# A Linear high-frequency AC voltmeter and AM detector

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## Abstract

The final-stage collector current of a push-pull complementary-symmetry amplifier is a half-wave rectified version of the signal being amplified. By biasing the amplifier in class C, and using a large amount of negative feedback, the rectification process can be made almost-perfectly linear. This provides the basis for a high-quality AM detector, or a high-frequency linear AC voltmeter. Both functions can also be realised simultaneously, with different time-constants, because there are two collectors from which signals can be extracted.


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## Acknowledgement

I would like to thank Dr John Ponsonby for pointing out an error in circuit analysis in the original (version 1) document.


1 Information partly derived from technical documents produced by the author (DWK) in 1981 and updated in 1987. Circuit description first made available via the internet in 2000. Introduction added in 2009, with minor corrections in 2013 and 2014 . Version 2.00 (March 2015) is the first major revision, with corrections to the circuit analysis, and discussion of design considerations, detector time-constants, and audio bandwidth.

## Introduction

The circuit described in this article was originally developed for use in an ultrasonic receiver. Such instruments are used, in conjunction with a piezoelectric transducer, for applications such as; the detection of bearing wear in electric motors; the detection of insulator breakdown in electricity substations; and for finding leaks in pipes and pressure vessels. Another use is in compensating for tool wear in automatic manufacturing equipment, by determining the point at which the cutter or grinding-wheel makes contact with the work-piece.

Industrial ultrasonic receivers require a linear voltmeter, sometimes used to trigger a level detector, for engineer call-out; and an AM detector, so that the engineer can use headphones and a directional transducer to find the source of the noise. The idea that both requirements might be combined in a single circuit came about because another design project, with which the author had been engaged shortly beforehand, had involved push-pull audio amplifiers; and the relationship between collector current and output current had been flagged as the basis of a possible rectification method.

The original ultrasonic receiver was designed to work at 40 kHz. A frequency-response measurement however showed that the new circuit worked perfectly well at more than 1 MHz. Hence the detector was also suitable for connection to the output of a superhet IF amplifier. The reason for wanting to do that relates to the problem of listening fatigue.

The half-wave germanium-diode detector is ubiquitous in broadcast receivers; but the translation from modulation level to output is seriously non-linear (and earlier valve / tube detectors are just as bad). The trick in getting acceptable audio quality is to drive the detector hard, but there is no truly linear region in the diode characteristic, and signals too weak to drive the AGC full-on will be detected close to the diode threshold. This does not bode well for the audio quality; and certainly, no one hearing a voice on AM radio will be fooled into thinking that the person is in the room. Still, everyone knows what AM radio sounds like, and so there is no great pressure to do anything about it.

The discovery that the new detector would work at superhet intermediate frequencies however provoked a certain curiosity. Also on hand was a Racal RA17C18, which is a 500 kHz to 30 MHz fax receiver with a 13 kHz maximum IF bandwidth, a linear IF phase response, and an IF output socket on the back. Hence, another version of the circuit was built (with different design considerations resulting in different component values) and connected as the interface between the RA17 and a hi-fi amplifier.

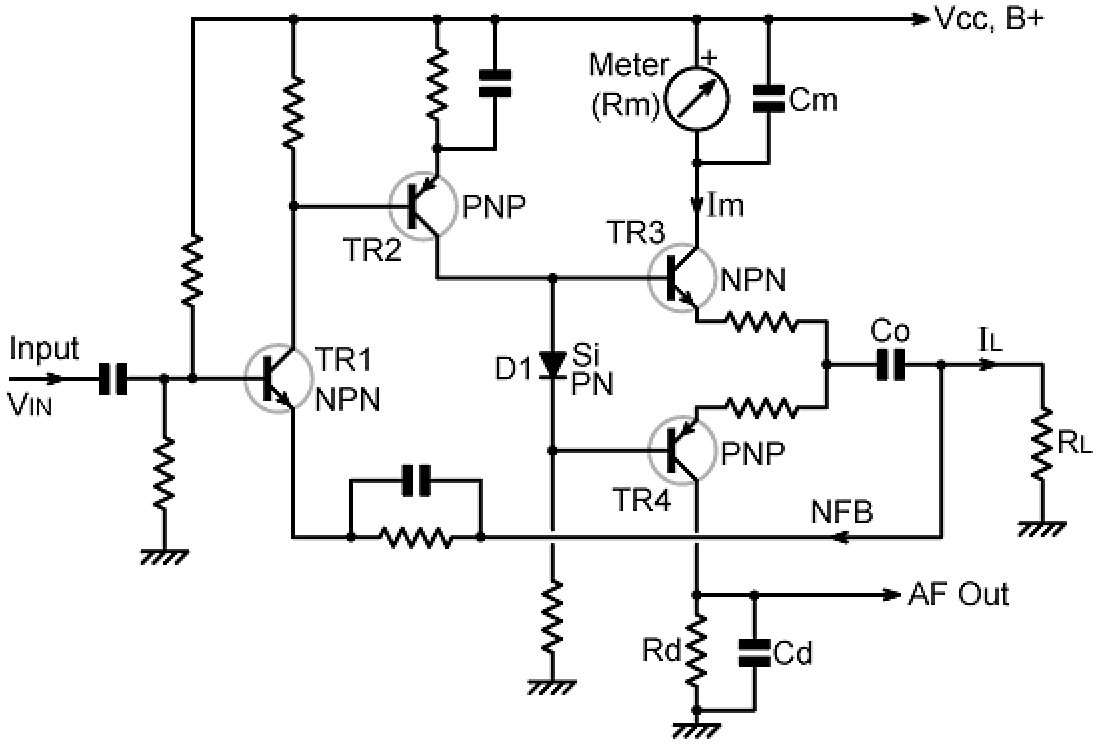
The obvious preconception with regard to broadcast AM is that the quality can never be any good, because the maximum usable audio frequency component is less than 9 kHz in the daytime and 4.5 kHz at night (according to the European band plan). In fact, the extreme treble content is not that important to the pleasure of listening, but the absence of high-order distortion products certainly is. It transpires that a great many broadcast transmitters do produce high-quality audio within the statutory bandwidth restriction. The sound of the linear detector on tuning around is also clean, and easy on the ear; and the intelligibility of weak stations is greatly enhanced. It is not much of a conjecture to say that the brain must do a considerable amount of processing to extract the intelligence from distorted signals; and the removal of the audio intermodulation products is like the lifting of a veil.

A linear rectifier is, of course, not the only solution to the problem of high-quality AM reception. Another way to do it is to phase-lock an oscillator to the carrier and use synchronous demodulation. Such circuits are complicated however, and jump disconcertingly from carrier to carrier when tuning across a band. The linear detector described here is simple (4 transistors), uses a single power supply, and just behaves like an ideal rectifier.

## Circuit description

Simple rectifier-type voltmeters require relatively-large inputs in order to overcome the effect of diode forward-threshold on linearity. A bridge formed with germanium or silicon Schottky diodes will provide a reasonably linear display with 2 V RMS input, but it will barely register at all until the input exceeds about 0.2 V P-P. Thus, using a conventional diode bridge to measure an input signal of (say) 1 μV requires a voltage gain of 126 dB in order to achieve modest linearity. Partial forward-biasing of the diodes can be used to improve matters, but such arrangements are prone to thermal drift. There are, of course, ways of achieving linear detection other than simple rectification, but the well-known methods have drawbacks. Conventional precision rectifier circuits, for example, have a restricted frequency range; and post detection linearity compensation methods are not successful for small inputs because they do not account for dynamic effects[2].

In circuit shown below , the signal is fed to a complementary symmetrical push-pull amplifier, with a feedback factor of 1, which delivers power to a load resistor $R_L$. A moving-coil meter, or some other indicating device having an input resistance (Rm) of the order of 1 kΩ, is placed in series with the collector of TR3.




The first two amplifying stages need to provide an open-loop voltage gain somewhat in excess of $10^3$, but this is not difficult to achieve using small-signal transistors. The output stage is biased for zero quiescent current (class B/C) because the voltage across D1 is insufficient to cause significant static conduction in the output transistors. The large feedback ratio forces the system into linearity, and reduces cross-over distortion to a negligible amount for the purposes of this application. Thus, assuming that the reactance of the output coupling capacitor, $C_O$ , is very small in comparison to $R_L$, the voltage appearing across $R_L$ is the same as the input voltage. Hence:

$I_L = V_{IN} / R_L$

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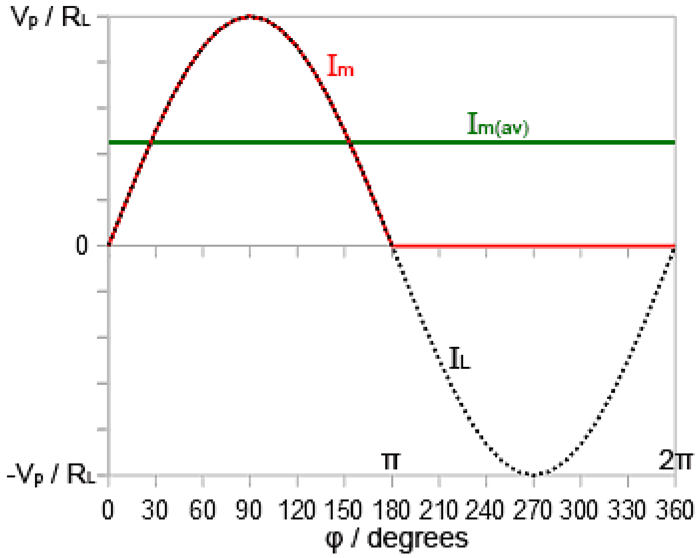
2 This is discussed in the author's article '**Diode detectors for RF measurement**', http://www.g3ynh.info/circuits/diode_det/index.html

When the circuit is driven by an AC waveform, TR3 drives the load current positive, and TR4 drives the load current negative. Hence the current flowing in TR3 collector is equal to the half-wave rectified load current. The meter however, due to inertia and the effect of the smoothing capacitor (Cm) will register the mean value. This is obtained by integrating the current over one complete cycle. Thus, if we call the time-varying phase-angle of the input waveform $\varphi$ (where $\varphi = 2\pi ft$ ) and analyse for a sinusoidal component with an upward zero-crossing at $t = 0$ , we get:




$V_{IN} = V_p \,\mathrm{Sin}\varphi$

where the peak voltage, $V_p$ , is given by:

$V_p = V_{IN\,(rms)} \times \sqrt{2}$

For the transistor supplying the positive half-cycle (TR3), conduction only takes place between 0 and 180° (0 and $\pi$ radians). Hence the meter reading, $I_{m\,(av)}$ , is given by:

$$I_{m\,(av)} = \frac{1}{2\pi}\int_0^\pi I_L \, d\varphi \quad = \frac{1}{2\pi}\int_0^\pi \frac{V_p}{R_L} \mathrm{Sin}\varphi \, d\varphi \qquad \text{where } \int \mathrm{Sin}\varphi \, d\varphi = -\cos\varphi + c$$

Thus:

$$I_{m\,(av)} = \frac{-1 \quad V_p}{2\pi \quad R_L}\Big[\cos\varphi\Big]_0^\pi \quad = \frac{-1 \quad V_p}{2\pi \quad R_L}[\,-1\ -1\,] = \frac{V_p}{\pi\, R_L}$$

Substituting for $V_p$ we get:

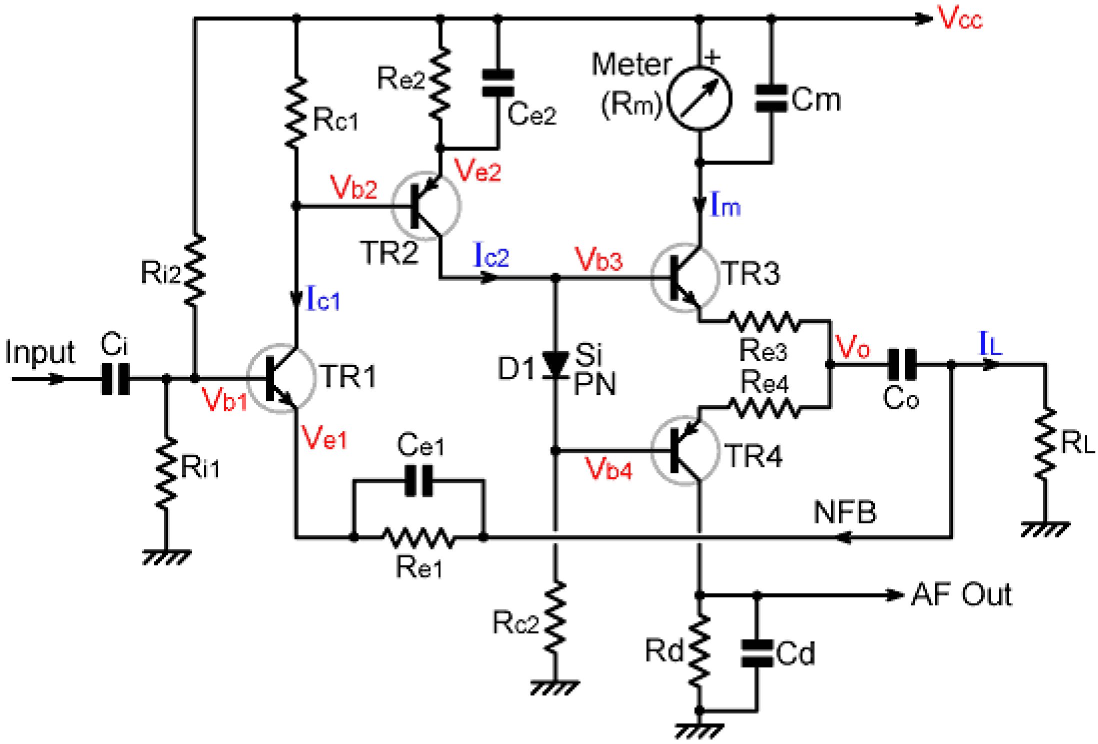
$$\boxed{I_{m\,(av)} = \frac{(\sqrt{2}) \quad V_{IN\,(rms)}}{\pi \quad R_L}} \qquad (1)$$

The meter sensitivity is thus defined by the choice of load resistor, and the scale is almost perfectly linear provided that the open-loop voltage gain is high. If, for example, the meter is chosen to have a full-scale deflection (FSD) of 100 μA, and the load resistor $R_L$ is 100 Ω, then FSD is obtained with a detector input of 22.1 mV RMS. It then requires only 87 dB of amplification prior to the detector to obtain FSD for a 1 μV signal.

Note that the meter and audio take-off points can be transposed. In mains powered equipment, where there may be noise on the power rail, the audio take-off point closest to ground is to be preferred. For a digital meter however, or input to an A-D converter, it will be necessary to use the lower (ground-referenced) output for signal level and take audio from TR3 collector.

## Basic design considerations

The detector can be built using four small-signal silicon transistors[3]. The choice is not critical provided that $h_{FE}$ comfortably exceeds 100 and $f_T$ exceeds about 100 MHz. The circuit does not need to produce a large power output, and so TR1 can be the same as TR3, and TR2 can be the same as TR4. The only critical component is $R_L$, which sets the gain and should have minimal parasitic reactance.




There is a great deal of choice in establishing the circuit DC conditions, and so the discussion to follow is merely for guidance. We might, for example, decide to use a 100 µA FSD meter with a resistance ($R_m$) of about 1 kΩ in series with TR3 collector. Then, for want of a good reason to do otherwise, we can make the resistance in TR4 collector ($R_d$) 1 kΩ as well[4]. With this symmetrical choice, best headroom is obtained if we make the output voltage $V_O$ to be half the supply voltage $V_{cc}$. Now note that, if we choose the collector current of TR2 ($I_{c2}$) to be somewhere in the range from 100 µA to 1 mA, the voltage across D1 will be about 0.6 V if a silicon PN signal diode such as a 1N4148 is used. Thus we establish the approximate static base voltages of TR3 and TR4:

$V_{b4} = (V_{cc} / 2) - 0.3$ and $V_{b3} = (V_{cc} / 2) + 0.3$

DC conditions can now be established by choosing a supply voltage and the collector currents for the transistors TR1 and TR2. Here, for the sake of illustration, we will assume a nominal rail voltage of +12 V. Also, in the absence of a low-power requirement, best RF performance is obtained if circuit impedances are kept reasonably low, and so we will set the collector currents of both TR1 and TR2 to about 1 mA.

3 Some examples of complementary small-signal transistors are: BC107B (NPN) + BC177B(PNP); BC183 (NPN) + BC213 (PNP); BC549 (PNP) + BC559 (NPN). A discussion of these devices is given at: http://en.wikipedia.org/wiki/BC548

4 $R_d$ also determines the audio output level relative to meter FSD (for a given percentage modulation), but the circuit is not noisy if the power-rail is clean, and so the detected audio signal can be amplified as required.

With a supply voltage of 12 V, we have $V_{b4}$ = 5.7 V. If $I_{c2}$ is 1 mA, then $R_{c2}$ = 5.7 kΩ. The nearest preferred value in the 5% tolerance series is 5.6 kΩ.

Now observe that in the choice of the TR2 emitter resistor, $R_{e2}$, there is a trade-off between stable biasing and signal headroom. The signal peak-to-peak amplitude however will be very small (a few 10s of mV) if a sensible choice is made for $R_L$ (say 50 Ω to 100 Ω ). Hence we can easily use about half of the available headroom in setting the voltage across $R_{e2}$. Also note that the voltage $V_{b2}$ will be about 0.6 V below the voltage $V_{e2}$. Therefore, since we are also going to set the collector current of TR1 to be 1 mA, we can choose preferred value resistors that have a difference of 0.6 kΩ for $R_{c1}$ and $R_{e2}$. Thus $R_{e2}$ = 2.7 kΩ and $R_{c1}$ = 3.3 kΩ. The voltage $V_{e2}$ is now $V_{cc}$ - 2.7 = 9.3 V , and the voltage $V_{b2}$ is now $V_{cc}$ - 3.3 = 8.7 V . (It is assumed that all transistors have large $h_{FE}$ for the purpose of this analysis, and so we ignore base current. The biasing errors generated by this approximation are negligible).

For stable biasing of TR1, maintaining the emitter at about 3 V is a reasonable choice. We can therefore choose $R_{e1}$ = 3.3 kΩ. Now note that TR1 gets its emitter ground return via the load resistor $R_L$. This is done to ensure that there are no impedances in the path from the load to the negative feedback input, so that the load voltage follows the detector input voltage accurately. If we choose $R_L$ to be be ≤ 100 Ω however, then this is < 5% of the value of $R_{e1}$, and we can ignore the small DC voltage that appears across it[5]. The voltage $V_{e1}$ is therefore approximately 3.3 V, and so the voltage $V_{b1}$ will be approximately 3.9 V. Now all we have to do to complete the biasing arrangement is to choose the input resistors to form a potential divider giving 3.9 V. If we choose the potential divider current to be (say) 100 µA, then the total resistance will be 120 k Ω. Thus $R_{i1}$ = 39 kΩ and $R_{i2}$ = (120 - 39) kΩ = 81 kΩ. Using the nearest preferred value of 82 kΩ for $R_{i2}$ is of course acceptable.

Note that, due to the 100% negative feedback, the input transistor acts as a bootstrap follower. Therefore the input resistance is essentially determined by the parallel combination of $R_{i1}$ and $R_{i2}$ (parasitics and circuit-board strays will also add some parallel capacitance). Using the resistor values calculated above, the input impedance will be somewhat high for an RF circuit, but the impedance can be reduced in various ways. We might, for example, increase the potential divider current so that smaller resistor values are required. This however, will increase the power consumption and is therefore not a good idea for battery-powered circuits. Alternatively we can place a load resistor or a low-value calibration pot. on the AC side of the input.

The resistors $R_{e3}$ and $R_{e4}$ are included to protect the output from short-circuits. A value of (say) 10 Ω will protect the transistors against momentary shorts during test and measurement, whereas a value of around 100 Ω will give protection for long periods. The resistors might be omitted altogether, but should be included if it is intended, for example, to connect additional networks to the load terminal, or to connect the load to the detector via a socket. Note that if the load is disconnected, the circuit will be switched off except for the input biasing network. If intending to allow load disconnection, check that the collector-base reverse breakdown voltage ($V_{CBO}$) for TR1 is sufficient to withstand the supply voltage.

The coupling capacitor $C_i$ is chosen to have a very small reactance magnitude in comparison to the input resistance. If we use the values calculated above, we have: $R_{in}$ = 39 k // 82 k = 26.4 kΩ . If the operating frequency is (say) 470 kHz, and we use a 10 nF ceramic (i.e., non-inductive) coupling capacitor, its reactance will be.

5 Larger values of $R_L$ will have to be added to $R_{e1}$ when calculating $V_{e1}$.

$X_{Ci} = -1 / 2\pi f\, C_i = -33.9\ \Omega$

For the load coupling capacitor, we might use (say) a 10 µF tantalum-bead (non-inductive) electrolytic. Thus:

$X_{CO} = -1 / 2\pi f\, C_O = -0.034\ \Omega$

The emitter decoupling capacitors should have reactance magnitude that is small relative to the resistors they shunt. A 0.1 µF ceramic capacitor, for example, has a reactance at 470 kHz of -3.4 Ω . The detector and meter load shunt capacitors however control the audio frequency-response and the meter damping. It is therefore not sufficient simply to make them 'large', and they will be discussed in detail in the next section.

## Detector time-constants

The output-transistor collector load resistances (one of them being a meter), and their associated smoothing capacitors, are equivalent to the load placed across the output of a conventional diode detector. When the output device is a meter, the RC time-constant sets the meter damping. In the case of an audio output, the RC time-constant dictates the audio bandwidth. Note that, for the push-pull detector, the collector loads do not need to be identical. The two time-constants can be greatly different, allowing wide audio bandwidth to be obtained in conjunction with a heavily-damped meter response (if so desired). If one of the outputs is not required moreover, the collector load can be replaced by a short-circuit.

### Meter damping

When a capacitor is discharged via a resistor, the voltage decay law is given by:

$V(t) = V_0 \exp\{-t/CR\}$

where $\exp\{x\} = e^x$, $V_0$ is the starting voltage, and $V(t)$ is the voltage after time t.

The product of the resistance and capacitance, CR, is known as the 'time constant'. When t = CR:

$V(t) = V_0\, e^{-1} = 0.368\, V_0$

The time constant is thus the time it takes for the voltage across a resistor in parallel with a capacitor to decay to 37% of it starting value. From this we can obtain an idea of how the needle of a panel meter will behave for different values of damping capacitance[6].

If we have a meter with a resistance of 1 kΩ and place a 0.1 µF ceramic capacitor across it, the level of RF signal present will be greatly attenuated. The time constant however, will be 100 µs, and the only significant damping achieved will be due to mechanical inertia and the inductance of the meter coil. This is acceptable when measuring slowly changing signals, and desirable when trying to detect transient events or null a bridge; but if the signal is unsteady, it is often preferable to obtain an average reading. Thus, if we were to place a 100 µF capacitor across the meter, the time

6 This discussion neglects magnetic damping, which is due to the tendency of the collapsing magnetic field to oppose the current-decay in the meter coil.

constant would increase to 0.1 s; which would have a noticeable steadying effect on the meter reading. It would also help to prevent meter damage during extreme overload, by reducing needle acceleration. Increasing the capacitance to 1000 μF would give a time constant of 1 s, which would result in a fairly slow response. It is often useful to provide a switch, giving a choice between fast and slow meter responses.

**Audio bandwidth and de-emphasis**

The detector time constant is also a parameter that controls the audio frequency response. This is because the half-power (-3dB) point, or 'corner frequency' occurs when the magnitude of the capacitor reactance is equal to the resistance. Thus we have:

$R = | -1 / 2\pi f_{-3dB} C |$

so that:

$f_{-3dB} = 1 / 2\pi RC$

Below is a table showing the corner frequencies for various AM detector time constants:

| Time const. | $f_{-3dB}$ | Notes |
|---|---|---|
| 10 μs | 15.9 kHz | |
| 15 μs | 10.6 kHz | |
| 20 μs | 7.96 kHz | |
| 25 μs | 6.37 kHz | |
| 50 μs | 3.18 kHz | Communications bandwidth |
| 75 μs | 2.12 kHz | NRSC de-emphasis |
| 100 μs | 1.59 kHz | |

From the table, it can be seen that a suitable choice of time constant for communications receivers is 50 μs. This was certainly the author's practice many years ago; but it perhaps subscribes to the misconception that, just because 3 kHz bandwidth is sufficient to convey the intelligence of human speech, then there is nothing to be gained by retaining frequencies greater than 3 kHz. That is obviously not true, and so it might be better to choose 20 μs, or even 15 μs, and use a variable IF bandwidth to cut the high audio-frequencies in the event of adjacent-channel interference.

For AM broadcast reception, it might seem that a time-constant shorter than that used for communications traffic is required. That is not the case however, because broadcasters usually use pre-emphasis. Effectively, the broadcaster boosts high-frequencies by using a parallel CR network in series with the audio signal, so that a reduction in receiver noise can be achieved by using a CR network in parallel with the audio signal. Hence the detector time constant should be selected to cancel the transmitter pre-emphasis and restore a flat response. Unfortunately however, a universally accepted standard for AM broadcast pre-emphasis is hard to find. The NRSC recommendation[7] is that receivers for AM broadcast should have a de-emphasis time constant of 75 μs, with an audio notch filter at the channel spacing frequency (10 kHz for the Americas, 9 kHz for Europe). AM broadcast radio receivers used in Europe usually have a detector time constant in the 50 μs to 100 μs range, but most writers on receiver design seem to think that the capacitor serves only to filter-out the IF signal. The apparently arbitrary choice of time constant therefore

7 NRSC-R10 AM pre-emphasis standards, 1986. http://www.nrscstandards.org/reports.asp

seems to emanate from a belief that the smoothing capacitor can always be 10 nF regardless of the parallel resistance.

Inability to determine a universal choice of time-constant might lead the designer to contemplate providing a number of choices via a rotary switch. An alternative however is to use a short time-constant (say 20 μs), with some high-order filtering to get rid of the IF if necessary, and then provide a tone control further along the audio signal chain.

## Providing a meter scale in dB

Since the meter reading is linear, the problem of diode-law correction is eliminated and it is a straightforward matter to generate artwork to produce a meter scale in dB. If the reading in dB is N, we have:

$$N = 20 \, \mathrm{Log}_{10}\{ V / V_{ref} \} \quad [\mathrm{dB}]$$

and

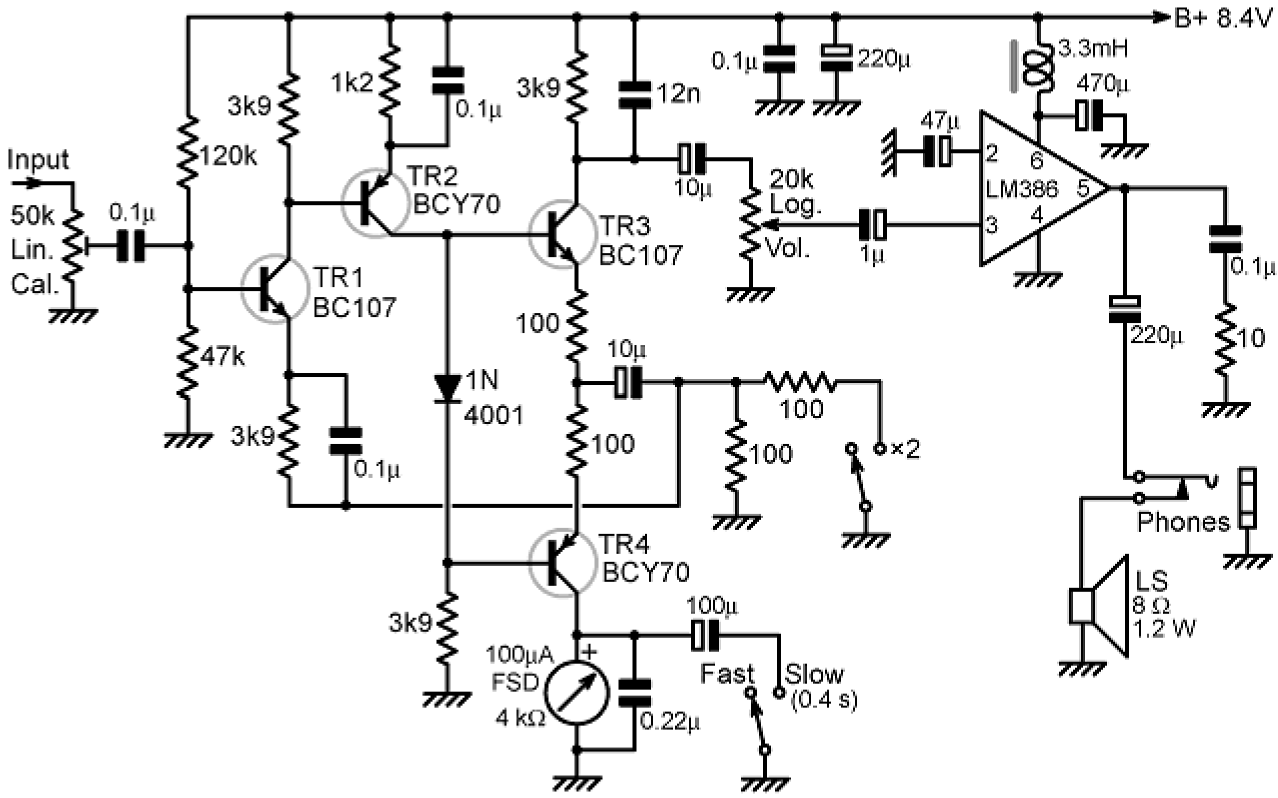
$$V / V_{ref} = 10^{N/20}$$

where $V/V_{ref}$ is the proportion of the full-scale reading that corresponds to N. Thus, if we decide (say) to make FSD correspond to +3dB, the scale comes out as follows:

| dB scale | dB below FSD | $V / V_{ref}$ |
|---|---|---|
| +3 | 0 | 1.0000 |
| +2 | -1 | 0.8913 |
| +1 | -2 | 0.7943 |
| 0 | -3 | 0.7079 |
| -1 | -4 | 0.6310 |
| -2 | -5 | 0.5623 |
| -3 | -6 | 0.5012 |
| -4 | -7 | 0.4467 |
| -5 | -8 | 0.3981 |
| -6 | -9 | 0.3548 |
| -7 | -10 | 0.3162 |
| -8 | -11 | 0.2818 |
| -9 | -12 | 0.2512 |
| -10 | -13 | 0.2239 |
| -15 | -18 | 0.1259 |
| -20 | -23 | 0.0708 |

## Example circuits

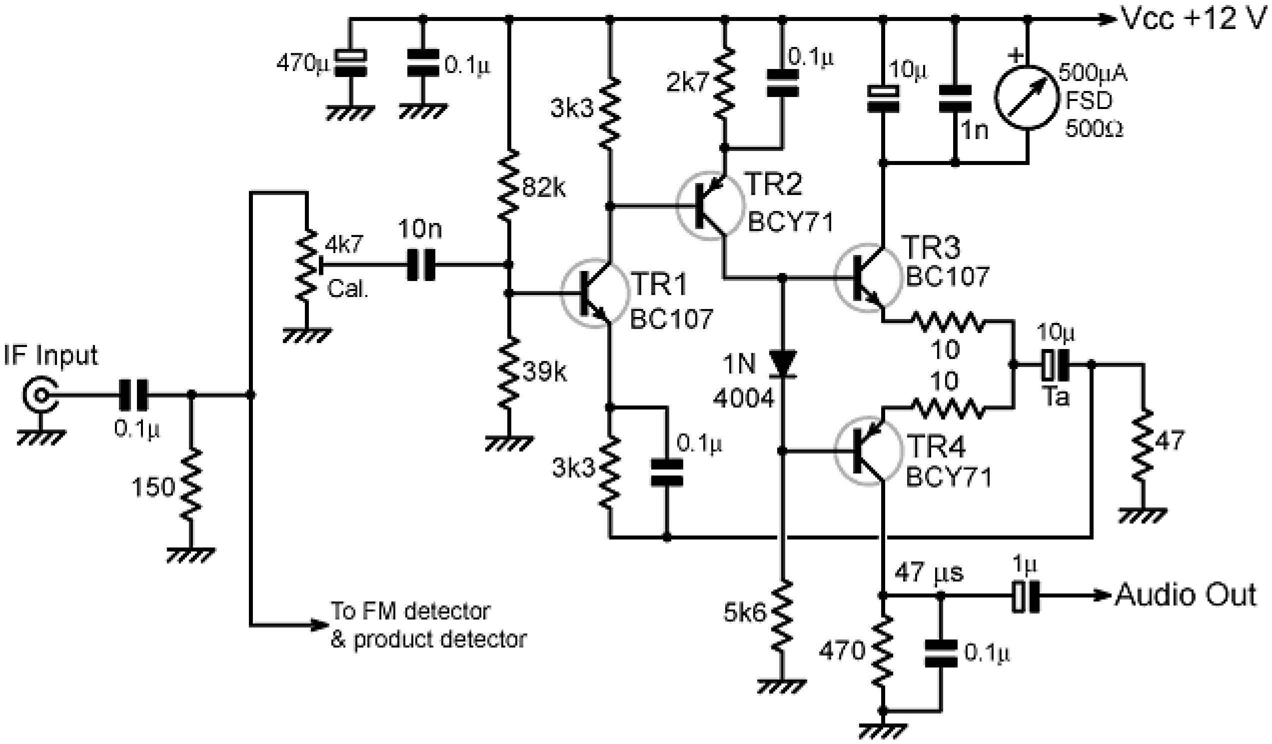
### Battery-powered AC voltmeter and detector with audio amplifier




In the circuit above, the load resistor is 100 Ω, and the meter movement is 100 µA FSD, giving a sensitivity of 22 mV RMS FSD at the wiper of the input calibration potentiometer. The circuit also has a ×2 (+6 dB) scale expansion switch (giving 11 mV RMS FSD), implemented by reducing the load resistor to 50 Ω. The LM386 audio amplifier provides 26 dB (×20) voltage gain. The inductor in the LM386 supply feed prevents RF distortion products from the audio amplification process from appearing on the supply rail. Note that this is a battery powered circuit, and so the the audio is taken from TR3 and the meter is in series with TR4. This arrangement was chosen to give the lowest possible level of spurious radiation from the meter wiring, the detector being part of a high gain ultrasonic receiver[8] with the meter mounted close to the input socket. In a mains powered instrument, the audio should be taken from TR4, since that is the arrangement giving best supply (mains hum) rejection in the audio output. The ultrasonic receiver was intended to operate with an input at 40 kHz. The detector however, when built with wire-ended components but a compact layout, gave a response that was was flat within 0.5 dB up to 1.1 MHz, and -3 dB at 1.6 MHz.

8 Detectaids 900 series ultrasonic receivers. D W Knight. http://www.g3ynh.info/circuits/detectaids/DT900ser.pdf

**AM detector for general-coverage radio receiver**



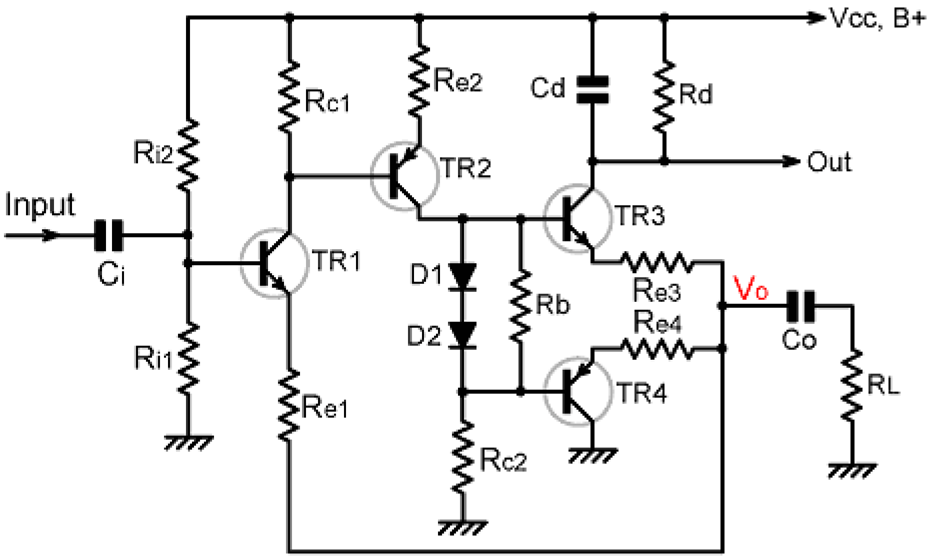

This is the circuit of the AM detector used in an all-mode (AM, FM, SSB, CW) adapter[9] built by the author for use with the Racal RA17. The IF of the RA17 is 100 kHz; but as before, the frequency response of the detector extends well beyond 1 MHz, and so the circuit will work perfectly well at 470 kHz. Note that the meter is largely redundant in radio receiver applications, because absolute signal strength is best related to the AGC voltage. It can be replaced with a short-circuit if not required; but it is useful for making accurate comparative measurements of RF level when the AGC is switched off. The sensitivity pot. (cal.) at the input is used to make the subjective audio volume reasonably constant on switching between demodulation modes.

9 http://www.g3ynh.info/Racal/RA17adap/RA17adp2.html

## Further development

The original circuit was very much a child of its time; and once the prototypes were found to work, it did occur to the author that some simplification ought to be possible, particularly the elimination of the emitter bypass capacitors. This was not done at the time however, firstly because the associated projects were finished, and secondly, because possible variants were not quite so unequivocally linear.

The most promising change is to take the emitter of TR1 directly to the junction of the output transistor emitters. This allows the biasing to be controlled by the feedback loop, permitting either the elimination of the TR1 and TR2 emitter resistors, or at least their reduction to the point at which they no longer need bypassing. The circuit below shows how the idea might be implemented.




The emitter resistors $R_{e1}$ and $R_{e2}$ can be of low value, or even short-circuited. The input resistors $R_{i1}$ and $R_{i2}$ are chosen so that the base of TR1 is held at $V_{cc}/2 + 0.6$ V. The DC output voltage $V_O$ is then $V_{cc}/2$, because TR4 pulls down the emitter of TR1 and turns it on, then TR1 turns on TR2, which pulls-up the base of TR4 until the system is in equilibrium. Note that this inevitably leads to some quiescent current in TR4, and so there will be some DC offset if the collector of TR4 is used as a detector output.

TR3 however does not need to have any standing current. If we use two silicon PN diodes for D1 and D2, TR3 base current can be controlled by adjusting $R_b$. If the biasing of TR3 is set so that the transistor is just at the threshold of conduction (as can be determined by measuring the voltage across $R_d$) then TR3 (rather than TR1) will supply most of the current in $R_L$ during positive half-cycles, and linear detection without significant zero-offset will result.

Note that either a meter or a resistive load can be used in the collector of TR3. Also, although shown connected directly to ground, the use of TR4 collector as an output is not ruled out. If $R_{c1}$ is relatively large, the standing current through TR4 will be small. A small offset at the beginning of an otherwise linear response will have no noticeable effect on the linearity of an AC coupled audio output.

An approach very similar to that described above has been used by Robert Batey for the design of a high-fidelity AM broadcast detector[10]. Robert's design also includes a notch filter to get rid of adjacent-channel carrier heterodyne interference, and the DC level at the audio terminal is used for overall receiver AGC. Using a compact layout, his circuit works with input frequencies in excess of 5 MHz.

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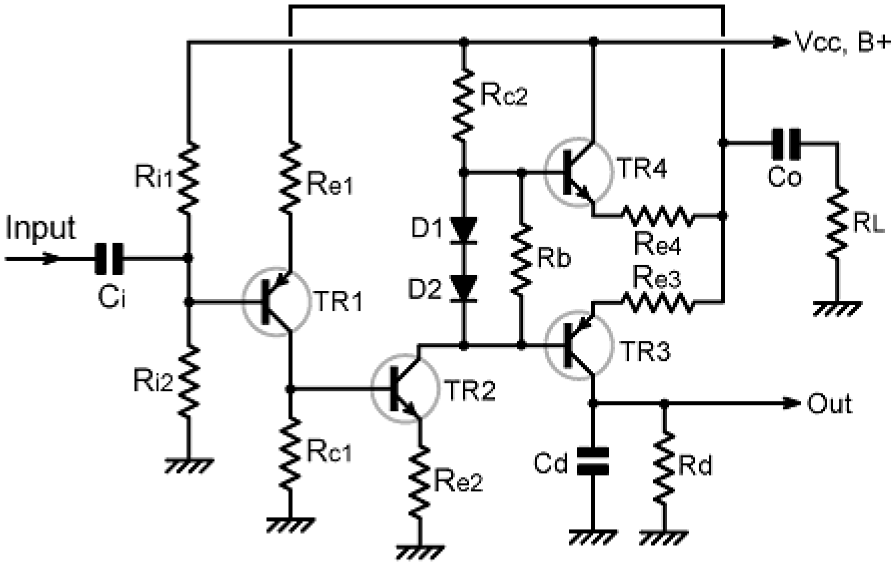
10 http://www.g3ynh.info/circuits/hi-fi_am.html

Note that, if it is desired to achieve linear detection for very small inputs, then the collector of TR3 (above) is the best point for taking the output. This is so however, with the drawback that the output is referenced to $V_{cc}$ instead of ground in the preceding circuit. Perfect supply filtration is not always achievable, and that might be a problem. A straightforward solution is to swap the PNP and NPN transistors and invert the circuit, as shown in the diagram below.




DWK