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Maddox et al.

[11] **3,868,857**

[45] **Mar. 4, 1975**

[54] **AUDIO DOSIMETER**

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[73] Assignee: **Teledyne, Inc.,** Los Angeles, Calif.

[22] Filed: **Nov. 16, 1972**

[21] Appl. No.: **307,082**

[52] **U.S. Cl.** **73/558**

[51] **Int. Cl.** **G01h 5/00**

[58] **Field of Search** .... 181/.5 NP, .5 AP; 179/1 N

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Assistant Examiner—J. V. Doramus

[57] **ABSTRACT**

An audio dosimeter for individual use determining exposure to sound energy as a function of both frequency and pressure level, with integration over the time of exposure and incorporating storage means preserving a quantitative measure of the exposure.

6 Claims, 8 Drawing Figures

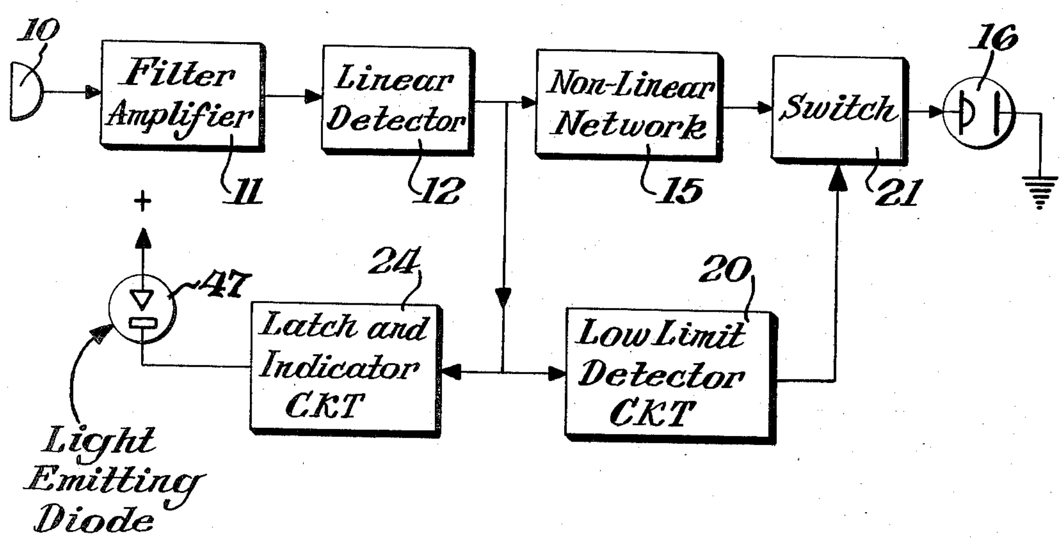
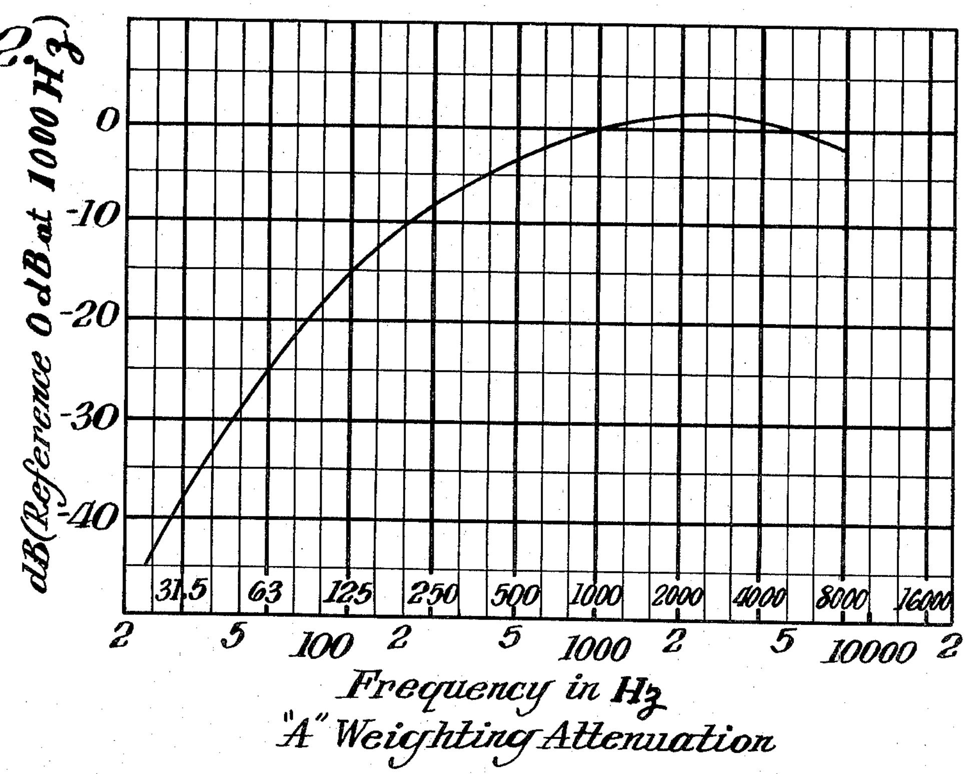


Fig. 2.



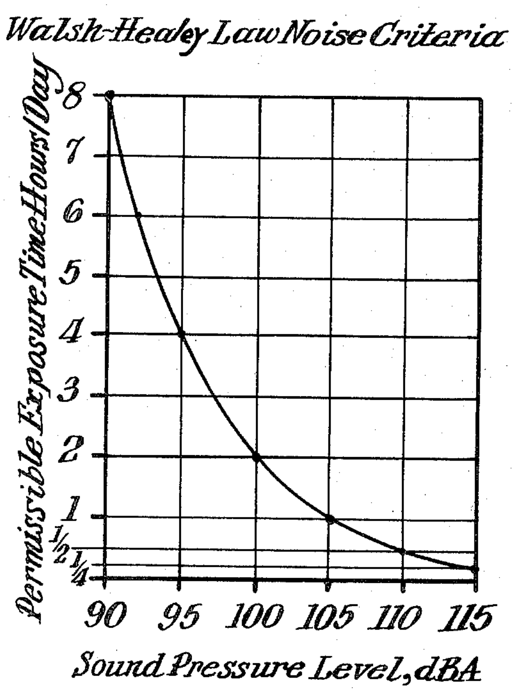


Fig. 1.

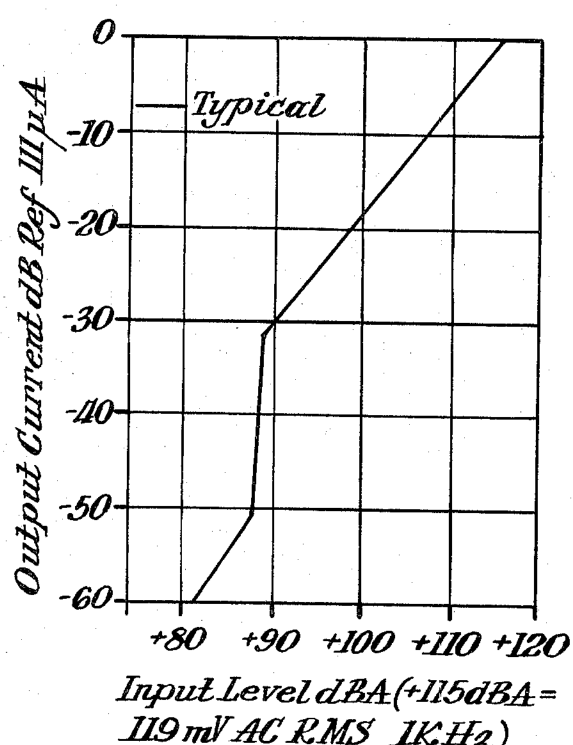


Fig. 3.

Fig. 4A.

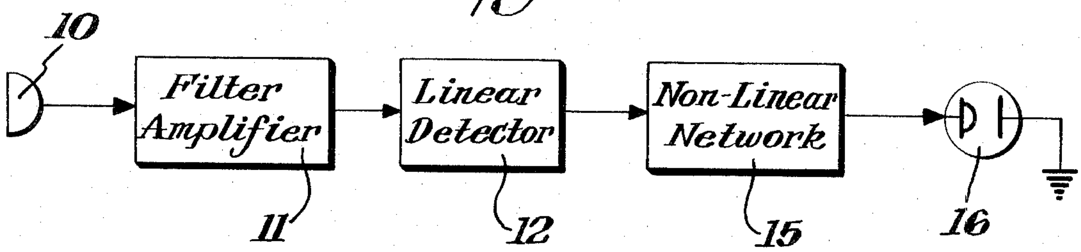


Fig. 4B.

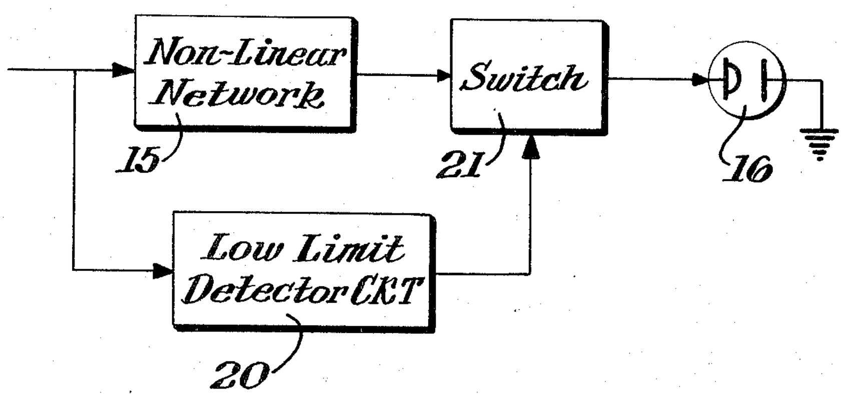


Fig. 4C.

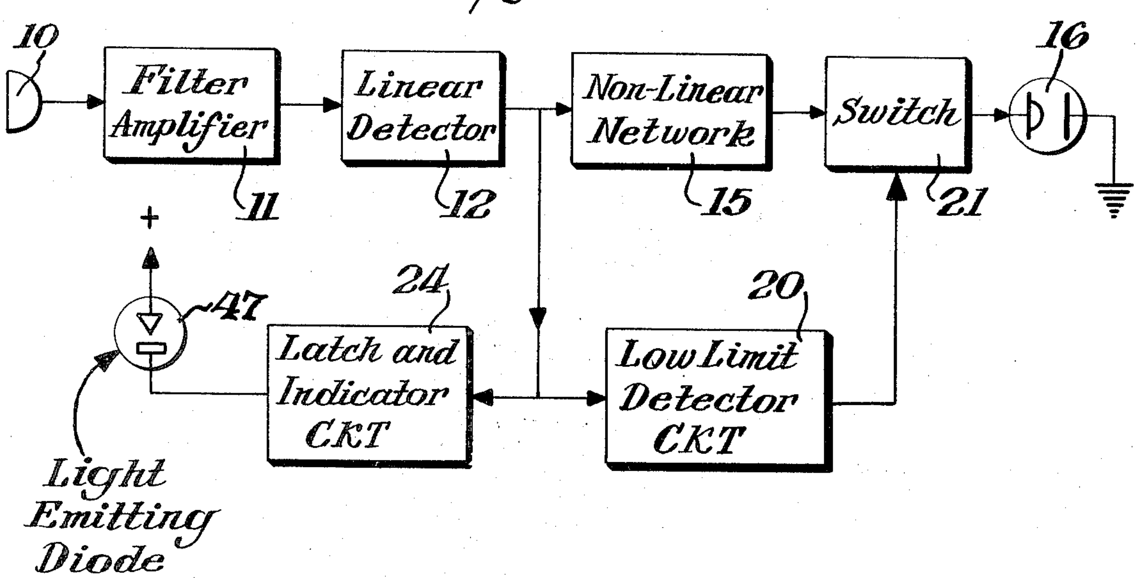


Fig. 5A.

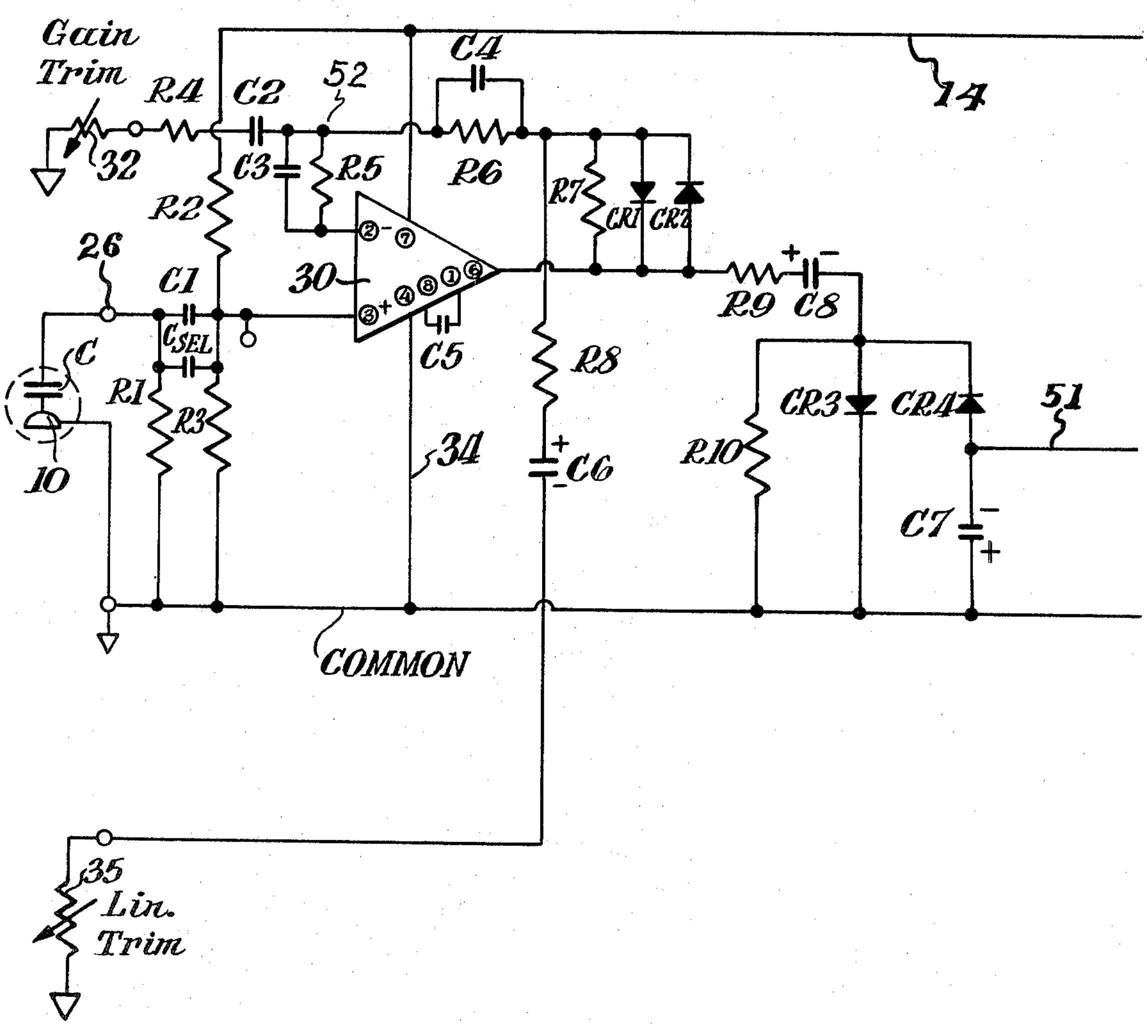
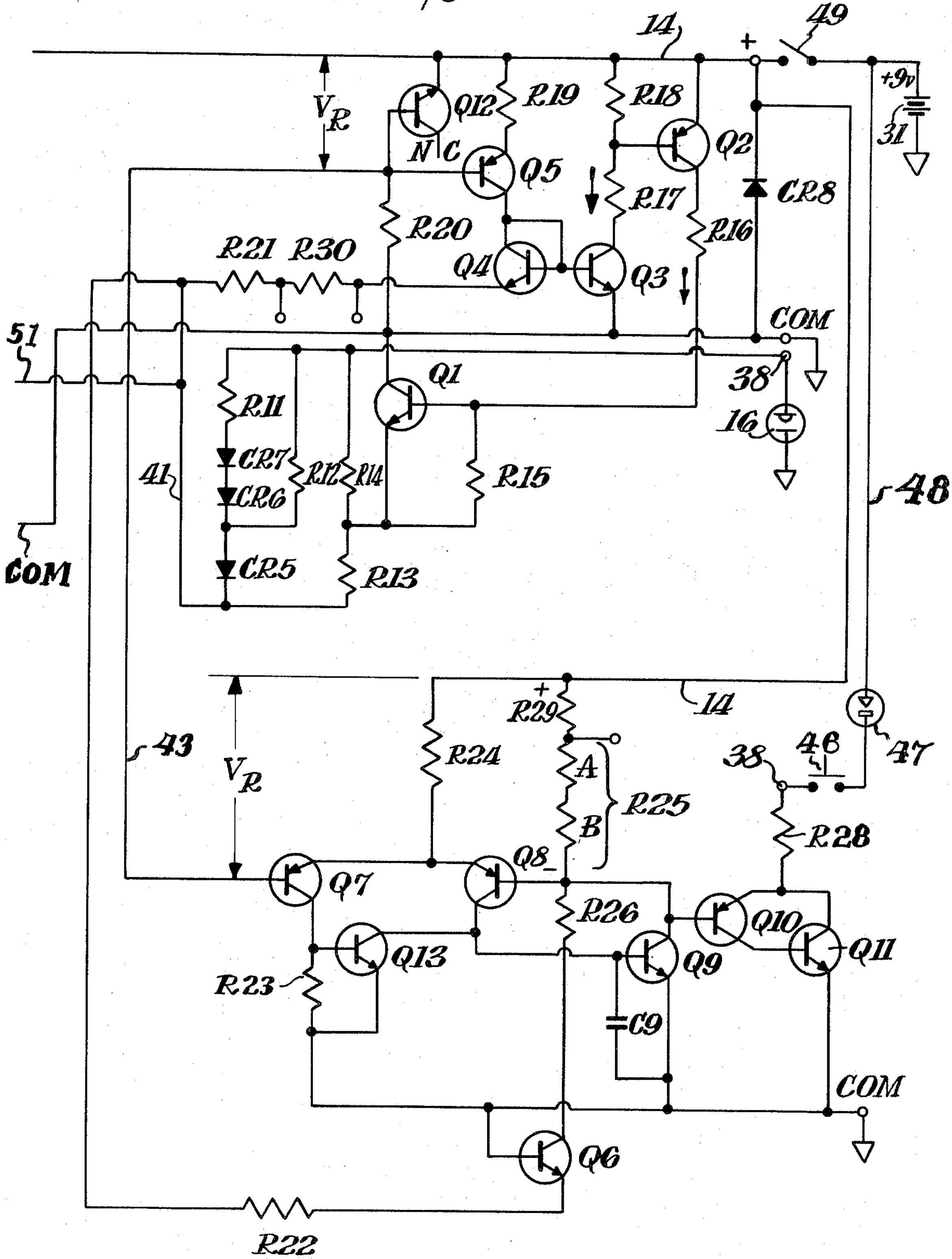

Gain
Trim
R4
C2
52
C4
14
32
C3
R5
R6
R7
CR1
CR2
R2
26
C1
30
C5
R9
C8
C
C_{SEL}
R8
CR3
CR4
51
R1
R3
R10
34
C6
10
C7
COMMON
35
Lin.
Trim

Fig. 5B.


14
49
+
+9v
31
VR
Q12
R19
R18
Q2
N
C
Q5
CR8
R17
R20
R16
R21
R30
Q4
Q3
COM
51
38
Q1
16
R11
CR7
R12
R14
R15
48
41
CR6
COM
CR5
R13
14
47
R29
43
R24
A
38
46
VR
R25
B
R28
Q8
Q7
R26
Q10
Q11
Q13
Q9
R23
C9
COM
Q6
R22

AUDIO DOSIMETER

BRIEF SUMMARY OF THE INVENTION

Generally, this invention comprises an audio dosimeter for individual use comprising, in series circuit in the order recited, electronic microphonic sound sensor means, a filter-amplifier receiving the a-c voltage output of the sound sensor means, said filter-amplifier comprising an operational amplifier connected in circuit with a plurality of a-c filter networks each having individual band pass characteristics preselected to collectively interact to shape the a-c voltage output during transmission by the filter-amplifier to conform to the pattern of a preselected weighting network incorporating in the a-c voltage output the otolaryngologically (and psychologically) harmful contribution of ambient sound frequency, linear detector means rectifying the a-c voltage output from the filter-amplifier, a non-linear network shaping the d-c output current from the linear detector to the function required to produce a substantially straight line of correct slope in a plot of decibels referred to a preselected current level versus sound energy input level in decibels, and an electrochemical integrating cell receiving the output current from the non-linear network measuring sound frequency in terms of sound pressurer with weighted frequency and time of exposure conjointly.

DRAWINGS

The following drawings, detail a preferred embodiment of the invention and the physical principles of operation:

FIG. 1 is a plot of the Walsh-Healey Law permissible human exposure time in hours/day v. sound pressure level in decibels "A" weighting network (i.e., dBA),

FIG. 2 is a graphic representation of "A" Weighting Attenuation in terms of decibels referred to 0 decibels at 1,000 Hz v. frequency in Hz (logarithmic scale),

FIG. 3 is Output (i.e., integrating) Current response in decibels referred to 111μ ampere v. Sound Input Level in dBA (+115 dBA = 119mv A.C. RMS 1kHz.) for apparatus constructed according to this invention,

FIG. 4A is a block diagram of a basic audio dosimeter according to this invention,

FIG. 4B is a block diagram of a low limit detection and measurement switch-off auxiliary adapted for use with the basic apparatus of FIG. 4A,

FIG. 4C is a block diagram of the basic apparatus of FIG. 4A, provided with the auxiliary of FIG. 4B and including, additionally, a latch and indicator circuit for high level detection and indication, and

FIGS. 5A and 5B, as to which the latter is an extension of the former, are detailed circuit diagrams of a preferred embodiment of this invention, the circuitry of FIG. 5A and the non-linear network and electrochemical integrating cell of FIG. 5B collectively constituting the basic audio dosimeter of this invention, whereas the remainder of FIG. 5B constitutes the low limit detection and measurement switch-off auxiliary and the latch and indicator circuit auxiliary for high level detection and indication.

GENERAL

The physiologically (and psychologically) injurious effects of sound energy have been appreciated for a long time; however, it has only been since the passage of the Walsh-Healey Law that quantitative limits have been set on human exposure. The official standards prescribed are set forth in the Federal Register, Vol. 35, No. 17 — Saturday, Jan. 24, 1970. These standards have now been embodied in American National Standards Institute (ANSI) standard Sl. 4-1971 (refer particularly Table 1 and FIG. 3, page 14).

The statutory Permissible (Human) Noise Exposures settled upon are (wherein "dBA" represents "A" Weighted Sound Pressure Level):

TABLE I

Duration Per day, hours	Sound Pressure Level, dBA slow response
8	90
6	92
4	95
3	97
2	100
1½	102
1	105
½	110
¼ or less	115

In explanation of Table I, the footnote (1) applicable thereto reads:

"When the daily noise exposure is composed of two or more periods of noise exposure at different levels, their combined effect should be considered, rather than the individual effect of each. If the sum of the following fractions: $C_1/T_1 + C_2/T_2 + \ldots . C_n/T_n$ exceeds unity, then, the mixed exposure should be considered to exceed the limit value. C_n indicates the total time of exposure at a specified noise level and T_n indicates the total time of exposure permitted at that level."

In addition, Section 50-204.10, "Occupational noise exposure," of the legislation requires that protection be provided to employees subjected to sound exceeding the limits of Table I, and that, in all cases where sound levels exceed the tabulated values, "a continuing effective hearing conservation program shall be administered."

The graphical relationship of permissible human exposure time in hours/day versus sound level in dBA set out in Table I is shown in FIG. 1.

From the foregoing, it is seen that individual employee monitoring analogous to that heretofore provided for workers exposed to nuclear radiation or the like is now mandatory as regards noise. This can only be provided by portable individual audio dosimeters, worn by the employee during his entire work day, not only in the work area itself but also in the cafeteria, change house, or anywhere else he may visit on either a regular or irregular basis, and also facilities for daily quantitative read-out and recording of consummated exposures to permit appropriate duty assignments in the conduct of hearing conservation programs, as well as the identification of work areas of potential auditory peril.

DETAILED DESCRIPTION

The audio dosimeter of this invention is small (typically 1⅛ inches × 2¾ inches × 4¾ inches) and compact in size, light in weight (typically less than 7 ounces), and can be carried comfortably by the employee (as by neck band, belt or pocket clip or the like) without inconvenience or hindrance to work activities. Moreover, the dosimeter is reasonable in cost and rugged in design, so that it is well-suited to service in demanding industrial environments.

Referring to the block diagrams of FIGS. 4A–4C, inclusive, FIG. 4A shows the basic arrangement of audio dosimeter according to this invention wherein the sound signal is sensed by electronic microphone **10**, typically a Shure Bros., Inc., ceramic precision microphone, Model 99A 401B having a capacitance of 460 pF. at 80° F. and a nominal level characteristic of 59.5 dB below 1 v. per microbar at 400 cps measured in a free field at a distance of 12 inches from the sound source.

The a-c voltage signal output of microphone **10** is passed to filter-amplifier **11**, which, in addition to amplifying, shapes the output to conform to the preselected frequency response pattern of an "A" weighting network modeling the otolaryngologically (and psychologically) harmful contribution of ambient sound frequency.

The signal then passes to linear detector **12**, which rectifies the signal and passes the resulting d-c output current to non-linear network **15**. It has been found that, due to the fact that the function required to convert the voltage signal to a dB signal and the function required to convert a dB signal in turn to the requirements of the Walsh-Healey Law (which latter entails a factor of two change in signal current output for every 5 dB intensity sound signal change) are almost self-cancelling, only a small amount of shaping is necessary to make the signal conform to the requirements of the plot of FIG. **3**. Non-linear network **15** does this shaping and gives an output current producing a substantially straight line of correct slope in a plot of the d-c output current received from the linear detector in terms of decibels referred to a preselected current level versus sound input level in decibels, and passes the resulting signal to an electrochemical integrating cell **16** integrating sound exposure in terms of level-weighted sound pressure and weighted frequency and time of exposure conjointly.

Since the Walsh-Healey Law Noise Criteria extends only to a low limit of 90 dBA magnitude, it is desirable to preclude the measurement of sound emanations below this level. Thus, referring to FIG. 4B, an auxiliary has been devised for this purpose, which comprises a low limit detector circuit **20** connected in parallel with respect to non-linear network **15**, which operates switch **21** interposed between non-linear network **15** and electrochemical integrating cell **16** to cut out cell **16** from measurement service during any time interval in which the ambient sound energy level is below 90 dBA.

Similarly, since ambient sound energies above 115 dBA are particularly objectionable, the detection and recording of sound emanations in this excessive range is desirable. This is accomplished with yet another auxiliary comprising a latch and indicator circuit **24**, operating a light-emitting diode **47**, which auxiliary is shown in relationship to the basic circuit provided with a low limit detector circuit **20** and switch **21** in FIG. 4C.

Turning now to the detailed schematic circuit of FIGS. 5A and 5B, electronic micophonic sensor **10**, particularly the Model 99A 4401B hereinbefore cited as typical, includes in its internal structure a relatively large capacitance C of typically 460 pF value, which is indicated as being part of the microphone structure per se by the broken line enclosure. (If microphone **10** does not embody a capacitance of the magnitude recited, an appropriate size capacitor can be substituted in the apparatus circuit past plug-in connection **26**.)

It is preferred to encapsulate as much of the circuitry as possible in conventional potting resin to give a self-contained module, and the multiple open circle plug-in connections, such as **26**, drawn in FIGS. **5A** and **5B** denote points of electrical connection with circuitry encapsulated in modular form. This encapsulation contributes to the high inherent safety of the apparatus of this invention, particularly as regards service in atmospheres contaminated with explosive gases, which is aided by use of a low voltage source **31** possessing high internal resistance and low capacitances throughout the circuit.

The filter-amplifier of this invention comprises the operational amplifier **30** (typically an LM 301A) having feedback to the negative side only, in association with a plurality of a-c filter networks, each having individual band pass characteristics preselected to collectively interact to shape the a-c voltage output during transmission by the filter amplifier to conform to the preselected frequency response pattern of an "A" weighting network modeling the otolaryngocologically (and psychologically) harmful contribution of ambient sound frequency.

The embodiment of FIG. **5A** utilizes three individual a-c filter networks, as follows:

a. The C, R_1 network, wherein C has a capacitance of, typically, 460 pF as hereinbefore reported, where R_1 is typically a 7.5 megohm resistor constituting the filter resistive portion. C, R_1 constitutes the first high pass filter, passing frequencies above approximately 100Hz,

b. Parallel-connected C_1 and $C_{select)}$ in conjunction with resistor R_2 in parallel with resistor R_3 provides the second high pass filtering wherein typical values are $C_1 = 50$ pF, $C_{Sel} = 15$ pF and R_2 and R_3 each 8.2 megohms act in parallel for a-c signals, thereby passing frequencies above approximately 600 Hz. Resistors R_2 and R_3 coincidentally establish the d-c input voltage to amplifier **30** at substantially one half of the supply voltage provided by battery **31** (typically + 9v), and

c. Parallel-connected capacitor C_4 and Resistor R_6, typically 82 pF and 232 Kohms, respectively, constitute a low pass filter pole producing roll off at approximately 8KHz. Capacitor C_2 (typically 0.1 μf) in conjunction with resistor R_4 (typically 8.06 Kohms) and the variable Gain-Trim resistor **32** (typically 20 Kohms) provide the high pass filter action for this third filter.

The linear detection means now to be described includes resistors R_9, and R_{10}, capacitor C_8, diodes CR_3 and CR_4, and capacitor C_7.

The a-c output signal from amplifier **30** is passed via series-connected resistor R_9 (typically 1 Kohm) and coupling capacitor C_8 (typically 10 μf) to detector diodes CR_3 and CR_4.

The function of resistor R_9 is to provide a quasi-peak detector circuit having characteristics responding to noise signals in approximately the same manner as to sine wave signals.

Capacitor C_8 couples the a-c signal to the input of the detector circuit and stores the charge on positive excursions under the action of diode CR_3, thereafter discharging the signal into capacitor C_7 by way of diode CR_4.

Detector diodes CR_3 and CR_4 (typically both type IN4148) are connected with CR_4 in series and CR_3 in shunt. Thus, on the positive-going excursions of the signal, diode CR_3 conducts current to ground, charging capacitor C_8, whereas, on the negative-going excursions of the signal, CR_4 conducts to the output and filter capacitor C_7 (typically 56 microfarads) which constitutes the main filter on the detector output.

Turning back to amplifier **30**, the feedback is provided through diodes CR_1 and CR_2 (typically both type IN4148) connected back-to-back, which provide compensation for the forward voltage drop of detector diodes CR_3 and CR_4.

Resistor R_7 (typically 150 Kohms) affords an intentional leakage path around diodes CR_1 and CR_2, limiting the maximum gain of operational amplifier **30** for small signal cases. Resistor R_6 (typically a 232 Kohm metal film resistor) establishes the gain of the amplifier circuit in conjunction with resistor R_4 (typically an 8.06 Kohm metal film type) and gain-trim resistor **32.**

Capacitor C_2, previously described, is additionally an a-c coupling for the gain path such that the a-c gain is determined by resistor R_4 in series with gain-trim potentiometer **32** and in ratio with resistor R_6. However, the d-c gain of amplifier **30**, as connected, is nominally one, since there is 100 percent negative feedback at d-c.

Resistor R_5 (typically 3.6 megohms) is parallelconnected with capacitor C_3 (typically 0.01 μF) from point **52** to the negative input of amplifier **30.** Resistor R_5 is a bias compensation resistor used to equalize the biasing at the amplifier input, the value of R_5 being preselected to be nominally equal to the parallel value of resistors R_2 and R_3. Capacitor C_3 serves as a bypass capacitor maintaining low a-c impedance across resistor R_5.

Capacitor C_5 (typically 10 pF) is a damping capacitor on amplifier **30,** and conductor **34** connects the amplifier's negative supply terminal to ground. The several conductors denoted "COM" in FIGS. **5A** and **5B** are intended to be the "common" referred to, which can be instrument ground.

The filter-amplifier circuitry is completed by resistor R_8 (typically 6.8 Kohms) connected in series with capacitor C_6 (typically 2.2 μF) and linearity trim resistor **35.** Resistor R_8 is a dummy load resistor, which, in conjunction with linearity trim resistor **35,** imposes a loading on feedback diodes CR_1 and CR_2 which loading is adjusted for small signal level linearity trim. Capacitor C_6 serves as an exclusive a-c coupling for the linearity trim path inclusive of resistor R_8 and linearity trim resistor **35.**

The detection circuit is completed by resistor R_{10} (typically 1 megohm) which shunts rectifier CR_3 to common (or ground) thereby discharging capacitor C_8 when the signal level has decreased.

The signal is next routed, via conductor **51,** to a non-linear shaping network comprising diodes CR_5 – CR_7, and resistors R_{11}–R_{14}, both inclusive. The purpose of this non-linear shaping network is to bring the signal into straight line form as regards a plot of output current in dB referred to a given current value (e.g., 111μ amp) versus sound energy input level in dBA (e.g. 115 dBA = 119mv AC RMS 1KHz) as shown in FIG. **3** for levels above +90 dBA. FIG. **3,** for example, requires approximately 6 dB change in current for every 5 dB change in signal level.

The shaping effected by this non-linear network changes the response to achieve closely an approximate factor of two change in output current delivered to electrochemical integrating cell **16** for every 5 dB intensity of sound application to microphone **10,** which response is plotted for typical instrument performance in FIG. **3.** This represents close conformance to the Walsh-Healey Law Criteria (Some foreign countries have proposed, at least tentatively, different standards. Thus, the International Organization of Standards for certain European countries prescribes a two-fold increase in current for every 3 dB intensity of sound application to microphone **10.** Obviously, a different non-linear network would be required for accommodation of these different standards. Similarly, individual countries could require a different frequency response than that specifically obtainable with the "A" weighting network, and the substitution of such different weighting networks is completely feasible in this invention by the use of filter-amplifiers **11** having different parameters).

The non-linear shaping network comprises series-connected diodes CR_5, CR_6 and CR_7 (all typically types IN4148) connected also in series with resistor R_{11} (typically 57.6 Kohms) and thence to signal output terminal **38.** By-pass resistor R_{12} (typically 71.5 Kohm) is connected directly to terminal **38** from a point between diodes CR_5 and CR_6, and series-connected resistors R_{13} (typically a 33.2 Kohm metal film type) and R_{14} (typically a 51.1 Kohm value) are parallel-connected to outut terminal **38** with respect to diodes CR_5–CR_7 and resistor R_{11} collectively.

Diodes CR_5, **6** and **7** apportion current through the several resistors in the following sense. When the input voltage exceeds approximately 0.5v.,CR_5 operates to force current through R_{12} in addition to the path afforded by resistors R_{13} and R_{14}. When the signal voltage exceeds approximately 1.5v., CR_6 and CR_7 also conduct, causing current flow through resistor R_{11} in parallel with the existing paths through resistor R_{12} and through series-connected R_{13} and R_{14}.

Thus, the functions of the several resistors are as follows: R_{11} in conjunction with the R_{12} and the R_{13}, R_{14} path imposes the dynamic impedance for the large signal region, R_{12} in conjunction with the R_{13}, R_{14} path imposes the dynamic impedance for the medium signal region, and R_{13} and R_{14} in series constitute the path from the detector to the output for the small signal levels.

The quantitative output of the basic audio dosimeter circuit hereinbefore described can be integrated by a commercially available electrochemical integrating cell **16** (typically a Bissett-Berman Model S-214 rated for about 4,000 milliampere-sseconds as full-charge integral).

Turning now to a low level signal detection and switch-off auxiliary which is an advantageous adjunct for the basic audio dosimeter, a preferred design is depicted in FIG. **5B** between and above non-linear network **15** and readout cell **16.**

This comprises an NPN transistor Q_1 (typically a type 2N3860) having its emitter connected to a point in circuit between resistors R_{13} and R_{14} constituting the small signal path of the non-linear shaping network hereinabove described and its collector connected to common (or ground). Transistor Q_1 is utilized as a shunt switch on the output current path to effectively short the output current to zero when the detected signal falls below the limit threshold (e.g., below 88 dB). Resistor R_{15} (typically 1 megohm) shunts the base to the emitter of transistor Q_1 establishing the minimum drive current required to turn Q_1 on.

Transistor Q_2 is a PNP type transistor (typically a type 2N4249) connected through its emitter to the positive voltage supply bus **14** and through its collector and resistor R_{16} (typically a 100 Kohm current limiting resistor) to the base of Q_1. Q_2 is an amplifier in the path driving Q_1, whereas resistor R_{16} limits the maximum current supplied to Q_1 when it is turned on hard. R_{18} (typically 1 megohm) shunts the base to emitter of transistor Q_2, establishing the nominal drive current required to turn Q_2 on at 0.5μ amp.

Transistor Q_3 is an NPN transistor (typically a type 2N3707) connected base-to-base with transistor Q_4 serving as an input amplifier in the path driving transistor Q_1. Resistor R_{17} (typically 470 Kohms) is a current limiting resistor interposed between the base Q_2 and the collector of Q_3, which limits maximum current in case transistor Q_3 is turned on hard.

Resistor R_{18} (typically one megohm) is connected from the transistor Q_2 emitter to its base, thereby establishing the current level at which transistor Q_3 turns transistor Q_2on. This current level is preselected to be nominally equal to the operating current level of the transistor Q_5 current source stage.

PNP transistor Q_5 (typically a 2N4249 type) functions as a current source producing a nominal 0.5 microamperes for the path through diode-connected transistor Q_4 (typically an NPN type 2N3707) and through Select resistor R_{30} (typically 36 Kohms) and resistor R_{21} (typically 180 Kohms) to the output of the linear detector means, via conductor **51.** Select resistor R_{30} permits preselection of the detector output voltage at which the voltage drop across R_{21} and R_{30} due to the 5μ ampere current supplied by Q_5 causes the emitter of Q_4 to be at virtual ground.

The base-collector connection of diode-connected transistor Q_4 is connected to the base of Q_3. For the detector output voltage that results in the emitter of Q_4 being at virtual ground, the voltage at the base of Q_3 is appropriate to cause Q_3 to produce an emitter current equal to the emitter current in Q_4 (0.5 microamperes). This current, most of which appears as collector current in Q_3, is the required amount to turn on Q_2 and thus Q_1. Thus, a specific detector output voltage level, or less, produces the condition which turns on Q_1, thereby shutting off the integrating currents.

The circuit hereinbefore described, due to compensation of Q_3's base to emitter voltage (V_{BE}) by Q_4's V_{BE}, provides for switching of Q_1 at a constant detector voltage for a wide range of operating temperatures.

Resistor R_{19} (typically a 10 megohm resistor) is interposed in the transistor Q_5 emitter path and operates in conjunction with the reference voltage (V_R) established across the Q_{12} emitter to base junction to produce a specified current level from the transistor Q_5 collector.

Transistor Q_{12} is a NPN type (typically a 2N3563, 5.4v. V Base-Emitter Reverse Breakdown type), connected from bus **14** to the base Q_5, establishing reference voltage V_R. Q_{12} functions similarly to a Zener diode and this is why the collector is not connected (abbreviated N.C.). Resistor R_{20} (typically 100 Kohm) connected in circuit with the base of transistor Q_{12} provides operating current for the latter.

From the foregoing, it will be understood that, with Q_3 connected in base-to-base coupling with Q_4, the detector means voltage output is applied via resistors R_{21} and R_{30} to the emitter of transistor Q_4, which is a compensating transistor whose voltage drop, emitter-to-base, compensates for the transistor Q_3 emitter-to-base voltage drop.

Transistor Q_4 is supplied with current from transistor Q_5, which current is set by voltage reference V_R. The current of transistor Q_5 passes, via Q_4, through resistors R_{30} and R_{21}, which sets the detector output voltage for the cutoff point. Transistors Q_3 and Q_4 are at the same current at the threshold signal level.

When the negative polarity detector signal level decreases, the potential applied to resistors R_{21} and R_{30} is in the positive-going, or ground, direction, which turns on transistor Q_3, which latter turns on transistor Q_2. The collector of transistor Q_2 thereupon pulls in the positive direction, turning on transistor Q_1, which is the low sound signal level shunt switch precluding recording of low level sound by cell **16.** The circuit described provides switching of transistor Q_1 with a relatively small change in the detected sound signal.

Diode CR_8 (typically a type 1N457) is connected in reverse across supply bus **14** and common, thereby protecting the circuitry in the event of accidental reverse battery connection. Diode CR_8 can withstand the maximum current produced in this event due to the high internal resistance of the particular 9v. battery type **31** used.

Referring to FIG. **5B**, there is shown schematically a preferred design of latch and indicator circuit auxiliary adapted to furnish high level detection and retention of particularly harmful sound signal intensities, e.g., those exceeding the 115 dB level.

The detector output is introduced via conductor **41** in series circuit with resistor R_{22} (typically a metal film 200 Kohm type) connected to the emitter of transistor Q_6 (typically an NPN 2N3707 type), thereby establishing the current related to the detector output voltage which is employed to drive the latch circuit. Transistor Q_6 operates as a common base connected stage, the emitter of which is driven by the detector voltage through resistor R_{22}, producing an emitter current which is transferred as collector current through resistor R_{26} (typically 100 Kohms) to the base of transistor Q_8 (typically a PNP 2N4249 type). This current develops a potential across series-connected resistors R_{25} and R_{29}.

Resistor R_{25} conveniently consists of two series-connected separate metal film resistors A (typically 215 Kohns) and B (typically 215 Kohms), whereas resistor R_{29} is a "select" resistor (typically 56 Kohms). Resistor R_{29} is chosen to establish the current required through the resistor R_{22}, transistor Q_6, resistor R_{26} path such that there is generated a voltage drop across the resistor R_{25}, R_{29} combination which, at latch threshold level, will equal the reference voltage V_R level applied to the base of transistor Q_7 via conductor **43.**

Common emitter-connected PNP transistors Q_7 and Q_8 (both typically type 2N4249) form a pair sharing the current supplied through resistor R_{24} (typically a 470 Kohm resistor) connected to bus **14.** When the detector signal level is lower than the upper threshold level, transistor Q_7 conducts essentially all of the current from resistor R_{24}, and transistor Q_8 is then cut off. However, when the signal level exceeds the upper threshold, transistor Q_8 is turned on and then takes essentially all of the current from resistor R_{24}. The collector current of Q_7 is utilized to drive the base of NPN transistor Q_{13} (typically a 2N3707 type) so that the Q_7 collector current produces sufficient voltage across resistor R_{23} (typically 220 Kohm) to turn on transistor Q_{13}, which acts as a shunt switch discharging capacitor C_9 (typically 6.8 microfarad) and grounding the collector of transistor Q_8. Transistor Q_8's collector current is utilized so that, when Q_8 is turned on and Q_7 is turned off, which latter itself turns off Q_{13}, the Q_8 collector current is directed to charge capacitor C_9. Capacitor C_9 serves as a time delay capacitor in the latch circuit and is charged during the time the signal voltage exceeds latch threshold level, and is discharged, when the signal voltage falls below the latch threshold level, by the shunt switch Q_{13}. When capacitor C_9 is charged to approximately 0.5 volt, transistor Q_9 is turned on by Q_8's collector current.

Transistor Q_9 is an NPN type (typically a 2N3707) having its base-emitter junction connected across capacitor C_9 and, when C_9's potential is large enough (approximately 0.5v) to permit Q_9 turn on, transistor Q_9's collector current supplants the signal current fed to the base of transistor Q_8 through resistor R_{26}. When the collector current of Q_9 is sufficient to hold Q_8's base potential below the reference level at Q_7's base, the signal current supplied by way of the R_{22}, Q_6, R_{26} path is no longer needed and the circuit is latched.

The potential appearing at Q_9's collector is also applied to the base of PNP transistor Q_{10} (typically a type 2N4249) whose emitter is utilized to drive resistor R_{28} (typically 1 Kohm) connected with indicator output pin **38**, to drive, via normally open push button switch **46**, light emitting diode **47**. The power supply circuit for light emitting diode **47** is completed to the supply voltage source **31** via conductor **48**.

NPN transistor Q_{11} (typically a 2N3707 type) is connected at its base to the collector of transistor Q_{10} and at its emitter to common. Transistor Q_{11} augments Q_{10}'s emitter current as Q_{10}'s collector current drives the base of Q_{11}. Thus, transistor Q_{11} amplifies the current, delivering it as collector current back to the emitter of Q_{10} and the output resistor R_{28}. Transistor Q_{11}, in fact, delivers the major portion of the output current, Q_{10} being required to deliver only sufficient emitter current so as to produce a collector current sufficient to supply the base requirement of transistor Q_{11}.

It will be understood that the latch circuit hereinabove described detects the receipt by electronic microphonic sound sensor means **10** of sound energy in excess of a preselected high energy level, in the described instance 115 dB, preserving an indication and record of the fact by the latched condition of Q_8 and Q_9, which causes Q_{10} and Q_{11} to turn on the light emitting diode **47** when the push button switch **46** is closed manually.

At the end of the regular audio dosimeter service period, for example after a given 8-hour work shift the latch circuit is cleared of the illumination record by momentarily opening the power switch **49**, which removes power and restores the latch circuit to condition for reuse as desired.

In service, it is practicable to maintain an audio dosimeter bank from which each employee draws his own unit at the beginning of his work shift. At the completion of the work tour, the employee returns his audio dosimeter to the bank, where the integrating cell **16** is connected across the terminals of a commercial readout device (e.g., a Bissett-Berman Model 300 EDR) and the stored exposure in electrochemical integrating cell **16** read out and recorded as the sound exposure dosage to which the employee was subjected on the data involved. The duration of readout is timed so as to preserve the time correlation which is inherent in the employee's total hour shift exposure. Typically, a 10 ma deplating current applied for 10 seconds duration represents 100 percent exposure under the Walsh-Healey Law. At the same time, exposure to excessive sound levels, as indicated by the latch and indicator circuit auxiliary, can be noted and preserved.

It may be preferred to monitor only one employee of a given group and allocate identical sound exposure to all other persons in the same environment. Or, if desired, individual dosimeters can be mounted statically in specific work areas and the sound exposure profiles obtained for each area, independent of employee travel. Individual employee exposures can then be approximated on the basis of their residence times in the areas.

The practicability of encapsulating essentially the entire electronic circuitry into a compact module form is particularly advantageous from the standpoint of long service life under demanding environmental conditions, reliability in monitoring and consistent readings obtained with relatively large number of audio dosimeters.

What is claimed is:

1. An audio dosimeter for individual use comprising, in series circuit in the order recited, electronic microphonic sound sensor means,

a filter-amplifier provided with a feedback circuit receiving the a-c voltage output of said sound sensor means, said filter-amplifier comprising an operational amplifier connected in circuit with a plurality of a-c filter networks each having individual band pass characteristics preselected to collectively interact to shape said a-c voltage output during transmission by said filter-amplifier to conform to the pattern of a preselected weighting network incorporating in said a-c voltage output the otolaryngologically (and psychologically) harmful contribution of ambient sound frequency,

linear detector means rectifying said a-c voltage output from said filter-amplifier, said linear detector means incorporating a pair of oppositely connected diodes as rectifying elements, and said filter amplifier feedback circuit incorporating a pair of oppositely connected diodes preselected to compensate the forward voltage drops of said pair of diodes in said linear detector means and an impedance preselected to provide the desired linearity of rectification for small signals connected between signal ground and the junction of said compensating diodes on the feedback delivery side of said compensating diodes,

a non-linear network shaping the d-c output current from said linear detector to the function required to produce a substantially straight line of correct slope in a plot of decibels referred to a preselected current level versus sound energy input level in decibels, and an electrochemical integrating cell receiving the output current from said non-linear network measuring sound exposure in terms of sound pressure with weighted freqency and time of exposure conjointly.

2. An audio dosimeter for individual use according to claim **1** wherein said preselected weighting network is an "A" weighting network.

3. An audio dosimeter for individual use according to claim **1** provided with a latching circuit detecting the receipt by said electronic microphonic sound sensor means of sound energy in excess of a preselected high energy level and retaining a record of said receipt.

4. An audio dosimeter for individual use according to claim **1** provided with a latching circuit detecting the receipt by said electronic microphonic sound sensor means of sound energy in excess of a preselected high energy level provided with a light-emitting diode as indication means for readout of said record of said receipt.

5. An audio dosimeter for individual use according to claim **1** provided with low level detection means sensing the receipt by said electronic microphonic sound sensor means of sound energy below a preselected low energy level and means responsive to said low level detection means switching said electrochemical integrating cell out of measurement service during the receipt of said sound energy below said preselected low energy level.

6. An audio dosimeter for individual use according to claim **1** wherein said non-linear network shaping the d-c output current from said linear detector to the function required to produce a substantially straight line of correct slope in a plot of decibels referred to a preselected current level versus sound energy input level in decibels comprises a multiplicity of parallel-connected resistor paths automatically switched in at progressively higher signal voltage levels to provide preselected dynamic impedances collectively sufficient to obtain signal current doubling for preselected sound energy input levels in decibels conforming to a given permissible time exposure-sound pressure level pattern as standard.

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UNITED STATES PATENT OFFICE
CERTIFICATE OF CORRECTION

Patent No. 3,868,857 Dated March 4, 1975

Inventor(s) EDWARD L. MADDOX and ROBERT A. PEASE

It is certified that error appears in the above-identified patent and that said Letters Patent are hereby corrected as shown below:

Col. 3, line 62 - "micophonic" should be --microphonic--.
Col. 3, line 63 - "4401" should be --401--.
Col. 4, line 35 - "select)" should be --sel(ect)--.
Col. 5, line 28 - "parallelcon-" should be -- parallel-con- --.
Col. 6, line 54 - "sseconds" should be --seconds--.
Col. 8, line 54 - "Kohns" should be --Kohms--.
Col. 10, line 9 - "data" should be --date--.

Signed and sealed this 29th day of April 1975.

(SEAL)
Attest:

RUTH C. MASON
Attesting Officer

C. MARSHALL DANN
Commissioner of Patents
and Trademarks