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A feedforward side chain Limiter/compressor/de-esser with improved flexibility

A.J. Oliveira
Acutron Electroacoustica
Sacavem, Portugal

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A FEEDFORWARD SIDE CHAIN LIMITER/COMPRESSOR/ /DE-ESSER WITH IMPROVED FLEXIBILITY.

A. J. OLIVEIRA

ACUTRON ELECTROACUSTICA, LDA. P. 2685 SACAVEM PORTUGAL

A lo-cost, yet high perfomance dual Limiter/Compressor/De-esser package is described, which overcomes most limitations of conventional designs. Main features include: feedforward side chain construction; selectable balanced or single-ended RFI-protected input/output configuration; abrupt de-esser action; user selectable hard or soft-knee compression curves through independant threshold and compression ratio controls that won't fiddle with unit's before threshold gain, so avoiding time-consuming adjustments; a VCA design featuring a consistently high clipping headroom to avoid transient distortion; availability of very fast attack times, important to limiting action; and finally an unconventional dual-link control to allow for unusual applications.

Precision, temperature and drift-compensated circuits are used extensively to ensure consistent action in spite of use of standard components. The unit lends itself to partial integration and to applications like remote control and multiband compression.

0 INTRODUCTION

A fast survey of available dynamic controllers indicates that they are roughly divided in two categories: lo priced units of very limited usefulness and sometimes unfortunately of lo audio quality, featuring user controls that interact with each other making adjustments a continuous headache for the audio engineer, and, on the other hand, a few high perfomance units which have the drawback of being too much expensive for general-purpose use. So we thougth it was time to develop a cost-effective unit which perfomance should suit most dynamic controlling situations, and not fear comparison with very expensive designs in what concerns audio quality and freedom from side-effects.

Operation should be straight forward and essentially transparent to user as comparing to the nebulous perfomance of some available designs, both cheap an expensive. Finally the unit should be easily serviceable, modular and should be suitable for more complex systems, like multiband compression.

1 GENERAL DESIGN GUIDELINES

- Number of channels

The unit shall have two independant channels that can be coupled for stereo operation, so avoiding displacements of stereo image caused by different amounts of compression being applied to the two channels.

- Inputs

The unit shall have both balanced and unbalanced inputs, both of the electronic type. Any of inputs shall accept without overload a maximum level of +22dBu, regardless of unit's control settings. This is done to eliminate hard transient distortion, particularly at slow attack times. Impedance shall be 100 KOHM for balanced input and 50 KOHM for unbalanced input, ± 10%. Any of inputs shall have a first - order RFI protection which will preserve the audio bandwidth (-1 dB at 20 KHz). Commom mode rejection ratio of balanced input from 20 Hz to 20 KHz must exceed 70 dB.

- Outputs

Two outputs shall be provided,one balanced and one unbalanced of the electronic type. Maximum output level shall be consistent with maximum input level, i.e. + 22 dBu. Output impedance shall be 600 OHM for both cases.

- Gain

Unit's gain before threshold, regardless of input/output balanced or unbalanced configuration being used, and of control settings, shall be unity (0 dB ± 1 dB). This is done to make the actual adjustment of the unit threshold and ratio controls easy and transparent, eliminating input and output level controls.

- Frequency response

Unit's frequency response shall extend from 20 Hz to 20 KHz (useful audio bandwidth) with no more than 1 dB attenuation at the two extremes, regardless of gain riding being used and control settings. Lo frequency transient response shall be according with findings of [1]. Limitation on high end shall be of the passive type to avoid RFI and transient intermodulation distortion [2].

- Noise level

The noise level with the unit operating at maximum gain (unity) shall not be greater than — 70 dBu, unweighted, 20 — 20 KHz measurement bandwidth. This measure shall include hum components. Of course "A" weighted noise level shall be consistently lower.

- Crosstalk between channels

Crosstalk figure shall be — 70 dB unweighted, from 20 Hz to 20 KHz, measured at any level before overload and with unit operating again at unity gain (no compression).

- Gain reduction capabilities

The unit shall be capable of gain reduction in excess of 32 dB, to accomodate a variety of dynamic reduction situations. When the stereo link mode is selected, tracking between channels shall be better than ± 1 dB.

- THD

We shall impose an upper limit for THD of 0.2%, at any level below maximum input/output level. This limit is regarded as an acceptable level for usual applications [3].

2 THE ERGONOMICS ... USER ACESSIBLE CONTROLS

User acessible controls shall be, on each channel:

- A "threshold" control shall be adjustable from - 12 dBu to + 12 dBu, so to accomodate all commonly used working levels.
- An "attack" control shall permit adjustment from 0.1 to 10ms. The first figure represents a very fast attack time and can be used effectively to ease the fast limiting task.
- A "release" control that shall be variable from 0.1 S to 10 S. This control shall not interact with the "attack" control.
- A compression "ratio" control that shall be variable from 1:1 (no compression), to ∞ :1 (limiting); this ratio shall be measured by the following expression:

$$(1)\quad CR = \frac{IR}{OR}$$

Where: CR - compression ratio

IR - input excursion in dB, above threshold (for definition purposes IR = 20 dB)

OR - output excursion in dB, above threshold, corresponding to 20 dB over threshold excursion at input.

The reason for this definition lies in the fact that we chose soft-knee compression curves for the low ratio values, that convert to hard-knee as the compression ratio control setting goes higher. This is done both to minimize "pumping" and to eliminate the use of a separate limiter at lo control settings, so lowering total costs. In this way at all ratio settings but infinity, the asymptotic ratio shall progressively get higher as the input signal grows with respect to threshold. This asymptotic ratio can be defined as the local derivative of the compression ratio function (1)(previously defined) at point I :

$$(2)\quad CR'_{(I)} = \left[\frac{d\ IR}{d\ OR}\right]_I$$

Where: I — input expressed in dB above threshold.

$CR'_{(I)}$ — asymptotic compression ratio.

dIR — local increment of input signal, in dB, made as small as possible.

dOR — increment of ouput signal, in dB, corresponding to dIR.

This function 2) shall tend to infinity as we deviate from threshold. We will go back to this point later on.

The "ratio" control shall finally be completely independant from the threshold control, so as to permit a fast, foolproof adjustment of the unit, and of course, independant from time adjustmens.

- A double "link" control for the two channels shall be provided. This represents a different situation from the conventional setup, where only one link control is used to cross-couple the side chains of both channels so the control signal of the side chain which is higher takes control over both channels. With our arrangement we can keep one of the two channels completely independant and the other slaved to the first one, by pressing its link switch. In this way, the dynamics of the second channel are slaved to the first, while keeping the first one independant. This can be useful to implement special effects, particulary at the recording studio. Of course if the user presses the two "link" switches, he will get the conventional cross-coupling of channels in order to keep a stable stereo image.
- A de-esser on/off control on each channel shall be provided, enabling a frequency-discriminating S - supression by lowering the threshold 12 dB in 4 — 20 KHz band. From 4 KHz down, return to normal threshold shall process itself at a 12 dB / octave rate, enabling a discriminating de-essing action.
- A "bypass" control shall be incorporated in each channel,to allow for elimination of the gain control module in the chain; input and output amps shall be left connected to permit unity gain use in balanced/unbalanced input/output configuration and reverse situation.

- The unit, for simplicity, shall incorporate only the following optical indicators, on each channel:

An "overload" indicator will light when the input signal gets in the vicinity of + 17 dBu $\pm$ 3dB. A "threshold" indicator will light when the input signal gets over threshold. This indicator shall be very precise and temperature compensated, so as to permit accurate threshold control adjustment.

An "end of compression" indicator will light when 32 dB gain reduction is achieved, to prevent depassing of the unit's gain riding capabilities.

3 PHYSICAL CONSIDERATIONS: EXTERNAL AND INTERNAL

- Physical size: the unit shall be presented in a standard 19" rack-mountable case of 2 units height.
- Internal layout: three PCB modules shall be provided: one for VCA, input-output sections and power supply, and two for the side chains, directly coupled to front panel controls, enabling wiring minimization and ease of expansion to applications such as multiband compression and remote control of VCA section.

4 CHOISE OF SIDE CHAIN CONFIGURATION

Two basic options are available to the design engineer in what concerns the circuit topology used to implement DC gain control: the feedback approach and the feedforward approach. Let's examine the two to check their basic demands on VCA transfer curves.

If we assume unity (0 dB gain) of the VCA before threshold, we can write down the basic relationships, directly coming from expression 1), if we ignore, for the moment, the effect of detector time constants:

$$(3)\quad \frac{\mathrm{Log}\,(V_i/V_t)}{\mathrm{Log}\,(V_o/V_t)} = CR$$

Consequently:

$$\mathrm{Log}\,(V_i/V_t) = CR * \mathrm{Log}\,(V_o/V_t)$$

$$V_i/V_t = 10^{(CR * \mathrm{Log}\,(V_o/V_t))}$$

$$V_i/V_t = (V_o/V_t)^{CR}$$

$$V_i = \frac{V_o^{\,CR}}{V_t^{\,CR-1}}$$

And finally:

$$(4)\quad V_o = (V_i * V_t^{(CR-1)})^{(1\,/\,CR)}$$

If we try to rewrite expression (4) in terms of extracting attenuation $A = V_i/V_o$ as a function of V_t/V_o, leading to feedback control of VCA (FIG.1) we get:

$$\frac{V_o^{\,CR}}{V_i} = V_t^{\,(CR-1)}$$

$$\frac{V_o}{V_i} = \left(\frac{V_t}{V_o}\right)^{(CR-1)}$$

$$(5)\quad A = \left(\frac{V_o}{V_t}\right)^{(CR-1)}$$

If alternately we repeat the procedure in terms of extracting attenuation as a function of V_t/V_i (feedforward configuration, Fig. 2), we get:

$$\frac{V_o}{V_i^{\frac{1}{CR}}} = V_t^{\left(\frac{CR-1}{CR}\right)}$$

$$\frac{V_o}{V_i} = \left(\frac{V_t}{V_i}\right)^{\left(1-\frac{1}{CR}\right)}$$

$$(6)\quad A = \left(\frac{V_i}{V_t}\right)^{\left(1-\frac{1}{CR}\right)}$$

If we redo the all procedure from (3) to (6) this time starting from Log relationships between variables, we get (with "l" suffix denoting Log relationship).

$$(7)\quad \frac{V_{il} - V_{tl}}{V_{ol} - V_{tl}} = CR$$

$$V_{il} - V_{tl} = CR * V_{ol} - CR * V_{tl}$$

$$(8)\quad V_o = \frac{V_i + (CR - 1)\,V_t}{CR}$$

For the feedback configuration we get:

$$V_{il} = CR * V_{ol} + (1 - CR) * V_{tl}$$

$$V_{ol} - V_{il} = (CR - 1) * V_{tl} + (1 - CR) * V_{ol}$$

$$V_{ol} - V_{il} = (V_{ol} - V_{tl}) * (1 - CR)$$

$$(9)\quad A_l = (V_{ol} - V_{tl}) * (CR - 1)$$

And for the feedforward one:

$$V_{ol} - V_{il} = \frac{V_i}{CR} + \frac{CR - 1}{CR} * V_{tl} - V_{il}$$

$$V_{ol} - V_{il} = V_{il} * (\frac{1}{CR} - 1) + (1 - \frac{1}{CR}) * V_{tl}$$

$$V_{ol} - V_{il} = (V_{il} - V_{tl}) * (\frac{1}{CR} - 1)$$

$$(10)\ A_l = (V_{il} - V_{tl}) * (1 - \frac{1}{CR})$$

Plots of (9) and of (10) assuming V_{tl} = 0 dB can be seen respectively in Fig.3 and 4. Anyway, by simple mathematical analysis of expressions (9) and (10) we can deduct that, if we want a compression ratio variable from 1:1 to ∞:1, at least teoretically we must choose the feedforward approach, because it is the only one that, with finite side chain gain, will provide ∞:1 compression (limiting). To provide that, feedback approach must use a switching type function. This is, at least to the extent of my analysis, imcompatible with adjustment of side chain gain in order to provide lo compression ratio settings.

Thus, I feel that with feedback approach, only limiting aproximations will be possible and that adjustment of high compression values, because of dependance on an open-loop type transfer function, will be rather unstable and temperature - dependant.

I must not go on without considering the effects of feedback control on time constants of detector, wich tend to vary with side-chain gain, turning into another inaccuracy factor. The effect may be described in the following way: in the side chain of any configuration, feedback or feedforward, we got two time constants, the attack time constant T_a which controls the time in which the compressor gain changes from it's before threshold value to the one dictated by the incoming over threshold signal, and the release time constant T_r, which controls the time in which compressor gain returns to the idle value after the over threshold input has been removed.

If we return to expressions 5) and 6) and multiply respectively the control variables V_o and V_i by (see appendix):

$$(11)\ 1 - e^{-\frac{t}{T_a}} \qquad T_a \text{ — attack time constant.}$$

Or

$$(12)\ e^{\frac{-t}{T_r}} \qquad T_r \text{ — release time constant.}$$

We get, for the feedback case:

$$(13)\ A = (\frac{V_o}{V_t})^{(CR - 1)} * (1 - e^{-\frac{t}{T_a}})^{(CR - 1)}$$

Or

$$(14)\ A = (\frac{V_o}{V_t})^{(CR - 1)} * e^{\frac{-t}{TR} * (CR - 1)}$$

And for the feedforward:

$$(15)\ A = ((\frac{V_i}{V_t}) * (1 - e^{\frac{-t}{Ta}}))^{(1 - \frac{1}{CR})}$$

$$(16)\ A = (\frac{V_i}{V_t})\ e^{\frac{-t}{T_r} * (1 - \frac{1}{CR})}$$

By inspection of equations 13, 14, 15 and 16, we can deduct that the attenuation will process itself in the feedback case with the time constants divided by (CR - 1) while in the feedforward approach it will come divided by $(1 - \frac{1}{CR})$. If we check maximum variation of the two terms we will readily see that the first will vary between 0 and CR while the second will at most vary between 0 and 1, this for CR excursions from 1 to infinity. Thus we may see that the feedforward approach is also superior in this respect, eliminating the extreme variations of time constants with compression ratio settings found with feedback designs, unless special measures are taken to combat this. Furthermore, time constants are not fixed in feedback designs and vary with input signal amplitude. Let's check this in more detail: if we reverse the time constant situation, i.e. specifying fixed time constants for attack and release that the user can rely on, we should have an attack expression of the form, where V_{oc} is the compressed output voltage resulting from equation (4):

$$(17)\ V_o = V_{oc} + (V_i - V_{oc}) * e^{\frac{-t}{Ta}}$$

Which is easily converted by use of equation (4) in:

$$(18)\ V_o = V_i^{\frac{1}{CR}} * V_t^{(1 - \frac{1}{CR})} * (1 - e^{\frac{t}{Ta}}) + V_i * e^{\frac{-t}{Ta}}$$

Which, develloped in order to extract gain V_o/V_i gives:

$$(19)\ \frac{V_o}{V_i} = (\frac{V_t}{V_i})^{(1 - \frac{1}{CR})} + (1 - (\frac{V_t}{V_i})^{(1 - \frac{1}{CR})}) * e^{\frac{-t}{Ta}}$$

If we repeat the procedure in terms of the release time constant we are lea_ded to:

$$(20)\ V_O = V_{OC} + (V_I - V_{OC}) * \left(1 - e^{\frac{-t}{Tr}}\right)$$

$$(21)\ V_O = V_I + V_I^{\frac{1}{CR}} * V_T^{\left(1 - \frac{1}{CR}\right)} * e^{\frac{-t}{Tr}} - V_I * e^{\frac{-t}{Tr}}$$

$$(22)\ \frac{V_O}{V_I} = 1 + \left(\left(\frac{V_T}{V_I}\right)^{\left(1 - \frac{1}{CR}\right)} - 1\right) * e^{\frac{-t}{Tr}}$$

From equations (19) and (22) we can readily conclude that, to have fixed time constants within the unit the only kind of possible control is feedforward control (because the V_I term in equations (18) and (21) is present two times with different exponents). It is impossible to obtain V_O/V_I as a single function of V_O in this case, because a V_I control term will subsist, demonstrating my previous assumption that the time constants would be a function of input level for the feedback approach. Control approach using equations (19) and (22) is however difficult to implement, requiring complex circuitry. It is preferable to tolerate some variation on time constants through the use of a control approximation, taking care to minimize such variations to the extent of possible, to which previous information may prove useful.

Furthermore, transient problems are unavoidable in feedback designs and perfectly controlable in feedforward ones, through the use of delay in the VCA input, immediatly after the side chain insertion point.

For the reasons previously laden, the feedforward approach was chosen for this particular design.

As a final note, it will be obvious that mixed feedback/feedforward designs are possible, for variations of CR from 2:1 to ∞:1, and we will note that previous attenuation equations are valid only for perfect compression systems; in practice, things turn out to be different, depending on the control approximation used. In fact, examination of equations (5) and (6) shows that their exact implementation would necessitate rather complex circuitry to be made in the analog domain, being extremely simple in the digital one.

5 VCA ARCHITECTURE

As in this design main concern is to keep costs at a reasonable level and make the unit serviceable anywhere (so avoiding exotic parts), an OTA (operational transconductance amplifier) was chosen as the VCA gain control element [5], [6]. This leads, as we shall see, to advantages and drawbacks, as in any design.

If design objectives of section 1 are to be fullfilled in the VCA, this shall feature:

A - Constant +22dBU headroom to avoid transient clipping at 1:1 compression ratio setting.

B - As low a noise floor as possible in order to keep a high dynamic range.

C - As large as possible gain reduction without compromising requirements A and B.

D - High linearity of attenuation versus control voltage in order to extract accurate limiting from the side chain configuration chosen.

In presently available OTA designs, for example the ones outlined in [5] and [6] two gain control modes are available. One is implemented by transconductance (gm) control by means of a current (I_{ABC}). This dependance is of the type:

$$(23)\ gm = K1 * I_{ABC}$$

Usually this function is quite linear (constant K1) over a few decades, giving rise to very useful control.

Another available gain control can be implemented using current controlled bias predistorting diodes. This can be looked at as resistance (R_D) control by an external current (I_D). This function is also linear over a few decades, as the previous one, and is of the type:

$$(24)\ R_D = K2/I_D$$

Of course, as this resistance presents itself on the OTA input, it can be used to control gain in conjunction with an external series resistance we shall call R_B (Fig.5). This combination obviously forms an attenuator, whose voltage transfer function $H = V_D/V_I$ can be described by:

$$(25)\ H = \frac{1}{1 + R_B * I_D/k2}$$

Function (25) is not obviously a linear function of I_D (it's first derivative is itself a function of I_D).

It can only approximated by a linear function for a relatively small I_D excursions referred to total R_D linearity versus I_D. Thus this kind of control is not suitable for our purposes, and function (23) will be used instead. Since our need is a circuit providing an attenuation proportional to a control variable, circuit of Fig.6 will be used, using OTA as a controlled resistance REQ. By inspection we may write, using (19) and assuming a an ideal opamp:

(26) $V_D = H * V_O$

Using (23) we have

(27) $I_O = K1 * I_{ABC} * H * V_O$

If we remember that (-) input of A is a virtual ground node and assume infinite input impedance for the opamp we may write, posing $I_O = - I_I$ and $I_I = V_I/R_A$

(28) $$\frac{V_O}{V_I} = - \frac{1}{K1 * H * I_{ABC} * R_A}$$

Or from Fig.7 [7] we may write:

(29) $$\frac{V_O}{V_I} = - \frac{REQ}{R_A}$$

Wich in return gives

(30) $$REQ = \frac{1}{K1 * H * I_{ABC}}$$

We will now look at the dynamic constraints of this type of design, starting from the overload side. I_O is limited to some value equal to or lower than I_{ABC}, so we can set:

(31) $$\left| I_{O\,PEAK} \right| \leqslant K3 * I_{ABC} \qquad K3 \leqslant 1$$

And if we remember considerations posed prior to writing eq (28):

(32) $$\left| V_{I\,PEAK} \right| \leqslant K3 * I_{ABC} * R_A$$

Finally converting V_I to sinus RMS value and rewriting (32) in terms of R_A we get:

(33) $$R_A \geqslant \frac{\sqrt{2} * V_{I\,max}}{K3 * I_{ABC}}$$

Expression (33) puts a constraint on the minimum value R_A must have for over load-free reproduction, given a certain I_{ABC} or conversely the minimum I_{ABC} for a known R_A.
If we now look to predistortion diode network bias current, it will follow [5] that this current shall be equal to or greater than maximum network input current (through R_B) if distortion is to be avoided. So:

(34) $$I_D \geqslant \frac{\sqrt{2} * V_{O\,max}}{R_B + R_D}$$

Let's now look to what happens with the noise floor. Noise in the circuit configuration selected is governed by OTA equivalent input noise voltage $\bar{e}n$. This applies because most gain is concentrated in the controlled transconductance section. This means we can neglect noise contribution from opamp, input resistor R_A, predistorting diode network and finally OTA input noise current, because it is flowing through a fairly low input impedance (that of linearizing network). If we assume that $\bar{e}n$ has been calculated or measured for the concerned bandwidth, noise at the output will be resultant from the OTA'S output noise current flowing through REQ (eq (30)) so we can write, using (23) and (30)

(35) $$V_{on} = \frac{\bar{e}n}{H}$$

Or if we prefer, taking (29) into account and with $G = V_O/V_I$

(36) $V_{on} = K1 * I_{ABC} * R_A * G * \bar{e}n$

Equation (36) has the virtue of showing us that we should use as low I_{ABC} as possible for less noise, assuming $\bar{e}n$ constant with I_{ABC}.
If we look instead to equation (35) we shall readily see, using also equation (25) that to keep H high we must use the smaller possible value for I_D.
From the present constraints it is obvious that the best choise is to set expressions (33) and (34) to exact equalities. If we then take equation (28),

and substitute with equation (25) and with expression (33) set to equality we get:

$$(37)\quad G = \frac{1}{K1 * \left(\frac{1}{1+R_B * I_D/K2}\right) * \frac{\sqrt{2} * V_I\ max}{K3}}$$

Introducing now R_B as governed by (34) set to equality and taking into account that $V_O\ max = G * V_I\ max$ we have:

$$(38)\quad R_B = \frac{\sqrt{2} * G * V_I\ max - K2}{I_D}$$

Applying expression (38) to expression (37) yelds

$$(39)\ K3 = K1 * K2$$

This way, constraint obtained by setting expressions (33) and (34) to equality can only be applied when we can manipulate K3 to satisfy (39). In this case, maximum headroom is obtained. Of course if this is true, from (25) and (38) results:

$$(40)\quad H = \frac{K1}{K1 + \sqrt{2} * G * V_I\ max - K2}$$

Finally let us calculate the maximum dynamic range, from noise floor to overload. Using (33) set to equality, (28) and (36), we see that, setting D = $= V_O\ max / V_{on}$

$$(41)\ D = \frac{K3}{\sqrt{2} * K1 * \bar{e}n}$$

Dynamic range is thus directly proportional to K3 and inversely proportional to K1 and ēn, and a fixed carachteristic of an individual VCA design using an OTA, regardless of gain used. This also is an indication that the design will not suffer from self noise pumping.
Using all of the exposed criteria, I arrived to a final VCA design wich features a commercially available OTA and that was cabable of the following perfomance:

THD 1 ⩽ 0.2% AT + 20dBU
Bandwidth (-1dB) = 20KHz
Noise ⩽ -66dBU unweighted, 20KHz bandwidth
Max gain reduction = 34dB (consistent with lowest threshold and nominal headroom).
Linearity ⩽ 0.1% for above range

As it is obvious, noise floor was too high for a demanding application. Further noise decrease was obtained using complementary preemphasis/deemphasis, taking care not to degrade headroom so that upper threshold was still usable with pink noise, maintaining 3dB headroom. I finally got as result, using this technique, $T_{P/D}$ being the preemphasis/deemphasis time constant:

$T_{P/D}$ (US)	Noise (dBU)	Headroom degradation (dB)
50	-73	7
75	-74.5	8.5

As it is obvious, headroom loss using pink noise is in this case equal to the lowering of noise floor, so dynamics calculated through equation (41) still apply. However, perceived noise will be substantially lower. This tecnique, if deemphasis is implemented within the VCA feedback loop to lower parts count and PCB space, avoiding subsequent buffering, requires for extremely careful R/C shunt, phase compensation to avoid HF oscilation, and relies on pole/zero cancellation for accurately flat response. I choose, also for compatibility with broadcasting use, the 50US time constant used in Europe for FM transmissions (because it allows for deemphasis at the receiver).
During the experimental phase, a very good correlation was observed between theoretical calculated and experimentally obtained values. For example differences observed in the dynamic range were less than 2dB, and were attributed to resistor approximation to standard E12 values, confirming my noise neglecting approximations.

6 INPUT/OUTPUT CIRCUITRY AND POWER SUPPLY

Input differential to single ended converter is a straightforward unity gain circuit [7], where care in component matching results in a common-mode rejection in excess of 70dB through the useful bandwidth. First-order passive filtering is used at the input to provide imunity to stray RF and interference signals.

Separate output circuits are provided for unbalanced and balanced outputs, to insure that unity idle gain is maintained irrespectively of input/output configuration used. To provide this, output signal used to provide the unbalanced output is attenuated 6dB before being fed to an electronic single ended to differential converter, in order to provide the balanced output. In either output, impedance is fixed at 600 OHM by means of output resistors. Power supply was configured using oversized filtering and dual integrated circuit regulators in order to obtain a voltage regulation better than 0.1% and ripple rejection of over 70dB. It is built to work from 220VAC ± 20% standard power line maintaining the previously specified regulation figure, which enables unit use in quite poor power lines sometimes encountered in field work.

7 OTA/SIDE CHAIN/USER INTERFACE DESIGN

As the chosen VCA design is current-controlled rather than voltage controlled, and that we need for supply voltage and temperature independant VCA perfomance, two voltage to current converters were connected to the controlling inputs I_D and I_{ABC}. These are inverting converters where input current is equal to output current, because the two flow in opposite direction into a virtual ground node [7] .

Currents obtained through the V/I converters are used to drive the internal OTA current mirrors, thus providing at the same time level translation and improved control feedthrough of basic OTA design. This is mainly due to the dependance of control current on a voltage with respect to ground instead of a voltage with respect to one of the supplies. Anyway, supply voltages are stabilized to 0.1% in this design, avoiding other errors as we will see next.

The I_{ABC} input is in this design a compression input, because increased current results in gain reduction. I_D input is conversely an expansion input, for opposite reasons. As we are not interested basically in using the expansion input (as the noise floor is lo enough to avoid tricks in order to mask self VCA noise), we will drive the expansion input with a reference voltage (translated to a reference current driving the I_D input), and make the resultant current variable by means of a preset pot. This will alow for 0dB gain trimming ensuring 0.1dB absolute precision and matching between channels. This reference voltage was chosen to be coincident with the threshold point peak voltage for a -12dBU sine wave, since as we will see, peak detection was chosen in this design.

Reference voltage is derived from a temperature compensated zener driven at zero TC current through a resistor fed by one of the supply voltages (as the supply is stabilized this is equivalent to driving it with constant current at least to the extent of supply/zener voltage relationship). Zener output is then buffered to eliminate variable loading effects.

This same voltage is used to drive a comparator fed with the I_{ABC} control voltage wich in turn drives a led to alert user that threshold level has been depassed. It is used also, after driving a 32dB voltage amp to drive another comparator fed with the same I_{ABC} control voltage, and also driving a led, to indicate that maximum gain reduction has been achieved (maximum gain reduction indicator). These two indicators alow the user to know both when threshold level was entered and when he shall manually decrease the gain because of depassing of automatic gain riding capabilities.

Finally, another indicator is provided to alert user that maximum input level is been approached, so avoiding overload distortion.

Conjunction of the three preceeding indicators allow for precise unit adjustment without the redundancy of doubling the meters of the preceeding/following stage, and with the added convenience of positively allerting user for out of range condition, as it is not the case with continuous reading meters, where too much information requires too much attention to spot individual situations.

Main PCB configuration, containing VCA as well as power supply and reference voltage circuits is shown in figure 8. Also included are the input/output conditioning circuits.

8 SIDE CHAIN DESIGN

Beginning at the input, chosen side chain design features a switchable DS filter. As the required perfomance was rapid threshold lowering from 4KHz up, a second order filter was chosen. Output of this filter adds with input to provide proper frequency shapping (Fig.9). In this way, very heavy and discriminative de-essing is available, of the ducking type (gain reduction in the entire bandwidth), to be used when needed.

After the switchable de-esser section, a precision full-wave rectifier featuring variable gain controlled by an external user-acessible pot. Gain ranges from -23.3 to +0.6dB, alowing easy adjustment of threshold. To make adjustment easy and positive, an inverse log law potentiometer was employed, yelding the scale pictured in Fig.10. This type of scale permits great adjustment precision around the threshold levels most used today.

Following we have an envellope generator, where attack and decay times are defined. It consists basically of a peak detector optimized for this use, which block diagram is shown in Fig.11.
As seen, a comparator senses the difference between input signal from rectifier and capacitor voltage. Its output is used to drive two alternate switches, which, connected to the capacitor and to the supply voltages, charge it through R_A and discharge it through R_R, resulting in the time constants given by:

(42) $T_A = R_A * C$

(43) $T_R = R_R * C$

This way, time constants are program-independant and (another useful feature if we eventually want to work near to zero DC level with precision) ouput voltage charges or discharges to zero volts DC (neglecting comparator offset voltage) folowing time constants of (42) and (43). This is resultant of both the comparator control and the release drive with a negative voltage, bypassing switch losses.
One more feature of this circuit is the use of a high current switch to drive the capacitor in the attack mode, alowing the use of a very small resistor to drive C and consequently very short attack times without having to make use of ultra high value special potentiometers to achieve long release times.
Log curve pots are used to control attack and release time constants, resulting in the easy to read scales of Fig.12.
Peak detection was chosen for this design, for two reasons:

1- It provides faultless limiting if attack time is short and compression is set to maximum;
2- More response variations are induced in the program by varying detector time constants than the type of detector. This is apparent with most program material and eliminates apparent advantages of RMS detectors.

Next to the preceeding section comes a level detector, in the form of a precision maximum selector, that selects either the input or the threshold reference voltage (whichever is greater), determining level beyond which VCA will receive the control voltage generated by the preceeding stages or just the reference voltage. It provides the function pictured in Fig.13.

Following the level detector we have the section responsible for the variation of compression ratio. It consists of the circuit outlined in Fig. 14 . This circuit approximates function defined by equations (4) and (9) in a very effective way, providing at the same time the "soft-knee" compression curves that have the advantage of making the compression less audible, this at intermediate ratio settings. Let's analyse this control function extensively. As you can see in figure 14, the ratio pot, chosen to be a linear one, in connected between the reference voltage and the previously described level detector output. Output from this pot (wiper) drives a high impedance buffer to avoid loading effects. Defining:

V_L - Output voltage from level detector.
V_T - Reference voltage.
V_C - Buffer output voltage.
K - Fractional displacement of wiper of ratio potentiometer, refered to terminal V_T and normalized to unity pot resistance value.

We can now, assuming that the buffer has unity voltage gain:

$$V_C = V_T + K * (V_L - V_T)$$

Or, rearranging

(44) $V_C = K * V_L + (1 - K) * V_T$

Attenuation of our VCA with control voltage is:

(45) $A = \frac{V_C}{V_T}$

Replacing V_C in equation (45) with the value given by equation (44) gives:

$$A = \frac{K * V_L + (1 - K) * V_T}{V_T}$$

Wich rearranged results in:

(46) $A = K * (\frac{V_L}{V_T} - 1) + 1$

If we remember that $A = V_I/V_O$ and that V_L is at first analysis equal to $|V_{I\ peak}|$, we have, replacing for simplicity $|V_{I\ peak}|$ for V_I and $|V_{O\ peak}|$ for V_O:

$$V_O = \frac{V_I}{K * (\frac{V_I}{V_T} - 1) + 1}$$

Or rearranging:

$$(47)\quad V_O = \frac{1}{K * (\frac{1}{V_T} - \frac{1}{V_I}) + \frac{1}{V_I}}$$

So,dividing by V_T to calculate V_O level relative to V_T we have:

$$\frac{V_O}{V_T} = \frac{1}{K * (1 - \frac{V_T}{V_I}) + \frac{V_T}{V_I}}$$

Which in turn gives:

$$(48)\quad \frac{V_O}{V_T} = \frac{1}{K + \frac{V_T}{V_I} * (1-K)}$$

Expressing (48) in dB rather than in voltage ratio gives, assuming V_T = OdB reference:

$$(49)\quad V_{O\ (dB)} = -20 \log (K + 10^{-\frac{V_I\ (dB)}{20}} * (1-K))$$

Equation (49) is then our approximation of equation (9). Comparison of results is straightforward by looking to Fig.15, where equation (46) is plotted for several K settings converted to compression ratios given by (1) and for the usable gain reduction range, as compared to Fig.4, where in a similar scale and for the same values of compression ratio, equation (9) is plotted. It is obvious that aproximation of the straight lines of Fig.4 by curves of Fig.15 is very good and that in fact at intermediate ratio settings all curves tend to transform into a limiting line (growing asymptotic compression ratio defined by equation (2)). Final perfomance in terms of V_O (dB) versus V_I (dB) is pictured in Fig.16 .

This fact results in "soft-knee" compression. User can, using this curves and threshold level and ratio adjustments, implement any soft limiting function of his choise.

Finally, let's check the compression ratio pot scale, as pictured in Fig.17. It provides for detailed adjustment of low ratio settings, as required in dynamic reduction applications where input program dynamics are known, maintaining nevertheless easy adjustment at high (10:1 up) values.

As the final side chain stage, we have the link section. It consists again in a precision maximum selector, as used for the level detector previously described, fed from the output voltage of the ratio section and the output voltage of the alternate channel ratio section, available by the operation of a "link" switch. This obviously permits that if the alternate channel has a higher control voltage, it will replace at the output of the side chain the voltage of the concerned channel. In this way, perfomance described in section 2 has been achieved.

As a final note, I wish to remark that the side chain design chosen obviously permits total independance of controls and total DC control or parameters, independant of temperature and supply voltages and not relying on difficultly controlable and maintainable non-linear part functions for the best possible final accuracy. Final block diagram of side chain is shown in Fig.18.

9 CONCLUSION

A dual channel limiter/compressor/de-esser has been extensively described, using straight forward circuitry but capable of high dynamic control accuracy. It is my goal that the description and the general design approach outlined serves at the same time to help broadcast and studio engineers to fully use and understand this kind of devices, and to encourage manufacturers in designing transparent, accurate and psicoacoustically improved gain control devices. It is also hoped that OTA and VCA manufacturers can use the present information in order to improve their present products by manipulation of internal scaling constants, in order to improve dynamic range.

10 AKNOWLEDGMENT

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ACUTRON ELECTROACUSTICA, LDA.
Urb. Quinta Nova, Impasse I, Lote 134
2685 SACAVEM PORTUGAL

I wish to express here my appreciation to Acutron staff for help in experimental work and typing the final manuscript and to my wife M. Lurdes for constant encourageament all the way.

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11 APPENDIX

Derivation of time constant circuit time transfer functions:

(A) ATTACK TIME

This corresponds to a charging RC circuit of the form of Fig.19A.

Transfer function is:

$$H_{(S)} = \frac{1}{RC} * \frac{1}{S + \frac{1}{RC}}$$

Applying step function $\frac{1}{S}$ at the input results in:

$$Eo\ (S) = \frac{1}{RC} * \frac{1}{S * (S + \frac{1}{RC})}$$

Inverse transformation yelds [4]:

$$Eo\ (t) = 1 - e^{\frac{-t}{RC}}$$

(B) RELEASE TIME

This corresponds to a discharging RC circuit of the form of Fig.19B.

Which is similar in behaviour to an RC circuit of the form of Fig.19C to which is applied the step function previously described.

Which transfer function is:

$$H\ (S) = \frac{S}{S + \frac{1}{RC}}$$

Applying again step function $\frac{1}{S}$ at the input:

$$Eo\ (S) = \frac{1}{S + \frac{1}{RC}}$$

And inverse transform is [1]

$$Eo\ (t) = e^{\frac{-t}{RC}}$$

Feedback configuration

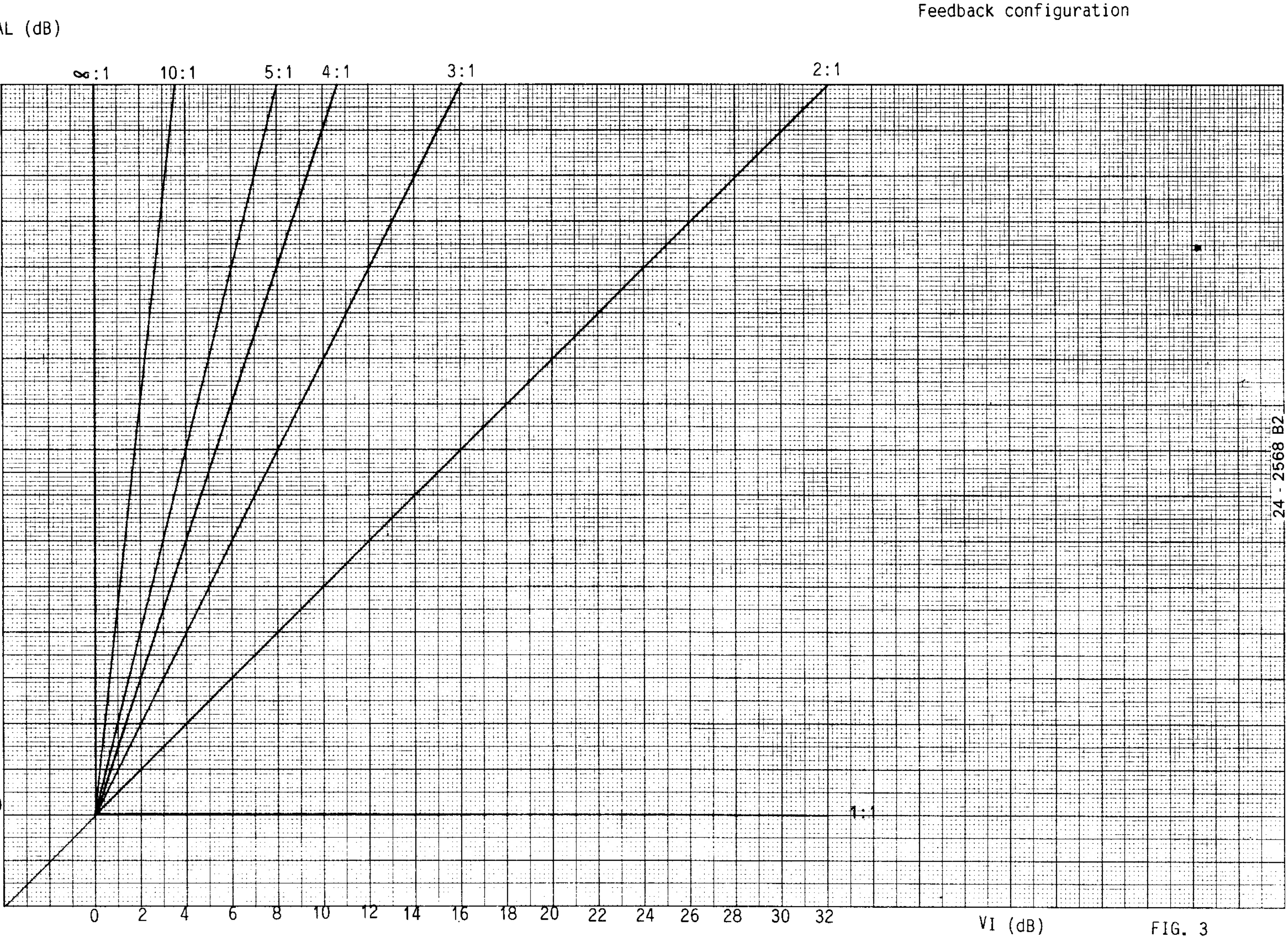

FIG. 3

Feedforward configuration

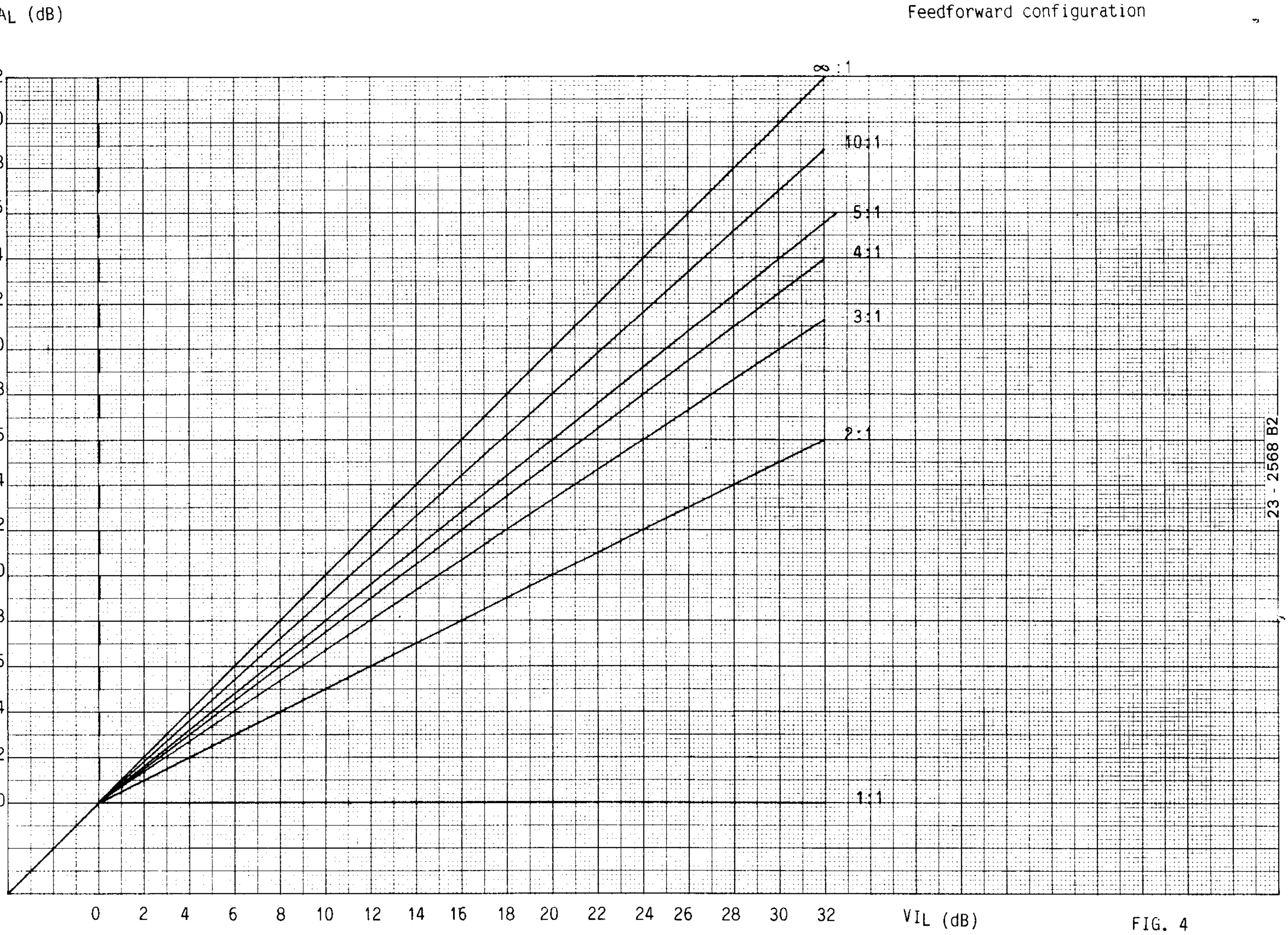

FIG. 4

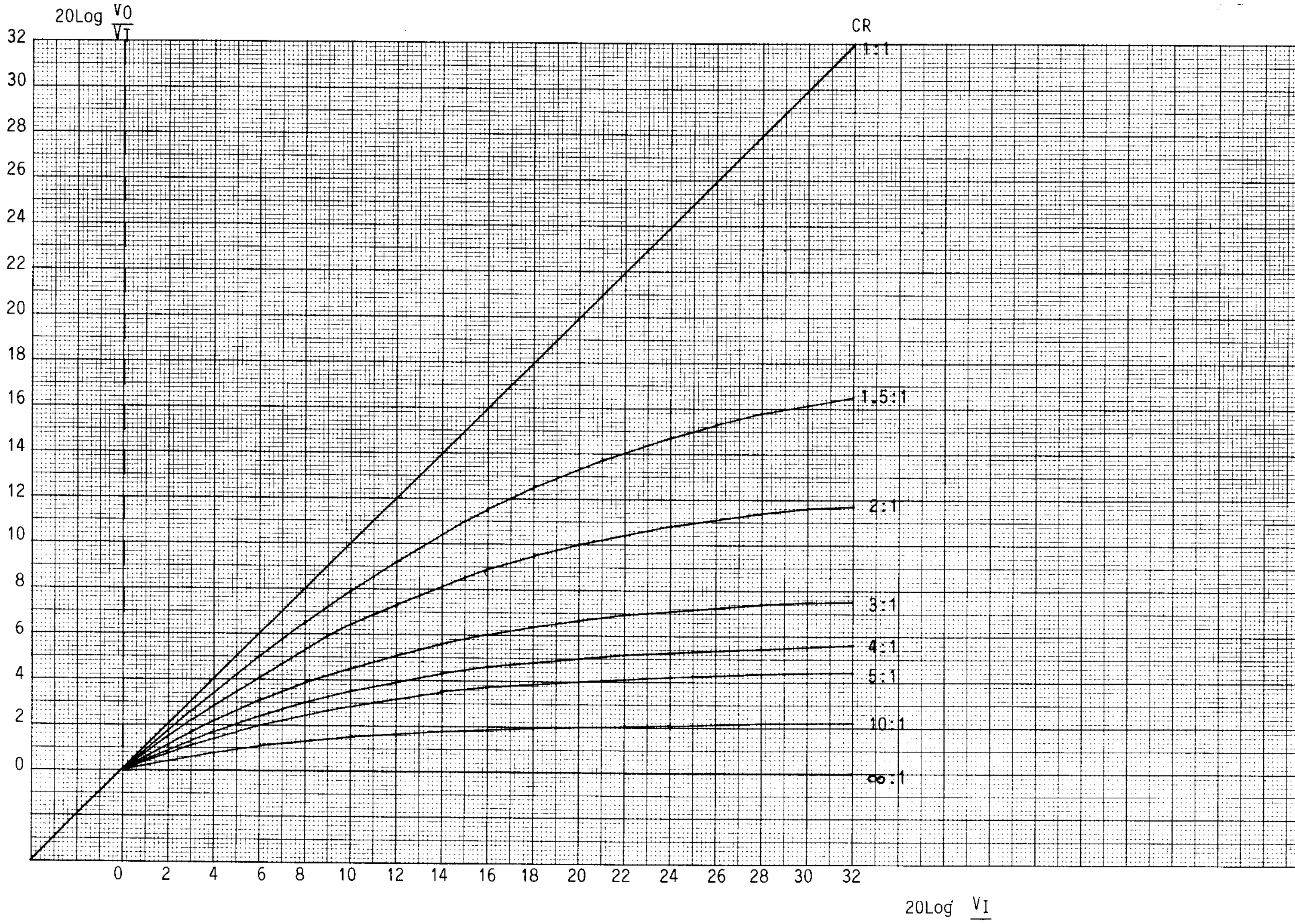

FIG.16

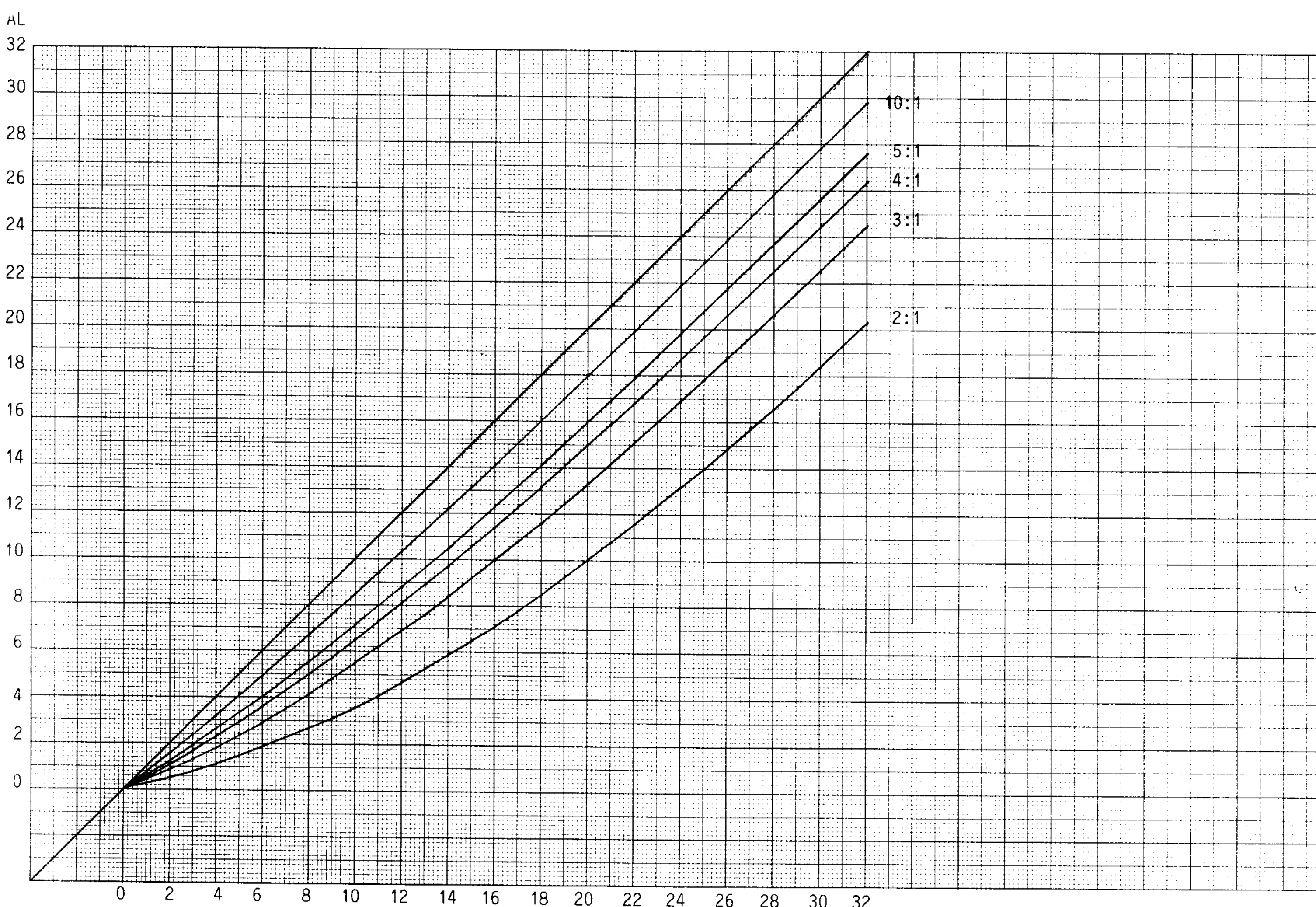

FIG. 15

Final attenuation versus input leve

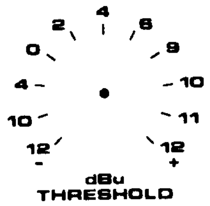

FIG.10

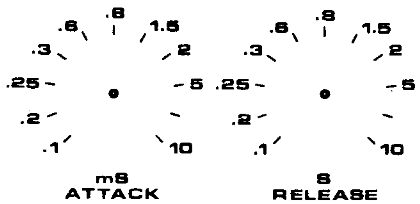

FIG.12

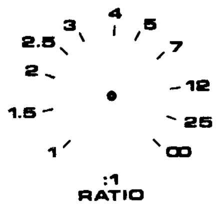

FIG.17

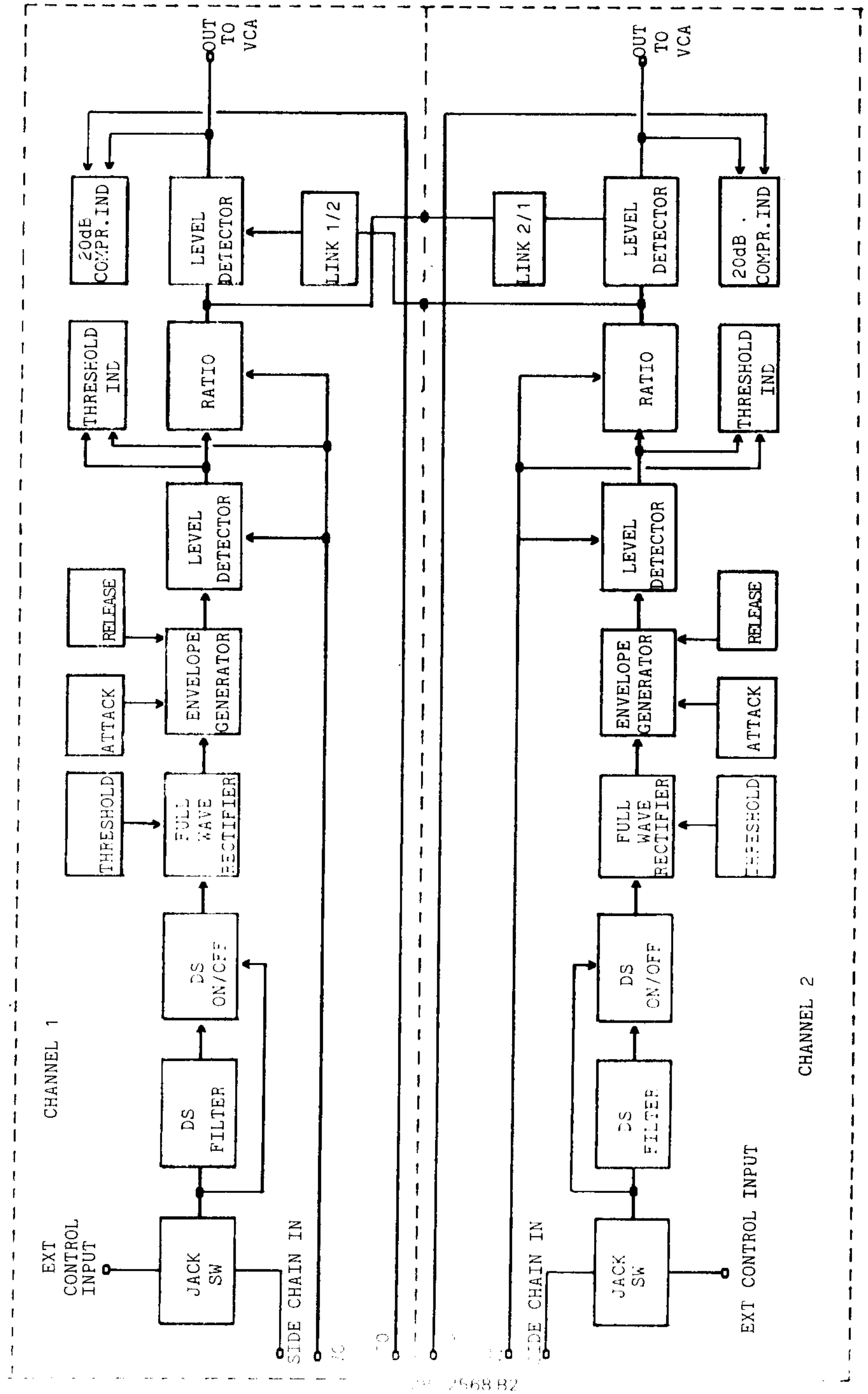

FIG.18

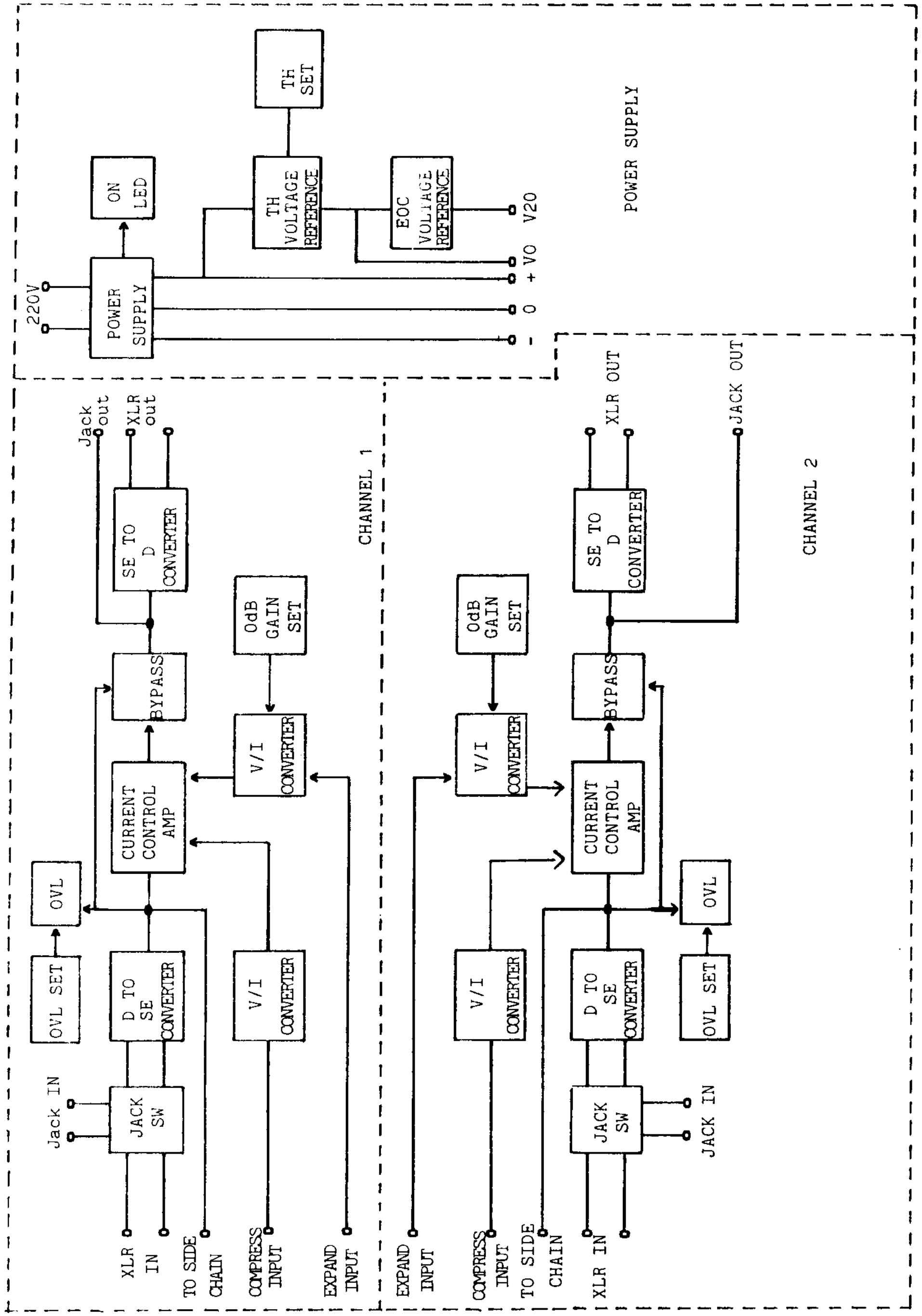

FIG 8 29-2568 B2

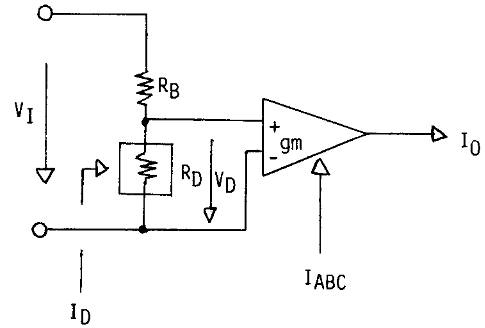

FIG.5

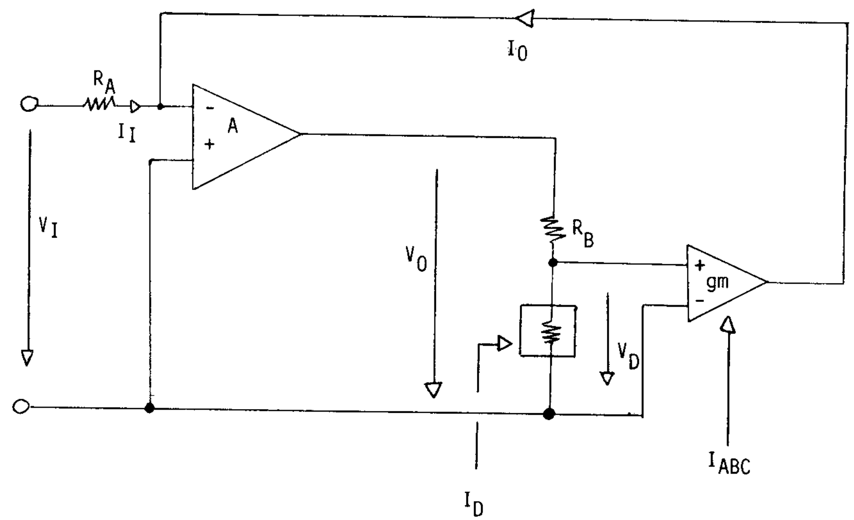

FIG.6

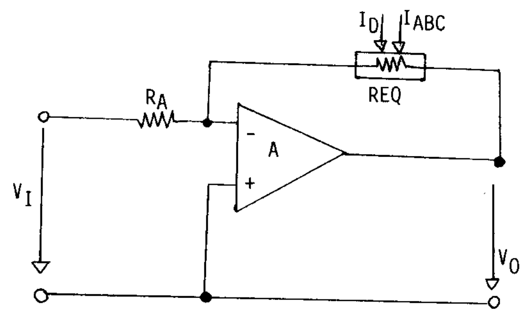

FIG.7

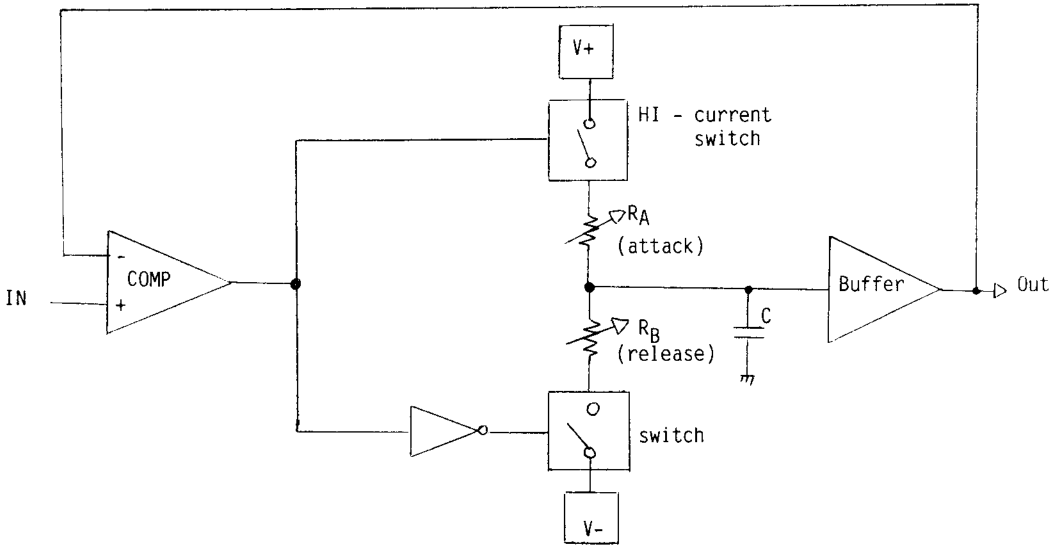

FIG.11

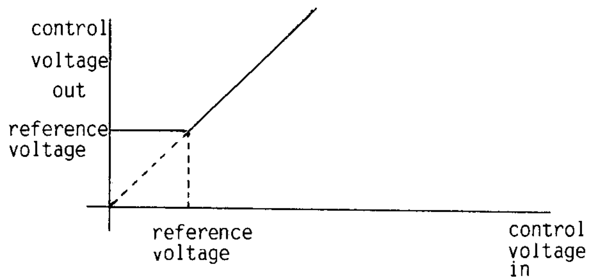

FIG.13

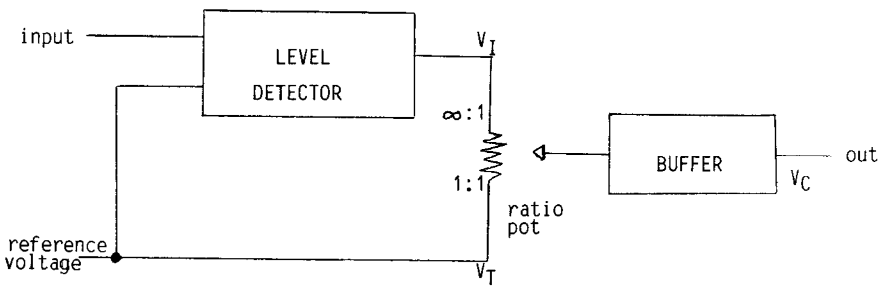

FIG.14 31-2568 B2

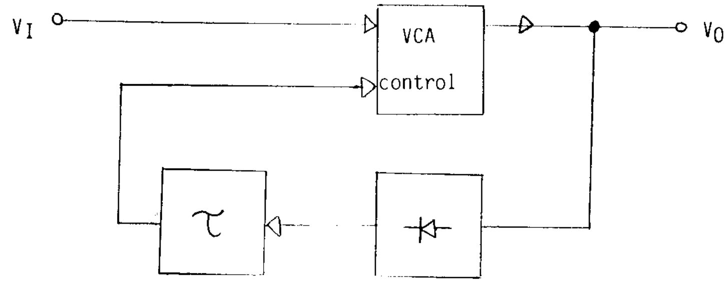

FIG.1

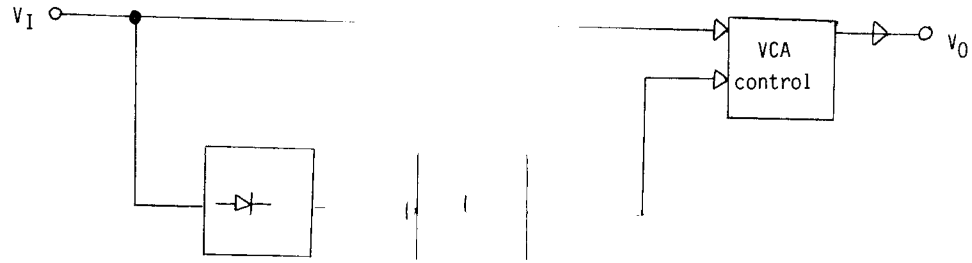

FIG.2

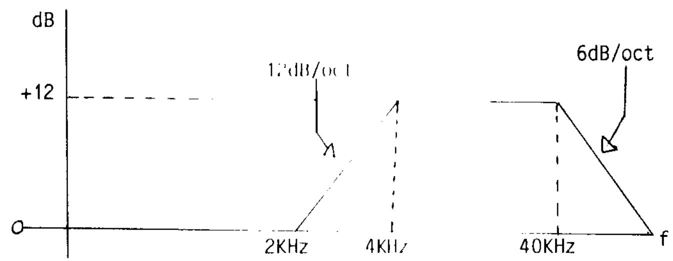

FIG.9

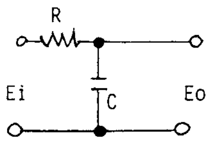

FIG.19A

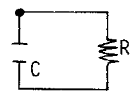

FIG.19B

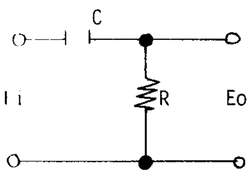

FIG.19C