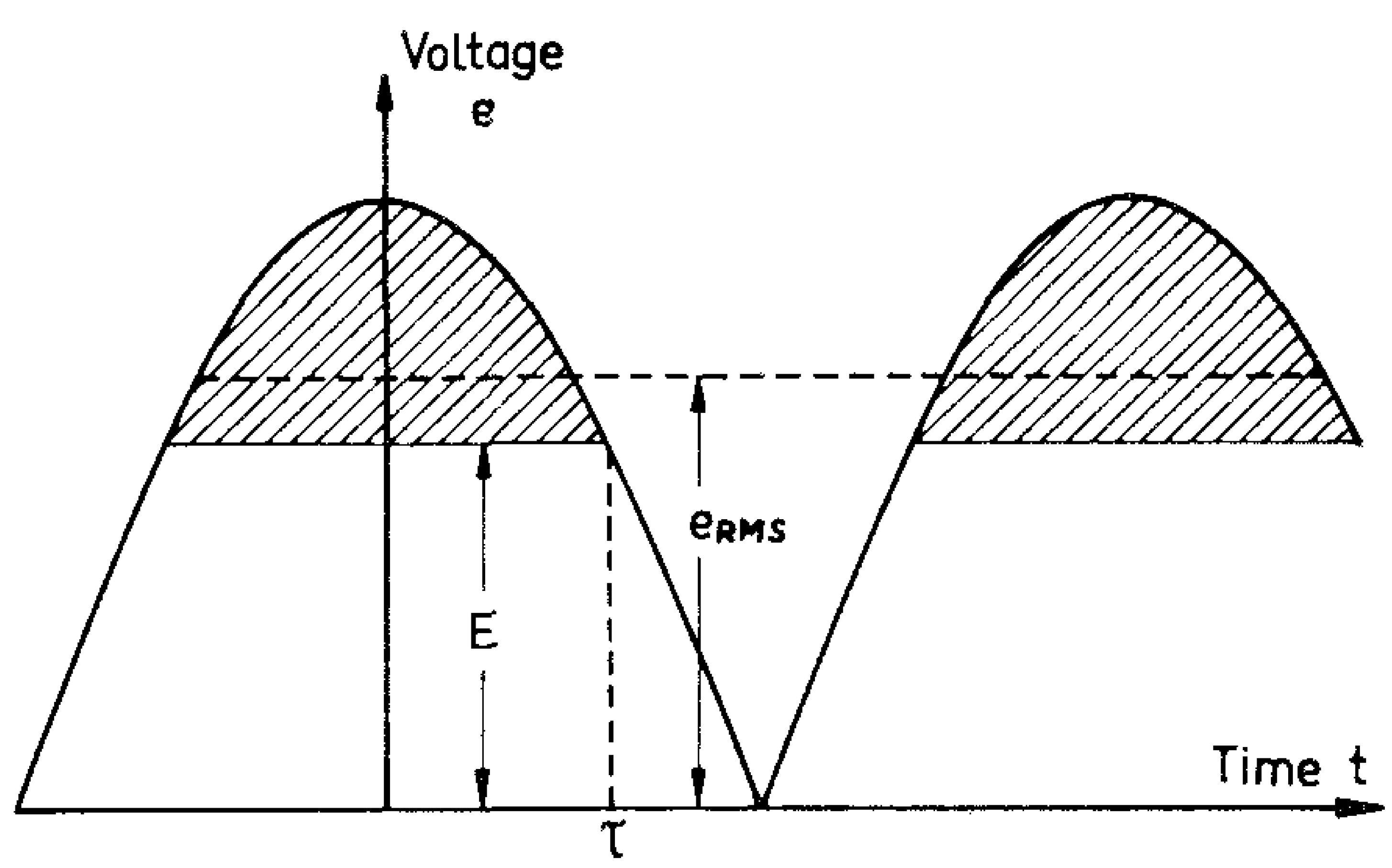


Fig. 8. Sine-wave with indication of the RMS value e_{RMS} and the average value of the capacitor DC voltage E.

the shaded areas will be supplied to the capacitor during each cycle. This must then equal the reduction in charge on the capacitor per cycle, whereby the current through the meter can be found.

$$\int_0^T I\,dt = \frac{E}{R}\,T = \frac{2\pi}{\omega}\int_0^T i\,dt = \int_0^\tau \frac{e-E}{r}\,dt$$

where: $e = e_{RMS}\sqrt{2}\cos(\omega t)$ and $\cos(\omega\tau) = \frac{E}{e_{RMS}\sqrt{2}}$

When r, E and e_{RMS} are given it is possible to find R.

In the same way the curves for square-waves and rectangular pulses with different crest factors can be found.

2. Random Noise.

In the case of random noise another way of attacking the problem must be used because the wave-form is not known, but only the probability density function

$$P(e) = \frac{1}{\sigma\sqrt{2\pi}} \times \exp\left(\div\frac{e^2}{2\sigma^2}\right)$$

where $\sigma = e_{RMS}$ (see fig. 9).

Here

$$I = \frac{E}{R} = 2 \int_0^{\infty} \frac{(e-E)}{r} \times P(e)\,de$$

The integration can be carried out by means of mathematical tables of the Gaussian error function and its integral.

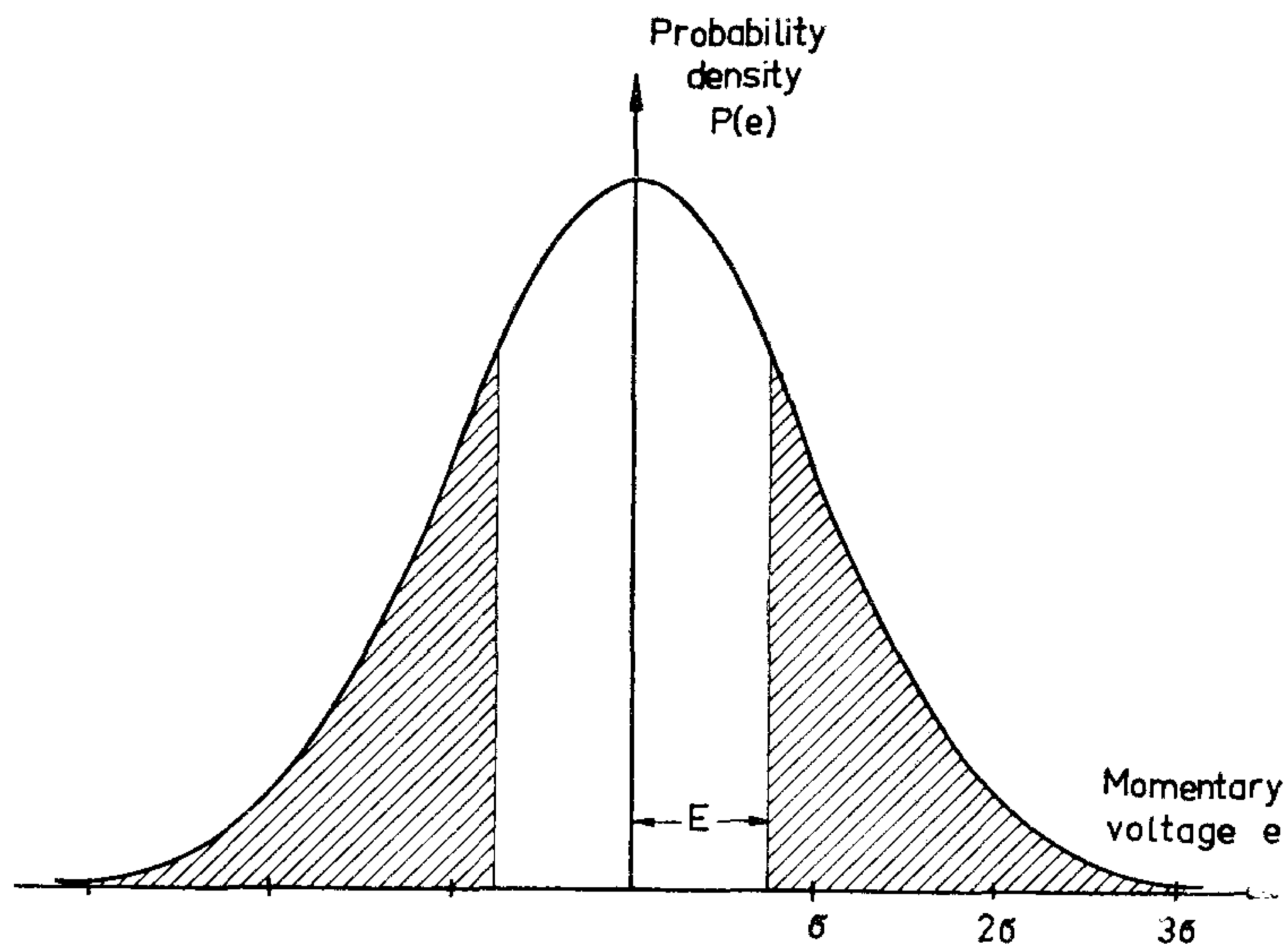


Fig. 9. Curve showing the probability density of random noise.

By looking at fig. 7 it is seen that the different curves intersect in different points. For example the curves for sine-waves and random noise intersect for the ratio $\frac{R}{r} = 4$. This means that an instrument having an $\frac{R}{r}$ ratio of 4, and which is adjusted to indicate correctly the RMS value of a sine-wave signal will also indicate correctly the RMS value of random noise.
Because the sine-wave and random noise are two rather important types of signals it would seem convenient to use a rectifier circuit with ratio $\frac{R}{r} = 4$ in practical instruments. However, this circuit will give a 1.5 db too high reading when rectangular pulses with a crest factor of 2 are measured and approx. 2 db too low for rectangular pulses with a crest factor of 5. At higher crest factors the error will be even greater, but such signals are not met very often in practice except in the special pulse technique. In this case, however, one is normally more interested in the peak value and the wave-form than in the RMS value of the signal.
It is now of interest to see if the simple circuit shown in fig. 5 can be modified to increase the measuring accuracy. From fig. 10 can be seen that a much better approximation to the parabola is obtained by means of a broken line and fig. 11 shows a circuit which can produce this type of characteristic. The diodes in the full wave rectifier causes the first "knee"

of the curve, diode A the second "knee", B the third one and so on. The advantages of the "displacement" of the characteristic with the DC voltage

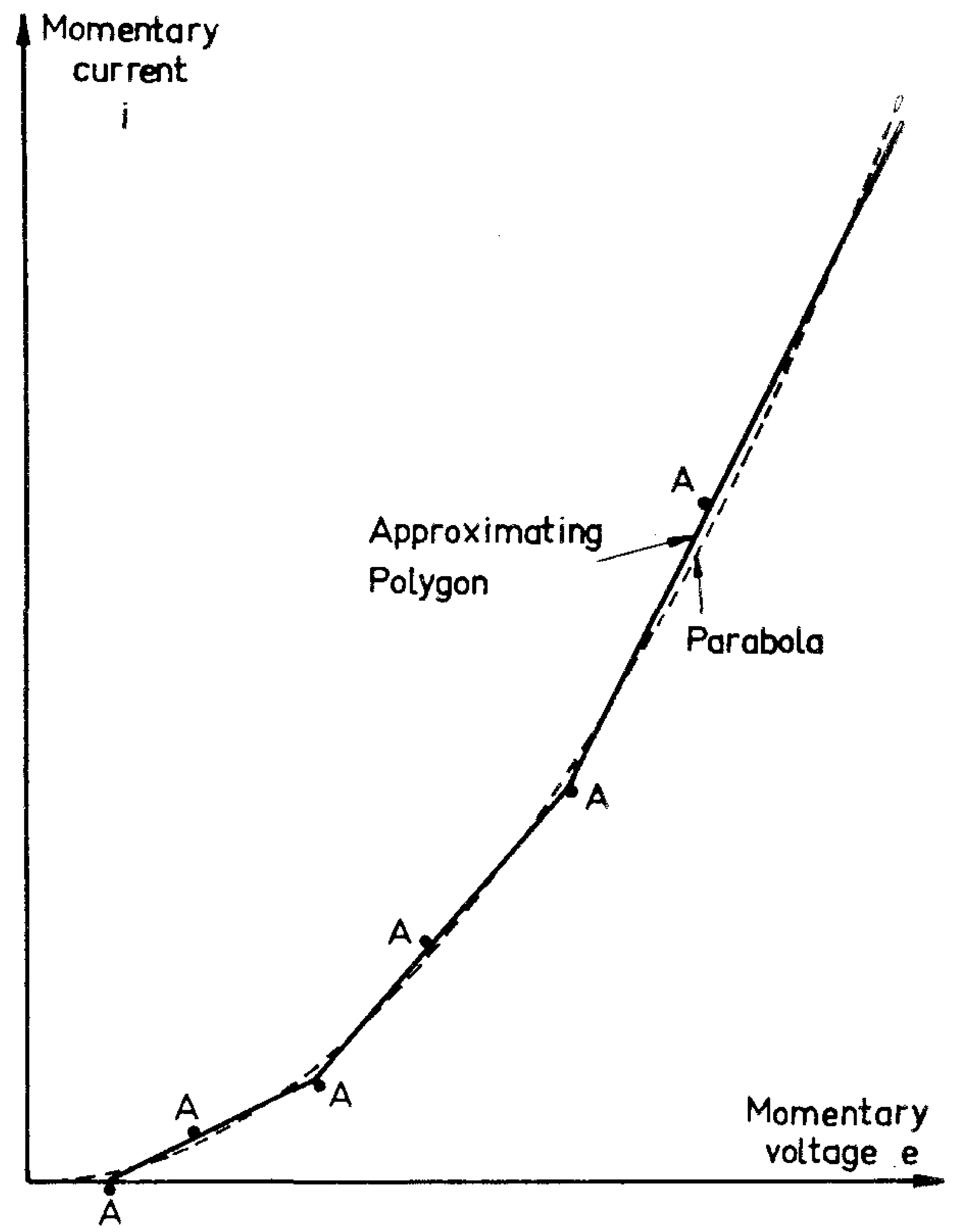


Fig. 10. Current voltage characteristic of a RMS rectifier circuit approximated by a polygon.

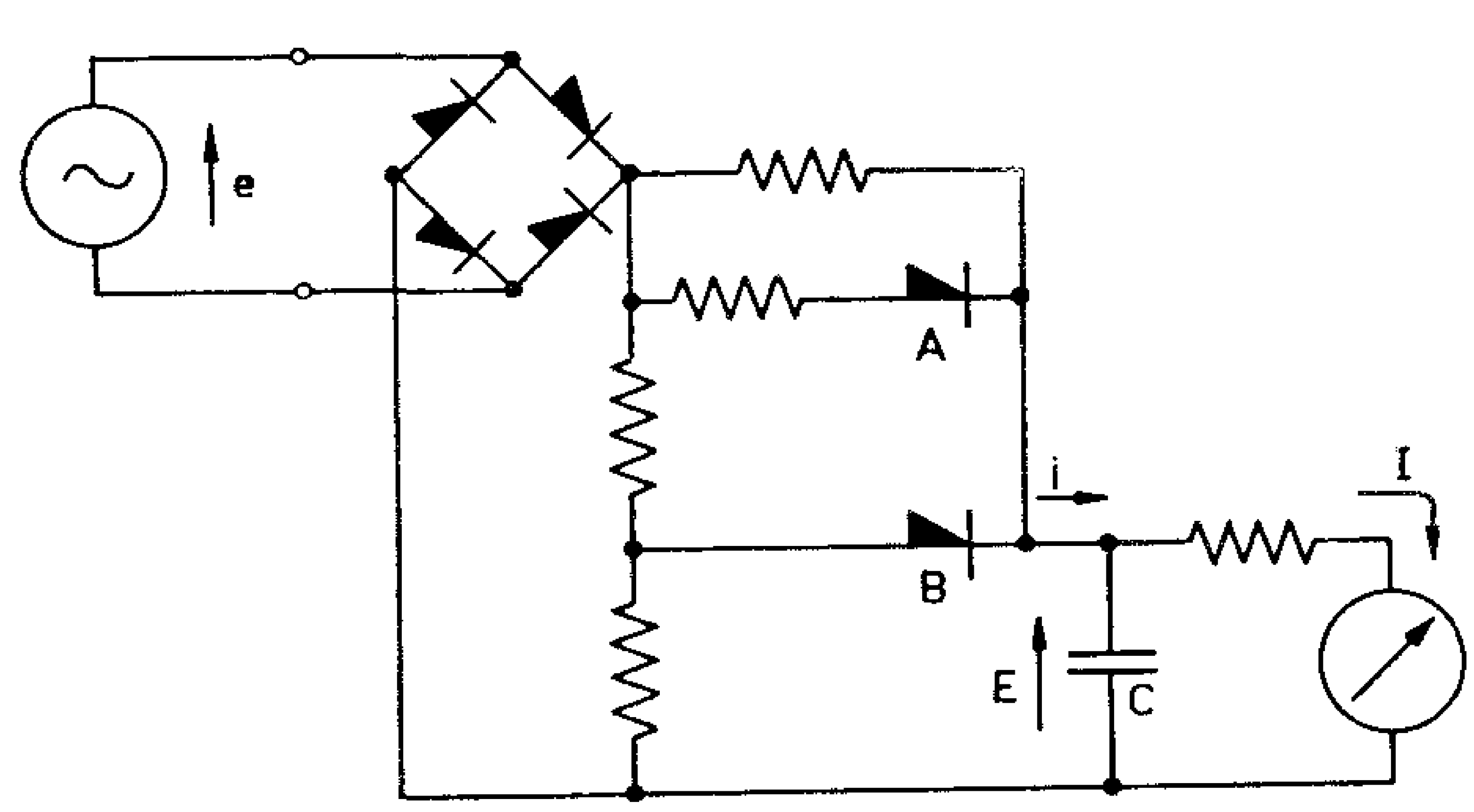
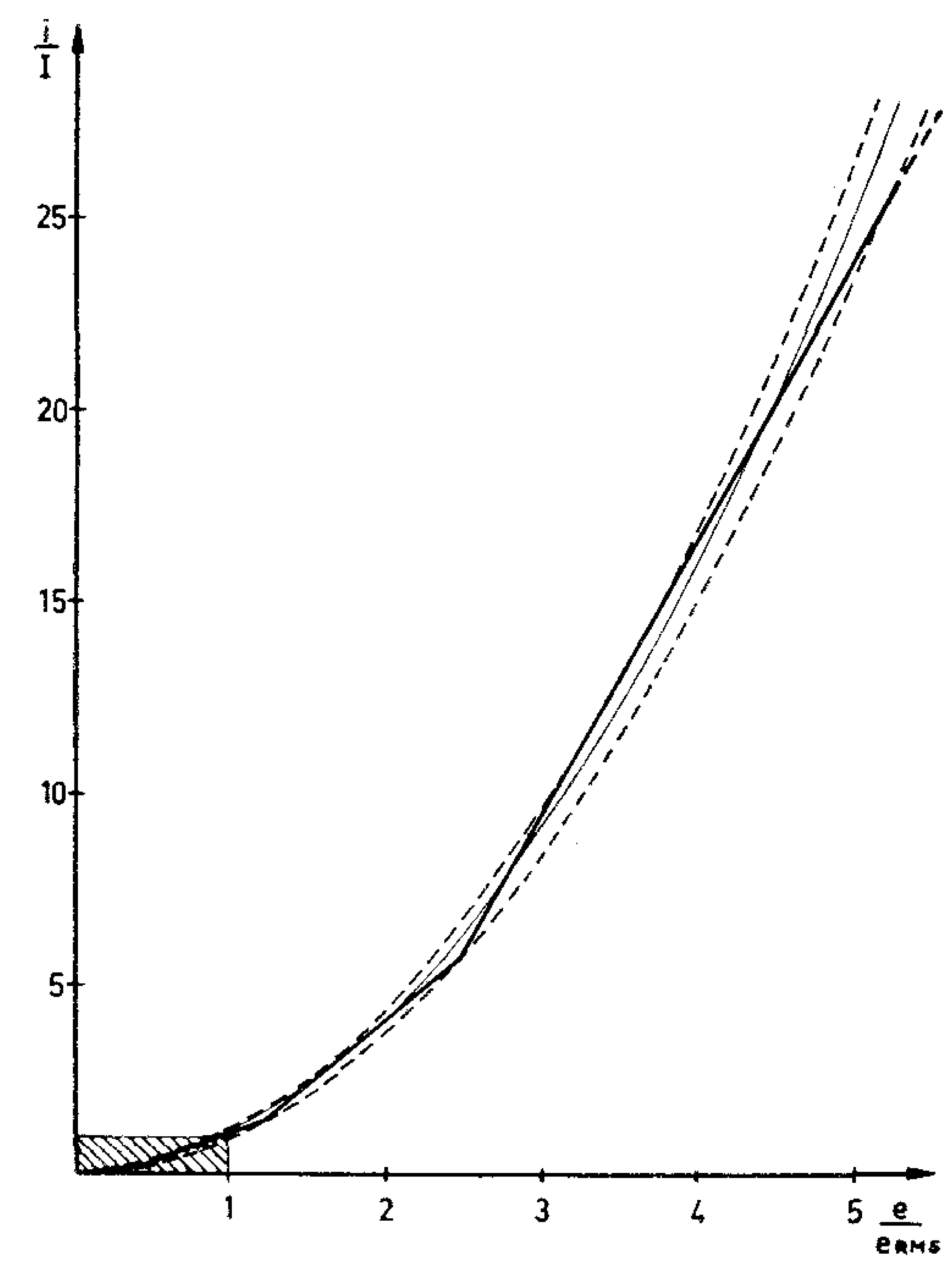


Fig. 11. Schematic diagram of the RMS rectifier circuit the characteristic of which is seen in fig. 10.

on the capacitor, as described previously for the simple circuit, are obtained also in this case, that is, the meter scale will be almost linear, and the relative error for a certain wave-form will be almost constant all over the scale. Compared with the similar squaring circuits using fixed bias this circuit therefore gives a higher accuracy with a smaller number of diodes.
To determine how many straight line portions are necessary to approximate the parabola with a certain accuracy the method outlined for the simple

circuit cannot be used because too many variables are involved. It is therefore necessary to seek other design criteria. From fig. 10 is seen that the greatest error is obtained for rectangular pulses having a certain crest factor so that they hit one of the points marked "A" on the parabola. It is therefore useful to draw the parabola together with two tolerance curves, and thereafter draw the polygon such that it lies between these curves, as seen from figs. 12 and 13. In fig. 12 is shown the full characteristic with tolerance curves corresponding to a certain percentage error, while fig. 13 shows a



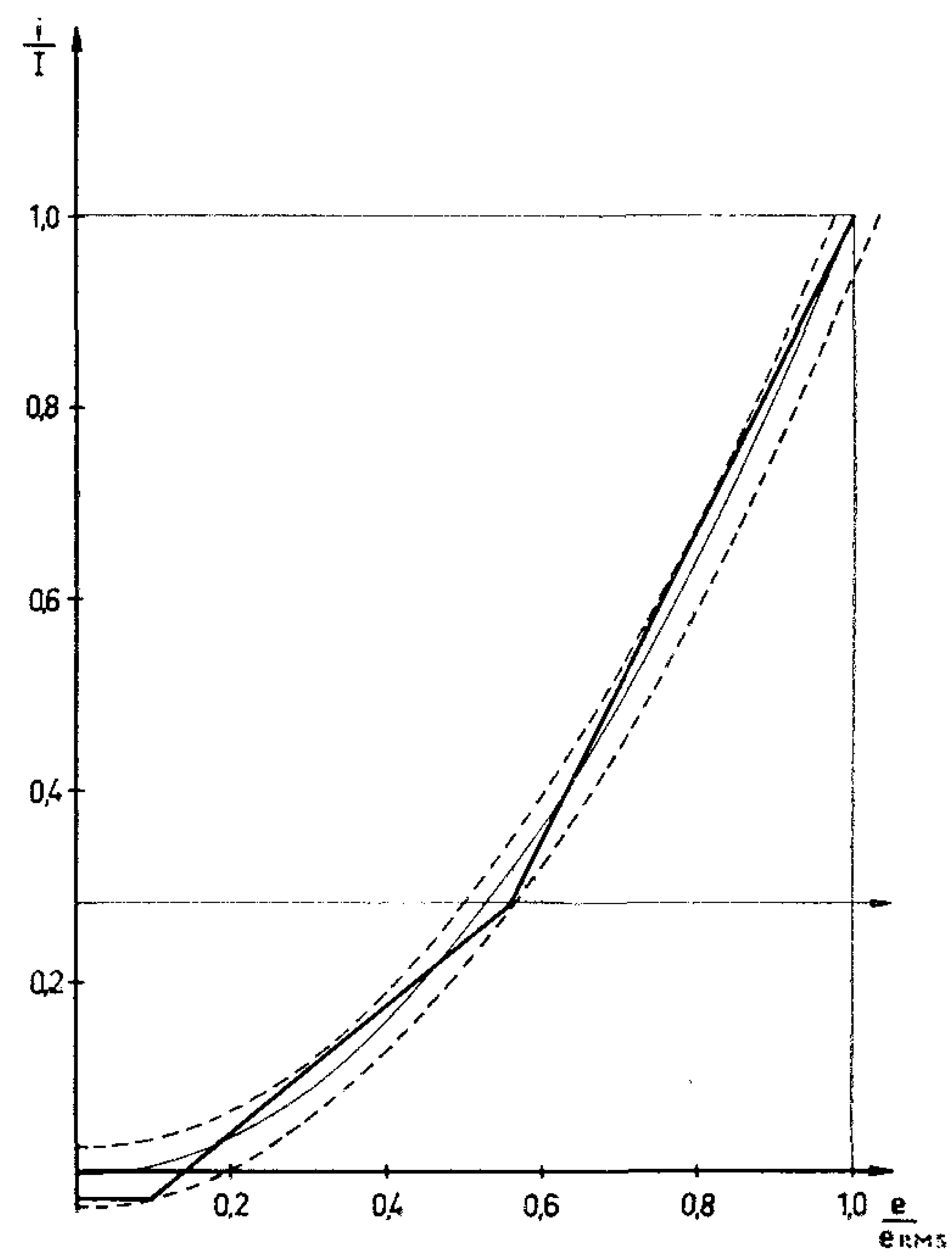


Fig. 12. Parabola with tolerance curves and approximating polygon for a RMS rectifier circuit.

Fig. 13. Close-up view of the inner part of the curves in fig. 12.

blown up view of the "inner part" of the characteristic corresponding to the shaded area in fig. 12. Because crest factors smaller than 1 do not exist, a signal wave-form other than the rectangular pulsewave must be used to determine the greatest possible error from this part of the polygon. In fig. 14 a wave-form of the desired type is shown (after rectification in a full wave rectifier). Assuming that the peak of the signal is correctly measured, that is, the peak value reaches one of the inter-section points of the polygon and the parabola, and that the signal itself has a crest factor of 4—5, the peak must be relatively high and thin. The lower part of the wave will therefore practically endure the whole period and the error which depends only on this "part" of the wave, can be found directly by means of fig. 13. The error in the momentary current should, however, be compared to the total DC, that is the current I. The tolerance curves must therefore in this case show a constant absolute deviation from the parabola instead of a constant relative deviation. Because also the peak may be measured (fig. 14) incorrectly the total error can, in the worst case, be equal to the sum of the two errors. The tolerance curve shown corresponds to a max. error of 4 % in fig. 12

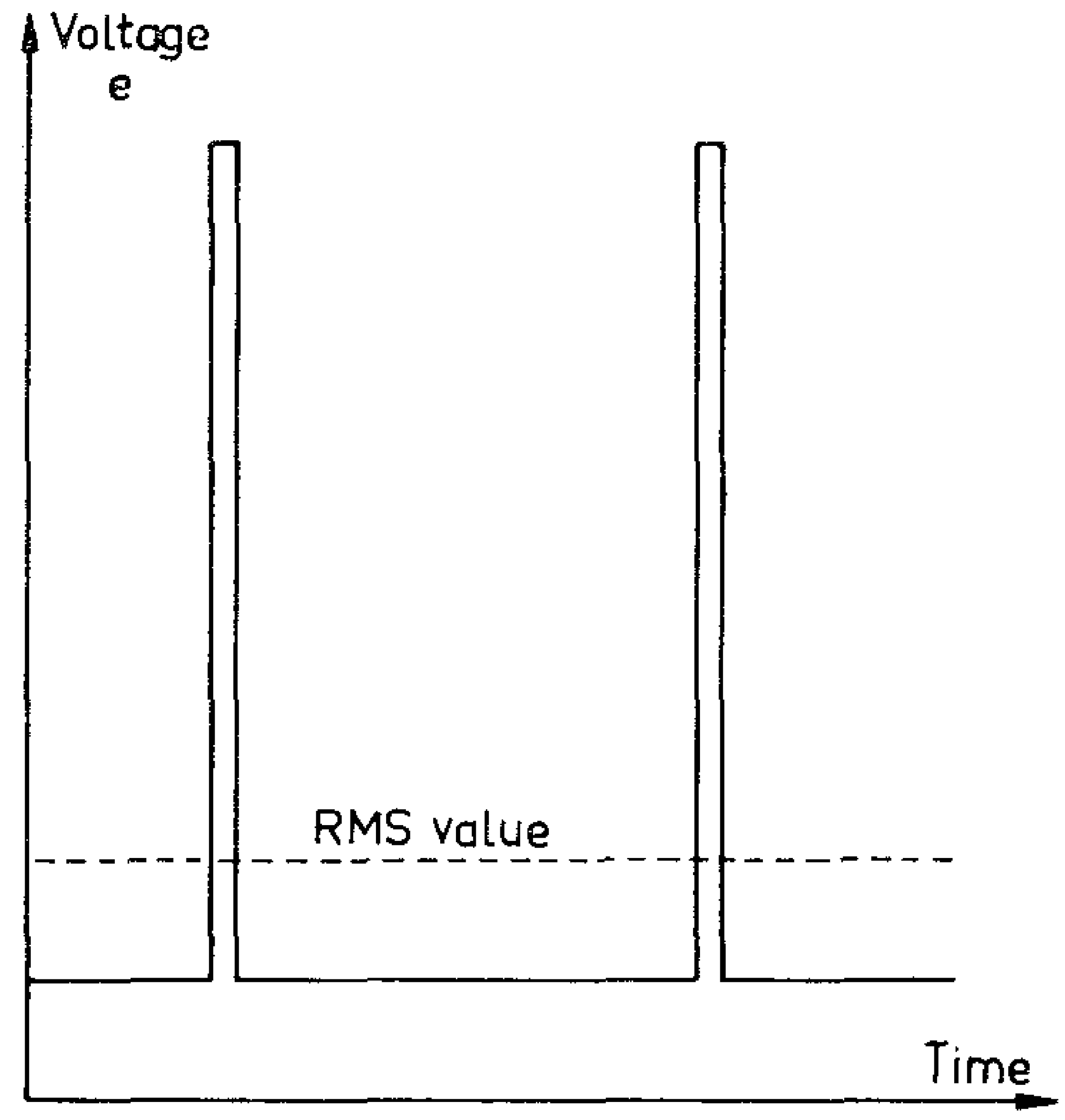


Fig. 14. Full wave rectified square pulse signal.

and 2 % in fig. 13, i.e. the max. total error may amount to 6 % (o.5 db). However, the wave-forms normally met with in practice will "cover" the polygon in such a way that the positive and negative errors compensate for each other to a certain degree and the RMS value will therefore normally be measured with a greater accuracy, possibly within ± 2 %.

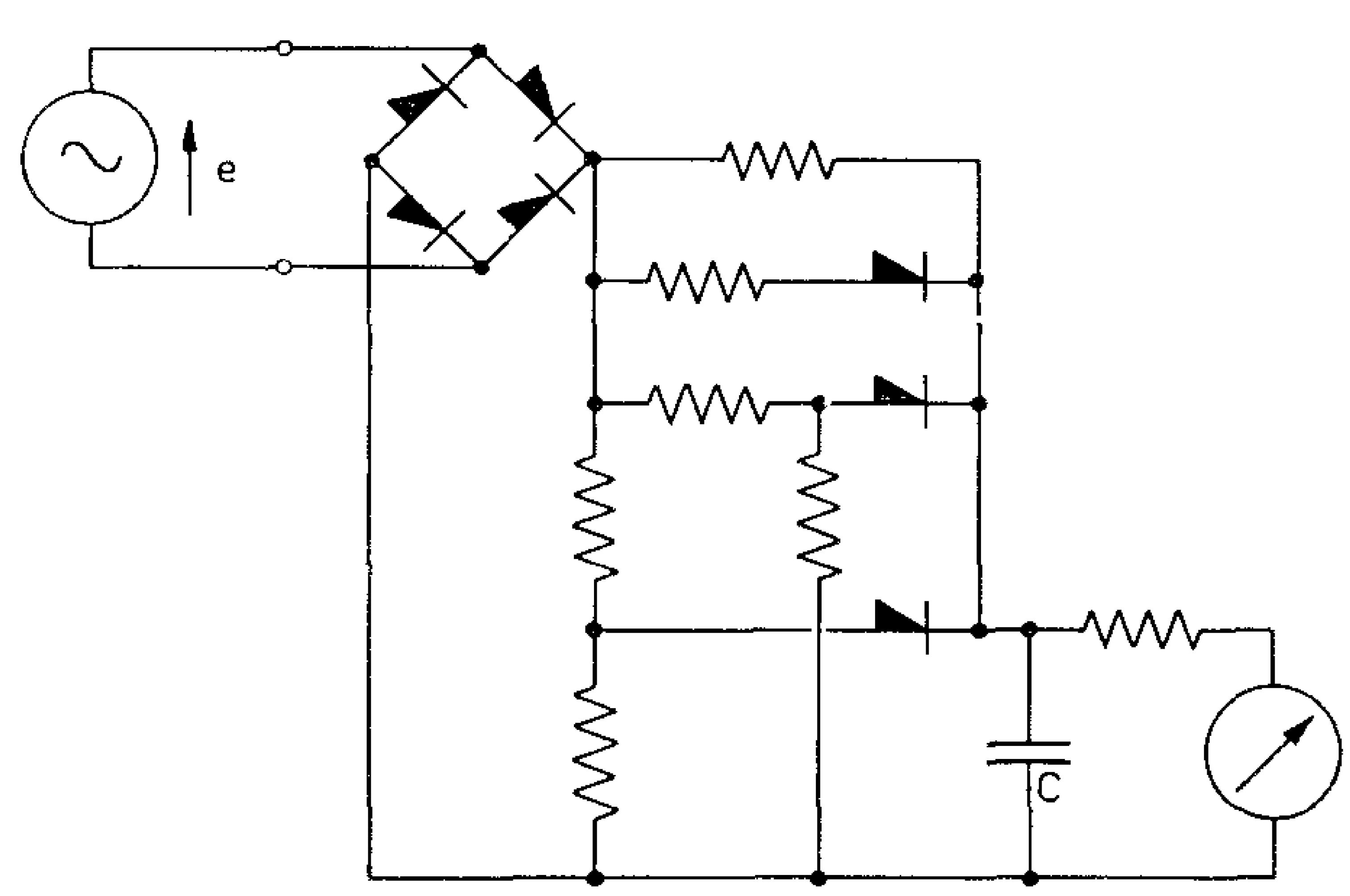


Fig. 15. Schematic diagram of the RMS rectifier circuit the characteristic of which is seen in fig. 12 and 13.

In fig. 15 a circuit diagram is shown which will give the above described broken line characteristic with 4 "knees". The approx. ratio between the different resistors can be found from figs. 12 and 13. It is necessary, however, to correct for the non-linear forward characteristic of the diodes. It should be mentioned that the characteristic of the rectifier network intersects with the voltage (abscissa) axis at the second bend due to the DC bias on the capacitor. This means, that the capacitor does not discharge only through the meter. The meter current will consequently be too small, which is illustrated by means of the extra axis shown in fig. 13. The displacement of the curve in vertical direction has no influence whatsoever on the RMS rectification. A displacement in horizontal direction, however, would influence the type of rectfication obtained because a voltage component proportional to the average value of the input signal would be added to the voltage across the capacitor.

It is of course possible to build still more accurate RMS rectifiers by using a greater number of straight lines to approximate the parabola. Other sources of error, however, will soon set a limit to how far it pays to go. The limited accuracy of the resistors being used causes deviations from the desired polygon and the non-linear forward characteristic of the diodes causes the polygon to be correct for a certain voltage (meter deflection) only. The latter source of error can be reduced by designing the circuit for great voltages and using high-ohmic resistors. If the value of the resistors are too great, however, the backward current of the diodes will influence the measurements especially at high temperatures. The voltage limit will normally be set by the amplifier feeding the rectifier circuit. Another improvement of the circuit would be to make it possible to measure wave-forms with higher crest factors correctly. This requires that the amplifier either must be able to supply a greater momentary voltage, and what is more important, a greater momentary current because the momentary current varies with the square of the momentary voltage, or a more sensitive meter must be used. The latter solution causes, however, the influence of the backward current of the diodes to increase. Furthermore, it becomes difficult to satisfy the dynamic requirements of the meter, which must be satisfied especially when sound measurements are taken, according to the standard ASA Z24.3 for objective sound level meters. As a practical compromise between all these factors the meter chosen for the Audio Frequency Spectrometer Type 2110 and the Microphone Amplifier Type 2603 has a current consumption of max. o.2 mA. This instrument satisfies the dynamic requirements of the standards. The input voltage is set to approx. 10 volts RMS for full scale deflection, and the max. crest factor is 5 which means that the amplifier must be able to supply a peak voltage of $\pm$ 50 volts and a peak current of $\pm$ 10 mA. Because the amplifier in these instruments can supply only approx. $\pm$ 8 mA correct measurements at full scale deflections are only obtained for waveforms with crest factors up to 4. At 4/5 full scale deflection and less, the crest factor may be as great as 5.

The lower frequency limit of the rectifier circuit is determined by the filter capacitor C and the resistors. The upper frequency limit will be set by the internal capacity of the diodes and their transient switching time which will cause phase shift and distortion in the partial currents. Finally it should be mentioned that the meter scale for the RMS rectifier will show just the same small curvature which is obtained in the case of an arithmetic average and a peak value rectifier circuit in which the same diodes are used for the full wave rectification plus one more series connected diode corresponding to the diodes causing the three upper knees of the RMS rectifier characteristic.

APPENDIX

In this appendix a brief review is given of the definitions of the different quantities used in the preceeding article.

The following matematical definitions all refer to an electrical signal e(t), and an integration time $T = t_2 - t_1$.

Peak Value:

$e_{peak} = e_{max}(t)$ is the maximum value of e(t) within the time interval T.

Average Value:

$$e_{average} = \frac{1}{T}\int_{t_1}^{t_2} |e(t)|\, dt$$

RMS Value:

$$e_{RMS} = \sqrt{\frac{1}{T}\int_{t_1}^{t_2} [e(t)]^2\, dt}$$

From these basic definitions are derived:

The crest factor:

$$F_c = \frac{e_{peak}}{e_{RMS}} \quad \text{and}$$

the form factor:

$$F_f = \frac{e_{RMS}}{e_{average}}$$

For a sinusoidal function $e(t) = E \sin \omega t$, the different values will be:

$$e_{peak} = E$$

$$e_{average} = \frac{2E}{\pi}$$

$$e_{RMS} = \frac{E}{\sqrt{2}}$$

$$F_c = \sqrt{2}$$

$$F_f = 1.11$$

In fig. 16 is shown the geometrical meaning of the matematical definitions. For the sake of convenience a periodically repeated function with the periode T is considered. Fig. 16a shows the peak value, and fig. 16b the average value. This is determined in such a way that the hatched area equals the total area between the e(t)-curve from t_1 to t_2 and the axis of abscissae. Finally the squared RMS value as indicated in fig. 16c is seen to be the average value of the squared signal e(t). The RMS value is naturally the square root of the one indicated on the figure.

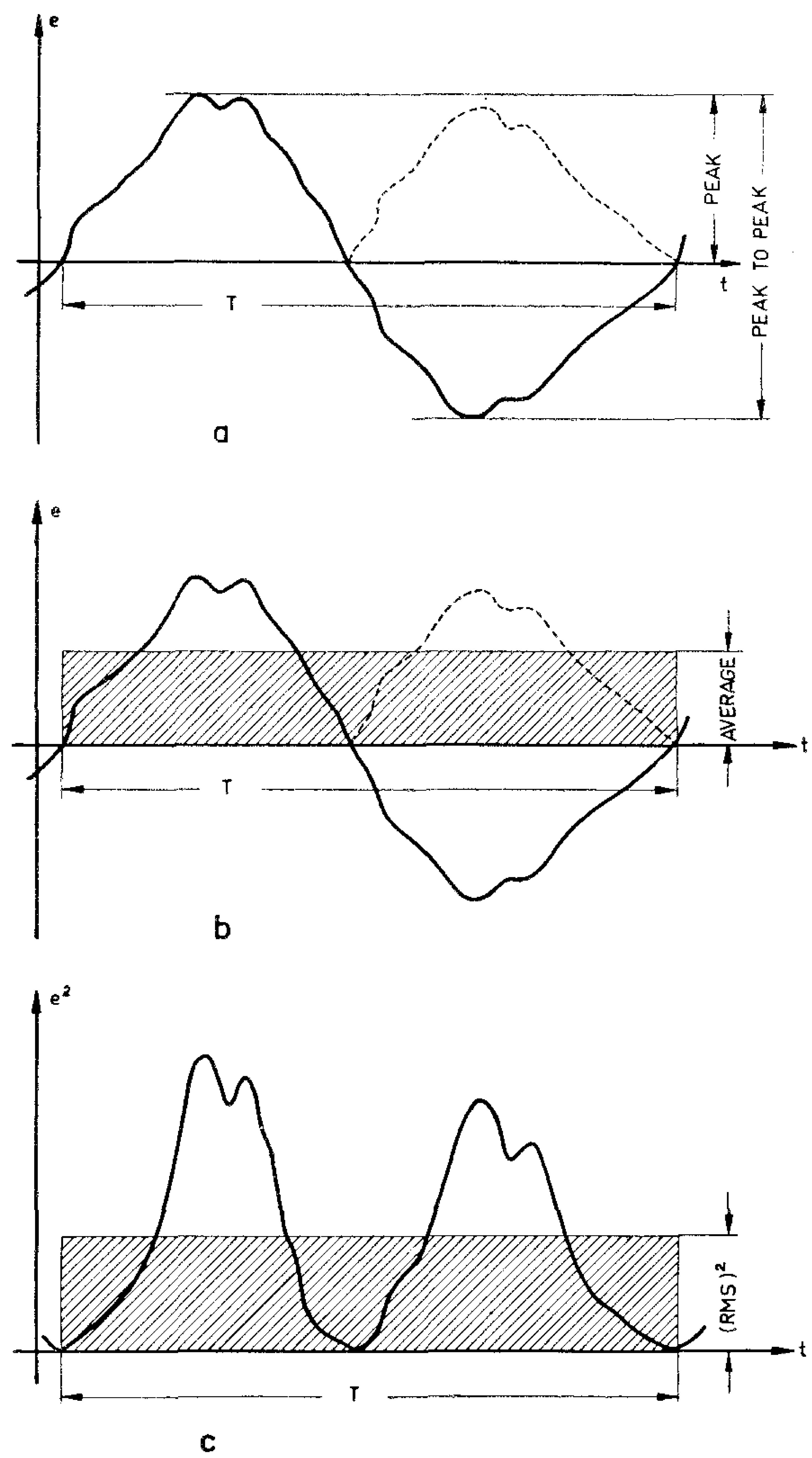


Fig. 16. a. Peak value of a periodical signal.
b. Average value of a periodical signal.
c. Squared RMS value of a periodical signal.

News from the Factory.

1/3 Octave Filter Set Type 1610.

The 1/3 Octave Filter Set Type 1610 is a redesign of the previous model Type 1609. It consists of thirty 1/3-octave filters with center frequencies from 40 c/s to 32000 c/s and the four internationally standardized weighting networks for sound level measurements. It is, furthermore, supplied with a 7-poled socket for connection to the Extension Filter Set Type 1619. The filters can be manually selected by means of the 50 position switch or automatically scanned when coupled to the Level Recorder Type 2304.

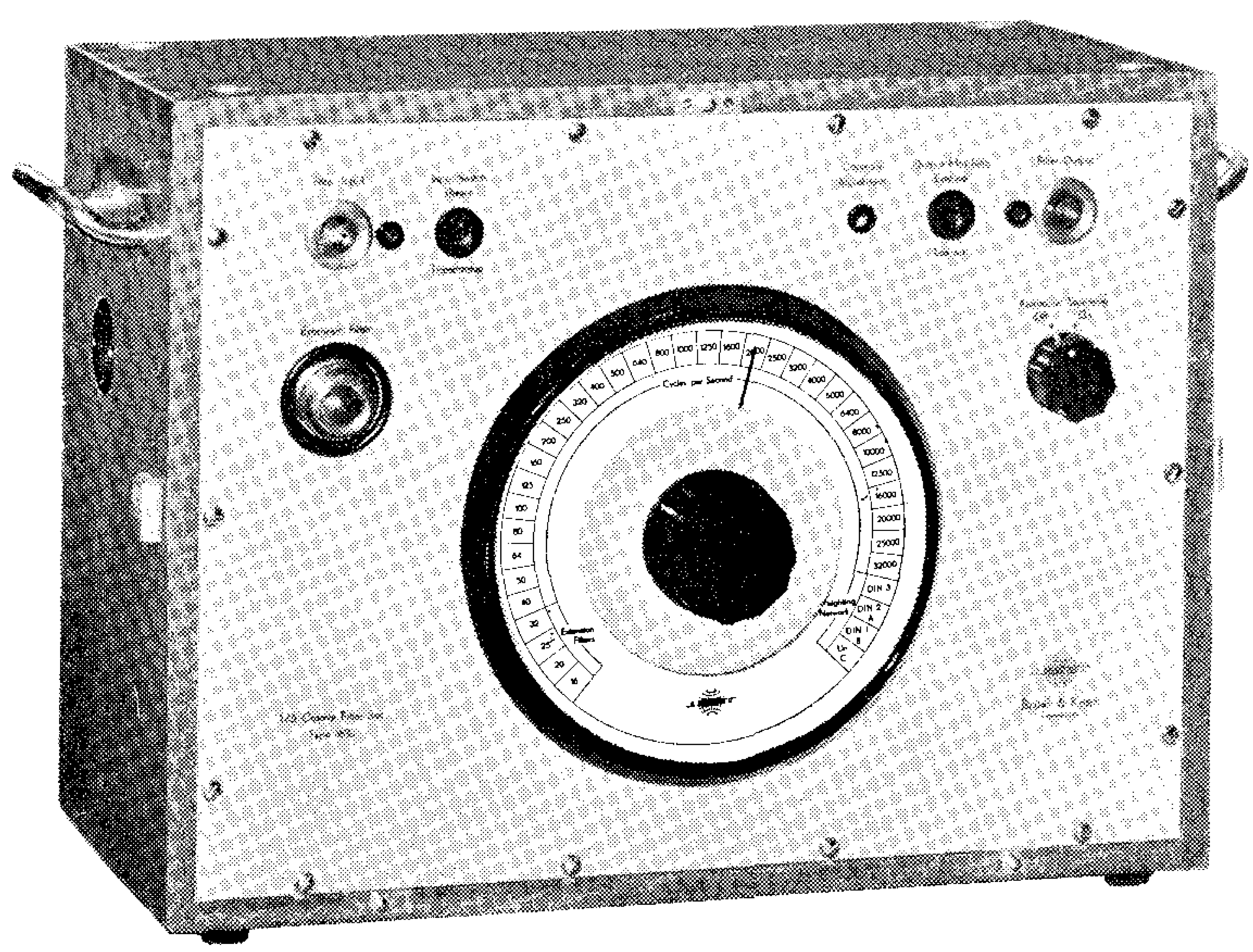


Fig. 1. Photo of the 1/3 Octave Filter Set Type 1610.

Type 1610 is delivered in lacquered mahogany cabinet with handles and lid.

Extension Filter Set Type 1619.

The Extension Filter Set Type 1619 consists of four 1/3-octave filters with center frequencies of 16 — 20 — 25 and 31.5 c/s. Type 1619 is supplied with a 7-cored screened cable with plug for connection to the Spectrometer Type 2110 and the Filter Set Type 1610.
Type 1619 is delivered in a lacquered mahogany cabinet with handles and lid. The Extension Filter Set can also be delivered *without mahogany cabinet* called *ZS 0145* which is similar to the previous model ZS 0045. These units

can be built into the rack-mounted models of the Spectrometer. See table below.

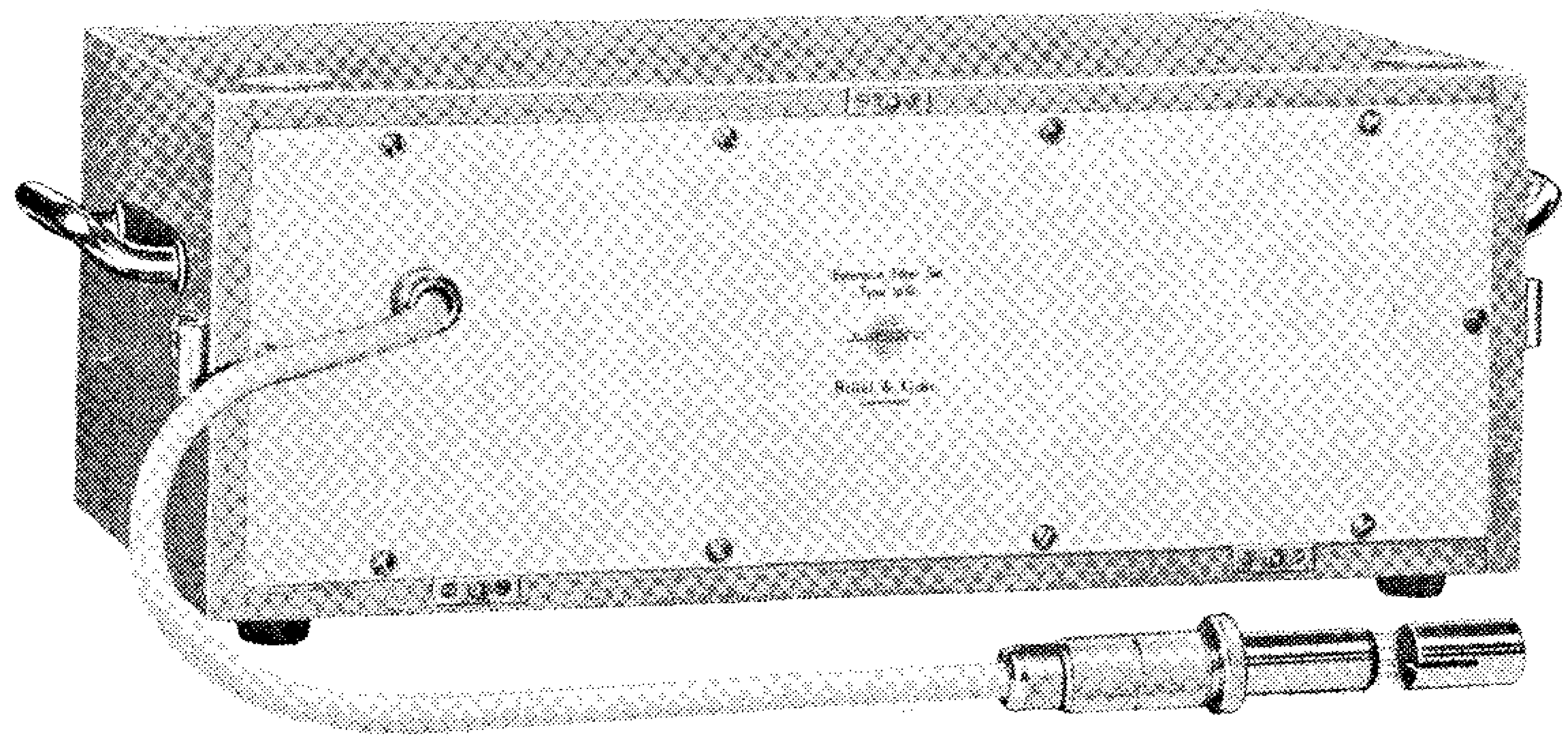


Fig. 2. Photo of the Extension Filter Set Type 1619.

	Instrument Type to be used in connection with Extension Filters							
Extension Filter	1609	1610	2109	2110	2311	3310	3320	3321
1619	—	+	—	+	—			—
ZS 0145	—		—		—	+	+	—
ZS 0045					+			+

— = not usable combination
\+ = usable combination
= may be used, not practical.

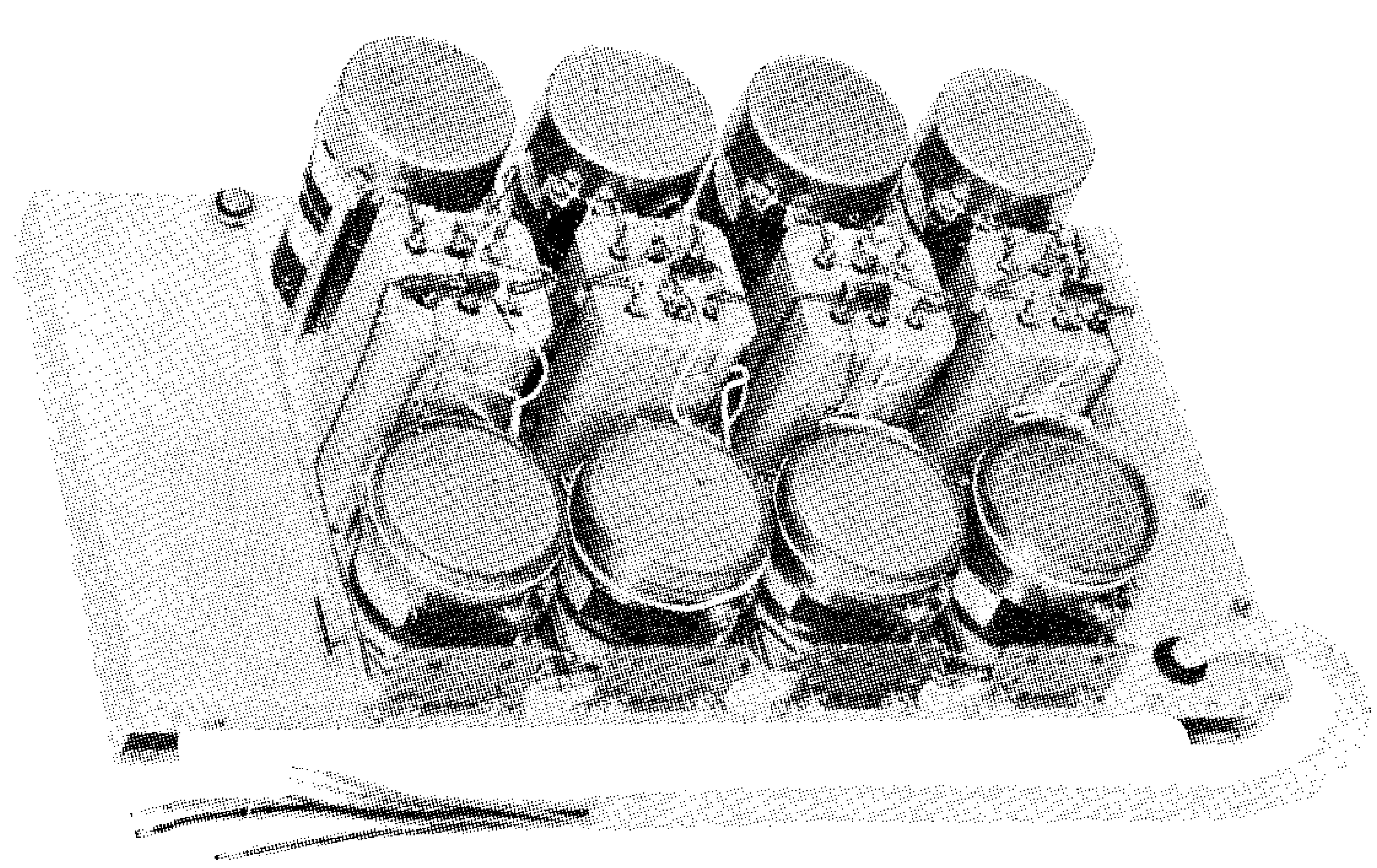


Fig. 3. Photo of the Extension Filter Set ZS 0145.

Audio Frequency Spectrometer Type 2110.

The Audio Frequency Spectrometer Type 2110 is a redesign of Type 2109. For further information is referred to page 2.

Modification of Frequency Analyzer Type 2105.

To make the Frequency Analyzer fit to other Condenser Microphones than the Brüel & Kjær Condenser Microphone Type 4111, the instrument is modified to include a polarization voltage which is continuously variable between 150 and 250 V. The previous model was only supplied with a fixed 150 V polarization voltage.

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