DESIGN AND PRE-PRODUCTION DEVELOPMENT OF THE DETECTAIDS 900 SERIES OF ULTRASONIC RECEIVERS

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29.8.81.

Detectaids (UK) Ltd.

Ultrasonic Test & Condition Monitoring

+ 1987 supplement

Model 950AF Ultrasonic Receiver with Audio Filter.
Models T800 mk II, T900X and T900V Ultrasonic Transmitters.
Design and Pre-production Development.

D W Knight BSc, PhD, Oct. 1987.

b

With corrections and additions.
Version 2.01, 15th March 2015

Contents

Diagrams

Note:
Some diagrams need to be rotated through 90º for viewing on a computer screen.
In Adobe Acrobat:
[Ctrl] [Shift] + rotates the display 90º clockwise
[Ctrl] [Shift] - rotates the display 90º anticlockwise

INTRODUCTION

1.1 The 900 series is a family of instruments intended for use in ultrasonic listening applications. This report considers the design of the 900 series prototype and its variants, most notably the model 950 which is intended to be intrinsically safe when used in the presence of explosive gas mixtures.

The prototype is intended to be used in conjunction with a piezoelectric ultrasonic transducer, resonant at approximately 40KHz and having an impedance of 12KΩ at this frequency. Other types of transducer may be used provided that tuning arrangements and resistive input lossing networks are altered.

A broad outline of the 900 series features is as follows:

1) A wide range of input signal amplitudes can be accomodated without overload
   a) by adjustment of the feedback ratio of the front end amplifier as a means of gain control
   b) by means of resistive lossing which can be applied at the input and is switched electronically.
2) The resonant nature of the transducer permits amplifier tuning to be used as a means of reducing the noise voltage presented to the detector. It is however possible to build a completely untuned version of the receiver if this should be desired. Production spread in the resonant frequency of transducers may necessitate some trade-off in the noise reduction attainable by means of bandwidth limitation.
3) A linear meter law is achieved by means of a negative feedback system. The meter monitors the supply current of a class 'B' amplifier which dumps energy into a load resistor. Altering the value of the load resistor gives an option for scale expansion.

4) This unit has a user A.F. gain (volume) control which enables the user to minimise the howl-'round sometimes experienced with airborne probes.

The functional elements of the receiver are shown in fig. 1.

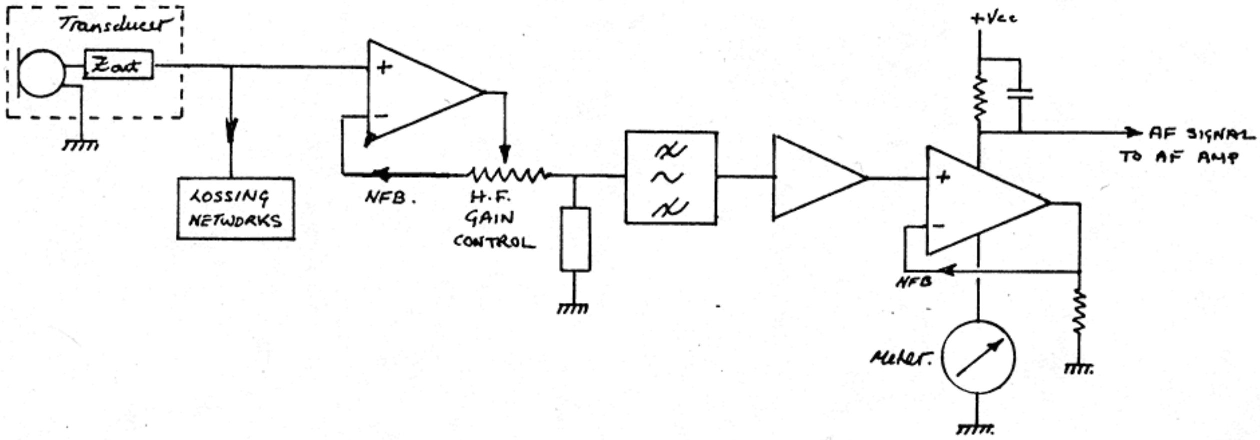


FIG 1.

The complete prototype circuit is given in fig. 2 (next page).

Note

The half-wave rectifier used in the Ni-Cd battery charging circuitry (see fig. 2) is a leftover from the earlier Detectaids 800-series. The author's attempt to change it to a bridge was rejected on grounds of cost. The use of half-wave rectification in conjunction with power transformers is not recommended.

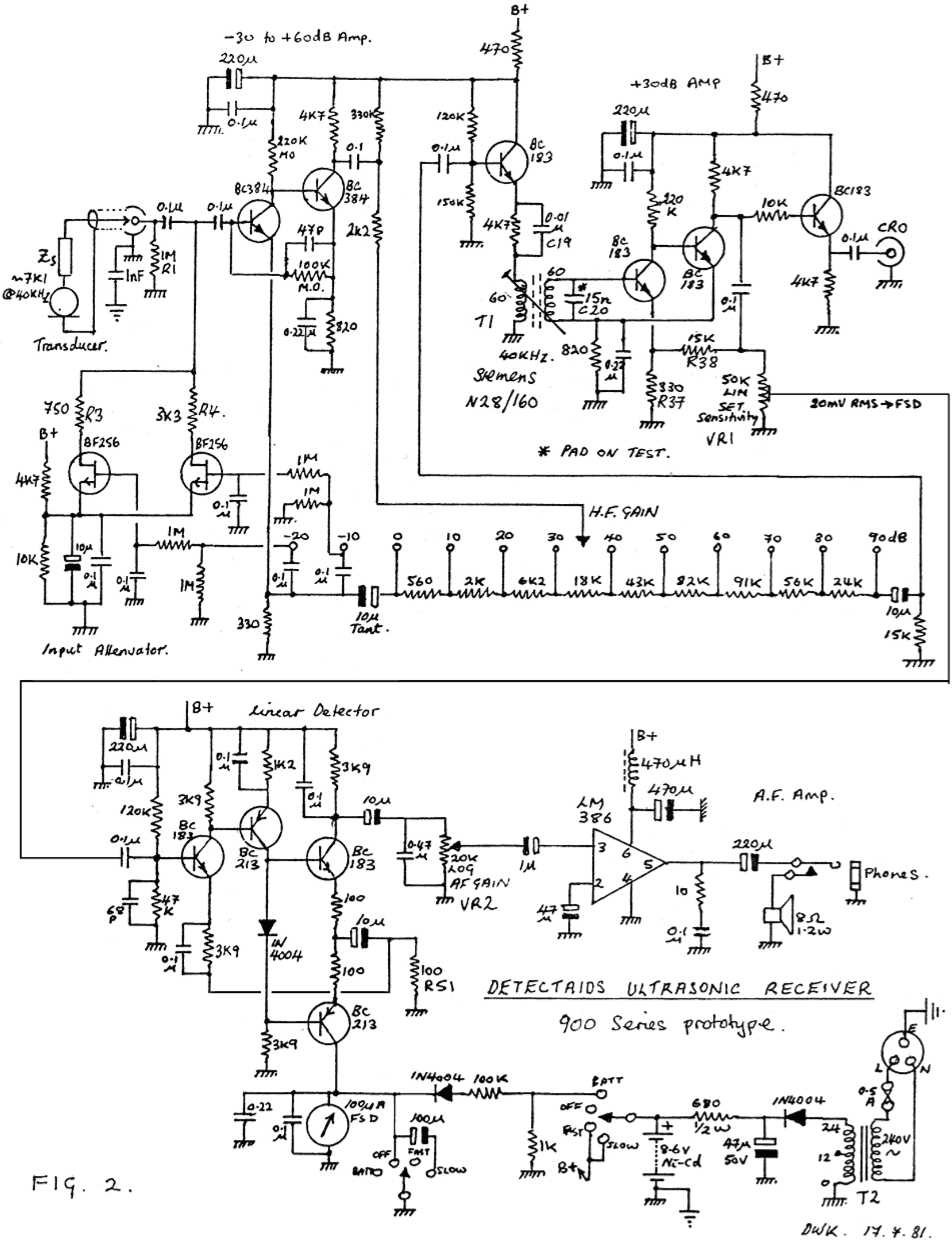


Fig. 2: 900 Series prototype circuit

CIRCUIT DESCRIPTION

2.1 Variable feedback gain control system

In order to achieve best overload protection an attenuator should be placed at the input to the system. Such an arrangement would however have the serious drawback that all of the inherent amplifier noise would be displayed by the meter. The only way to implement such a system in practice would be to limit the gain of the amplifier to the extent that the meter did not show a static noise reading. An overall gain of 60dB would be possible but this is much less than the theoretically attainable gain.

In order to utilise the maximum possible gain it is necessary to insert an attenuator into the signal chain at a point where all amplifiers after the attenuator provide insufficient noise to deflect the meter. If a simple resistive pad attenuator is used, then the amplifier before the attenuator must have a very wide dynamic range in order to operate in conditions of high ambient sound pressure.

In order to provide input protection and overcome inherent noise, most high gain amplifier designs employ attenuators both at the input and partway along the signal chain, either as ganged or as separate switches. This seems like an elegant solution to the problem except that; if the input amplifier has feedback to control its gain, which it must unless the gain is to vary with ambient temperature, then its output signal is in phase with its input signal and a consequent problem of stability arises in the associated wiring.

In this design the attenuator controls the amount of negative feedback delivered to the first amplifier simultaneously as it controls the drive level to the next. In this way an attenuator range of 90dB is achieved by means of a single pole rotary switch which varies the closed loop gain of the input amplifier between 0dB and 60dB. The input is thus only required to have a dynamic range of 30dB and the noise due to the following amplifier, which has a gain of 30dB, is not discernable on the meter.

The attenuator is considered as the following network.

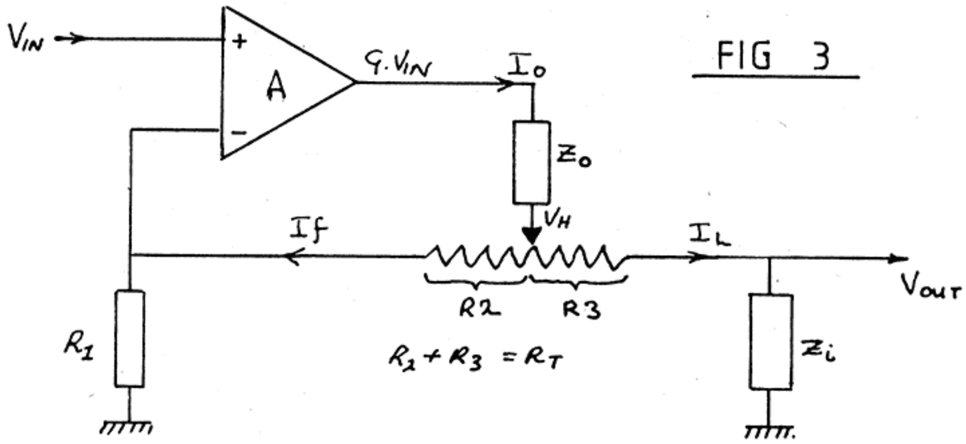


FIG 3

V_{IN} is the input voltage.

V_{OUT} is the output voltage.

A is the open loop gain of the amplifier.

G is the closed loop gain of the amplifier,

where $G = \dfrac{A}{1+A\beta} \simeq \dfrac{1}{\beta}$ if $A\beta \gg 1$

where β is the feedback ratio.

R_1 is the lumped input impedance at the NFB input.

Z_i is the lumped input impedance of the driven stage.

Z_o is the output impedance of the amplifier.

All other symbols are defined in fig. 3.

Now, $V_H = G.V_{IN} - I_o Z_o$

and $I_o = \dfrac{G.V_{IN}}{Z_o+(R_1+R_2//R_3+Z_i)}$

Therefore, $$V_H = G.V_{IN}\left[1 - \frac{Z_o}{Z_o+(R_1+R_2//R_3+Z_i)}\right] \qquad (1)$$

The feedback voltage $$V_F = V_H.\frac{R_1}{(R_1+R_2)} = G.V_{IN}\left[1 - \frac{Z_o}{Z_o+(R_1+R_2//R_3+Z_i)}\right]\frac{R_1}{(R_1+R_2)}$$

The feedback ratio $\beta \simeq \dfrac{V_F}{G.V_{IN}} = \left[1 - \dfrac{Z_o}{Z_o+(R_1+R_2//R_3+Z_i)}\right]\dfrac{R_1}{(R_1+R_2)}$

therefore

$$\beta = \left[\frac{\cancel{Z_o}+(R_1+R_2//R_3+Z_i) - \cancel{Z_o}}{Z_o+(R_1+R_2//R_3+Z_i)}\right]\frac{R_1}{(R_1+R_2)} \qquad \text{— —(2)}$$

Let the overall gain be

$$G' = \frac{V_{OUT}}{V_{IN}} = \frac{V_{OUT}}{G.V_{IN}}.G \qquad \text{— — — — — —(3)}$$

now, $\quad V_{OUT} = V_H.\dfrac{Z_i}{(R_3+Z_i)}$

therefore from (1)

$$V_{OUT} = G.V_{IN}\left[\frac{(R_1+R_2//R_3+Z_i)}{Z_o+(R_1+R_2//R_3+Z_i)}\right]\frac{Z_i}{(R_3+Z_i)}$$

therefore, $\quad \dfrac{V_{OUT}}{G.V_{IN}} = \dfrac{Z_i}{(R_3+Z_i)}\left[\dfrac{(R_1+R_2//R_3+Z_i)}{Z_o+(R_1+R_2//R_3+Z_i)}\right]$

Use with (2) and (3)

$$G' = \frac{V_{OUT}}{V_{IN}} \quad \frac{V_{OUT}}{G.V_{IN}}.\frac{1}{\beta}$$

$$G' = \frac{Z_i}{(R_3+Z_i)}\left[\cancel{\frac{(R_1+R_2//R_3+Z_i)}{Z_o+(R_1+R_2//R_3+Z_i)}}\right]\frac{(R_1+R_2)}{R_1}\left[\cancel{\frac{Z_o+(R_1+R_2//R_3+Z_i)}{(R_1+R_2//R_3+Z_i)}}\right]$$

therefore,

$$\boxed{G' = \frac{V_{OUT}}{V_{IN}} = \frac{Z_i(R_1+R_2)}{(R_3+Z_i)R_i}} \qquad \text{— — — — — — — —(4)}$$

Note that $R_2+R_3 = R_T$, a constant.

Referring to fig. 1,

as $R_3 \longrightarrow 0 \quad G' \longrightarrow G'max \qquad R_2 \longrightarrow R_T$

as $R_2 \longrightarrow 0 \quad G' \longrightarrow G'min \qquad R_3 \longrightarrow R_T$

$$G'min = \frac{Z_i}{(R_3+Z_i)} = \frac{Z_i}{(R_T+Z_i)} \qquad \text{— — — — — — — — — —(5)}$$

$$G'min = \frac{R_1+R_2}{R_1} = \frac{R_1+R_T}{R_1} \qquad \text{— — — — — — — — — — —(6)}$$

From (5), $G'min\ (R_T+Z_i) = Z_i$

From (6), $G'max.R_1 = R_1+R_T$

$$R_T = \frac{Z_i - G'min.Z_i}{Gmin}$$

$$\boxed{R_T = Z_i\left[\frac{1 - G'min}{Gmin}\right] = R_1(G'max - 1)} \qquad (7)$$

Hence by defining the required gain variation and R_1 or Z_i the network may be calculated from (4) and (7).

In practice, the chain of resistors attached to the HF gain control switch (see fig. 2) make up R_T, R_2 and R_3 being chosen by setting the switch. The 900 series gain control operates in 10dB steps from -30 to +60dB, with R_1 = 330R. Thus G'max = 60dB (1000 X) and G'min = -30dB (0.0316 X). Using equation (7), this gives Z_i = 12.33320KR. A family of solutions to a re-arrangement of equation (4), for the various choices of G' is as follows:

Gain in dB	V_{OUT}/V_{IN} =G'	R_2	R_3
60	1000	329670	0
50	316.23	305821	23849
40	100	248868	80802
30	31.62	156570	173100
20	10	71930	257740
10	3.16	26370	303300
0	1	8590	321080
-10	0.316	2540	327130
-20	0.1	580	329090
-30	0.0316	0	329670

Starting from the top of R_1, the calculated values for the resistor chain making up R_T are:

580 + 1960 + 6050 + 17,780 + 45,560 + 84,640 + 92,298
+ 56,953 + 23,849 = 329,670 R.

A practical chain using the nearest possible E2400 series resistor to the calculated value is:

560 + 2K + 6K2 + 18K + 43K + 82K + 91K + 56K + 24K = 322,760R.

Using this practical resistor chain to calculate G' for each switch setting gives:

R_2	R_3	G'	$G'(dB) = -20 \log_{10}(G')$
322760	0	979	59.82
298760	24K	302.1	49.60
242760	80K	96.08	39.65
151760	171K	30.22	29.61
69760	253K	9.618	19.66
26760	296K	3.198	10.099
8760	314K	1.0139	0.1203
2560	320K2	0.3163	-9.9967
560	322K2	0.0968	-20.279
0	322760	0.0358	-28.911

which is an adequate approximation to 10 dB step control for most purposes.

2.2 Input attenuators

The piezoelectric transducers used in conjunction with the 900 series reciever are capable of producing voltage outputs in the order of 10V P-P when subjected to large sound pressures. This large a signal cannot be linearly amplified by an amplifier having a supply voltage of approximately 8V, input attenuation is therefore needed to cope with such situations.

By placing a resistor between signal input and signal ground, an attenuator can be formed comprising the output impedance of the transducer and the shunt resistor as a divider network. In the 900 series, such shunt resistors can be placed across the input by means of F.E.T. switches (fig. 2). Each F.E.T. has its source maintained at approx. 2/3 B+. When the gate is at 2/3 B+ the attenuator is on, when the gate is at 0V, the attenuator is off.

Note that in fig. 2, the F.E.T.'s receive their gate bias via the HF gain switch. This arrangement allows the use of a single pole 12-way gain control switch and gives input attenuator facilities which, as far as the user is concerned, are simply settings of the normal gain control.

On models 900 and 910 an additional input attenuator is provided which is operated from the mode switch (S2). This is used to provide a X ½ facility. Note that for the model 910, TR1 is a logical inverter.

2.3 Tuning circuitry and intermediate amplification

In tuned versions of the 900 series receiver, (fig. 2), the output from the variable gain H.F. amplifier (section 2.1) is fed to an emitter follower stage. The emitter follower presents a high impedance load to the output of the attenuator, thus enabling the overall load impedance to be decided by suitable choice of resistors. Knowledge of this impedance is necessary in order to obtain the correct gain law for the H.F. gain control. The attenuator terminating resistor in parallel with the two base biasing resistors for the emitter follower stage thus constitute the quantity Z_i which appears in equations (4) and (7) (section 2.1).

The emitter follower drives a 40KHz tuned transformer (T1) via its emitter resistor and capacitor in parallel. The tuned transformer looks back into a low impedance port, therefore the value of the emitter coupling capacitor affects both the Q and the resonant frequency of the tuned circuit. Altering the value of this capacitor (designated C19 on all models - see Appendix) can be used as a means for choosing the selectivity of the unit. The smaller the value, the narrower the bandwidth. If the receiver is to be used with a transducer selected at random from a production batch, then its bandwidth must not be too narrow. There is, therefore, a limit to the improvement in signal to noise ratio that can be obtained by bandwidth restriction, this limit being dictated by production tolerances in transducer resonant frequency.

The output of the tuned transformer is fed to an amplifier, the gain of which is determined by negative feedback.

Thus: $$G \simeq \frac{R38}{R37}$$

ie. $$G \simeq 20\log_{10}\left(\frac{R38}{R37}\right) \text{ dB.}$$

The gain of this stage should be chosen so that the sensitivity control (VR1) has to be set in the top portion of its travel, thus avoiding unnecessary amplification of the signal.

In order to build an untuned version of the 900 series receiver, the emitter follower stage may be omitted and the intermediary stage rewired as a conventional bootstrap amplifier (fig. 4).

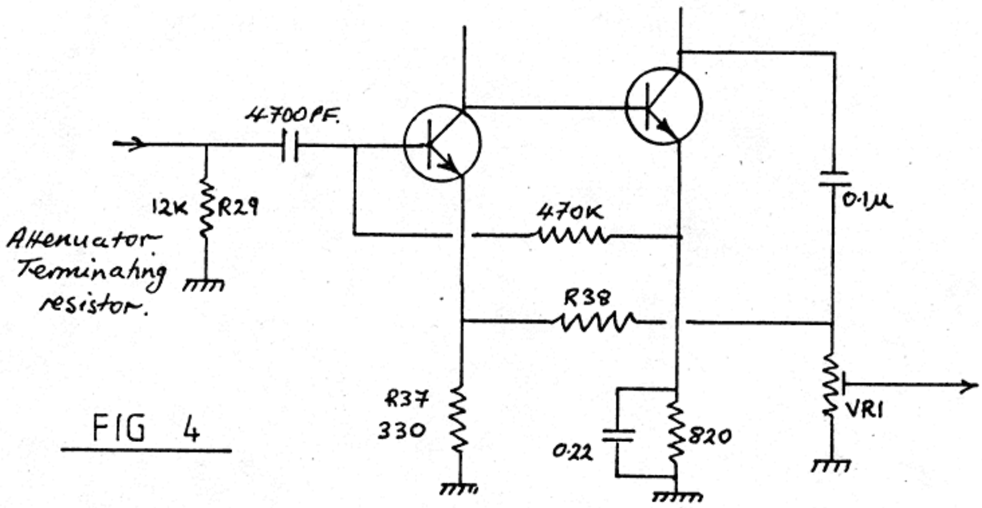


FIG 4

This option can be implemented on the existing PCB by putting the new 470KR base bias resistor into the patch intended for C20 and by connecting the 4700pF coupling capacitor from the attenuator output to the base of the first transistor using patches made available by the omission of the emitter follower and transformer.

2.4 Linear high-frequency voltmeter and detector

Conventional rectifier type voltmeter arrangements require high voltage inputs in order to overcome the effect of diode forward threshold voltages upon linearity. A bridge formed with germanium diodes will provide a reasonably linear display with 2V RMS input but it will not register at all until the input exceeds 0.4V P-P. Thus, using this arrangement, to amplify an input signal of 1uV requires a voltage gain of 126dB in order to achieve modest linearity. Amplifier stability is thus a serious problem with such an arrangement. Partial biasing of the diodes may be used to improve linearity, but such arrangements are subject to thermal drift.

In this equipment, the signal is fed to a complementary symmetrical push-pull amplifier, with a feedback factor of 1, which delivers power into a load resistor. Fig. 5.

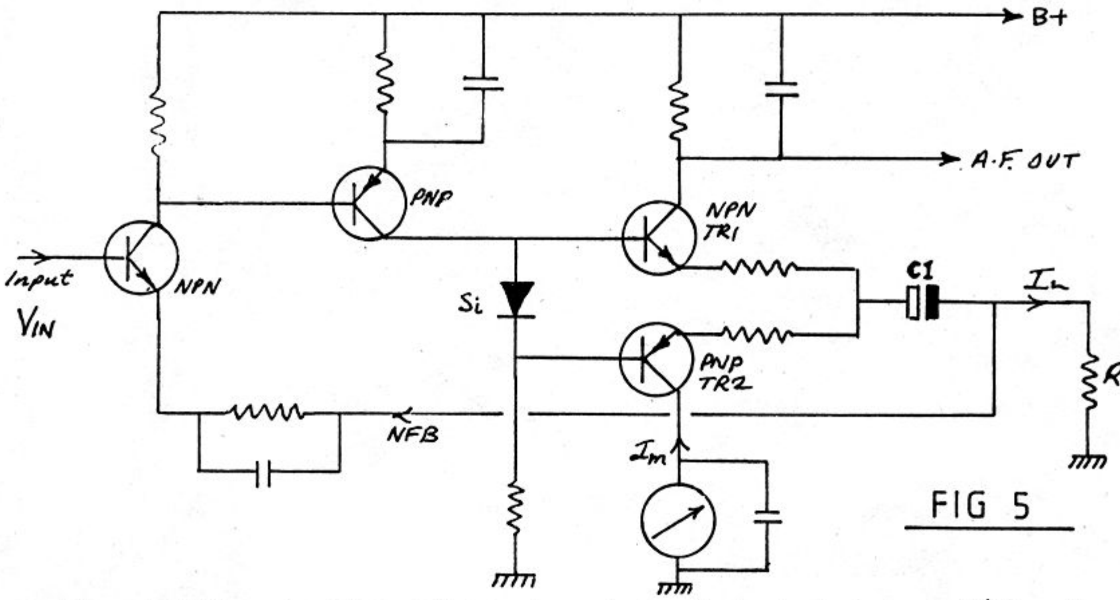


FIG 5

The amplifier is biased for zero quiescent current (class B/C). The large feedback ratio prevents crossover distortion. Thus for an input of V_{IN} volts, V_{IN} volts appears across R_L.

Hence

$I_L = V_{IN} / R_L$.

When the circuit is driven by an AC signal, TR1 drives the load current positive and TR2 drives the load current negative. Hence the average current flowing in TR2 collector is the mean half-wave rectified load current. This can be found by integration[1] and is given by:

$I_m = (\sqrt{2})\, V_{IN\,(rms)} / \pi\, R_L$

The meter movement is not at signal potential, which means that it can be mounted close to the input socket or attenuator without jeopardising circuit stability. Meter sensitivity is defined by the choice of R_L, and the scale is highly linear. Note that the meter could be placed in the collector of TR1 or TR2.

By placing a resistor in the collector of TR1 and shunting it with a suitable smoothing capacitor, the circuit also behaves as a highly linear radio detector. This output is used to provide the detected audio signal. The product of this resistance and capacitance is the detector time constant, which sets the audio corner (-3 dB) frequency according to the relationship:

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

By choosing R_L = 100 Ω and a meter movement of 100 µA FSD, the voltmeter has a sensitivity if 22.2 mV RMS FSD. A 1 µV signal thus only requires 87 dB of voltage amplification in order to give a full-scale reading.

By switching another resistor in parallel with R_L, a scale-expansion facility can be provided. This option is used in the model 950. By placing another 100 Ω resistor (R52) in parallel with R51, R_L is reduced to 50 Ω and the circuit sensitivity increases to 11.1 mV RMS FSD. The scale expansion obtained is thus ×2 (+6 dB).

The 100 µA FSD meter can, if so desired, be calibrated in dB. The following table gives dB points on a 0 - 100 scale assuming + 3dB = 100%. This can be used for constructing the meter-scale artwork.

dB	-20	-15	-10	-9	-8	-7	-6	-5	-4	-3	-2	-1	0	+1	+2	+3
% FS	7.08	12.59	22.39	25.12	28.18	31.62	35.48	39.81	44.67	50.12	56.23	63.10	70.79	79.43	89.12	100

1 The derivation is given in the article: **A Linear high-frequency voltmeter and AM detector**. D W Knight. http://www.g3ynh.info/circuits/AM_det.html

2.5 A.F. amplifier stage

The audio bandwidth of the receiver may be tailored by choice of capacitors C45 and C46 (model 900), C36 and C45 (model 910) and C34 and C37 (model 950). (See Appendix).

The audio output circuit utilises an LM386 integrated circuit, which has a voltage gain of 20 (26dB). The small inductor (L1) prevents distortion products of the A.F. amplification process from appearing on the supply rail. In the absence of L1, distortion products in the 40KHz region cause instability at high volume settings.

2.6 Model 950 electronic supply switching

In the presence of an explosive gas atmosphere, a piece of battery operated electronic apparatus might be expected to cause an explosion at switch-on because of the spark created at the instant the battery is connected to the circuit decoupling capacitors. The model 950 is designed to have a spark-free switch-on by virtue of electronic supply switching (TR15). When switched on, the small signal stages draw their supply current from TR15 base, base current causes TR15 to turn on, thus connecting the battery to the meter and A.F. output stages. Note that TR15 is also turned on in the battery check position, so that the meter and A.F. stages receive their power supply but the small signal stages do not.

ADJUSTMENT AND IMPLEMENTATION OF OPTIONS

3.1 Functions of the unit

The model 900 has PCB mounted switches and is therefore difficult to modify. If input attenuators are not required, the HF gain control end stop may be moved and TR1, 2, 3 and their associated circuitry omitted. In the event of an unusual customer requirement, there is room in the cabinet to mount hard wired switches as used in models 910 and 950. This arrangement has not been tried to date and care should be taken to avoid instability. Note that a patch is provided in the model 900 PCB (DT006) which can be used to place a resistor across R51 for scale expansion (see section 2.4). This method of scale expansion is used as standard in the model 950.

If so desired, the meter connections can be brought out to a panel socket. This option will provide a chart recorder signal of approximately 0.4V for FSD of the meter (assuming that the meter has a resistance of 4KR).

3.2 Input attenuators

Any amount of input attenuation can be chosen by altering the values of R2, R3 and R4. These resistors can be considered to form a divider network in conjunction with the output impedance of the transducer, since the input impedance of the amplifier is high enough to be neglected in most instances.

Taking R2 as example: if the transducer has Z_{OUT} = 12K , then if R2 = 12K the attenuation achieved is X $\frac{1}{2}$ (or -6dB).

In general, $\quad \text{attenuation} = \dfrac{R2}{R2+Z_{OUT}}$

or $\quad \text{attenuation} = 20 \log_{10}\left(\dfrac{R2}{R2+Z_{OUT}}\right)$ dB.

3.3 H.F. gain control law

Alternative switch increments may be calculated by using equations (4) and (7) in section 2.1.

3.4 Choice of H.F. bandwidth

Reducing the value of C19 reduces the bandwidth of the receiver. Any change in C19 necessitates retuning of C20/T1. Bandwidth reduction improves signal to noise ratio but cannot be carried out indefinitely due to production spread in transducer resonant frequency.

3.5 Adjustment of tuning

This is required on production test. Suggested procedure:

1) Choose C20 = 22 nF.
2) Connect a signal generator to the input of the receiver and sweep through the region 20 - 60 KHz.
3) Alter or pad C20 appropriately.
4) When a resonance close to the desired frequency is achieved, final tuning may be accomplished by adjustment of T1 core.

3.6 Intermediate gain and sensitivity adjustment

Rotate VR1 until the desired sensitivity is achieved. 1 µV FSD on the highest sensitivity setting is appropriate. If the desired sensitivity is not attainable and there are no electrical faults present (eg. input attenuators operating when they should not be or incorrectly tuned H.F.) then increase the value of R38 (see section 2.3). R38 should in any case be chosen so that VR1 is normally towards the top end of its travel.

3.7 Signal to noise ratio

The unit should have a signal to noise ratio of approximately 8dB for a 1μV input, with the input terminated in 12KR. If the unit is adjusted for 1μV FSD for maximum gain, place a 12KR resistor across the input terminals and see that the noise is less than 0.4μV. If this is not the case, then investigate the input amplifier and its associated circuitry or look for circuit instability.

The capacitor C10 (models 900, 910) or C7 (model 950) must be a low leakage tantalum type. If this is not the case, low frequency noise will be heard in the AF output at all settings of the H.F. gain control. Any leakage current is injected into the inverting input of the first amplifier and is consequently not subject to the limitations of loop gain.

3.8 Scale expansion

Choice of the resistor switched in parallel with R51 determines the scale expansion ratio.

Using model 950 nomenclature,

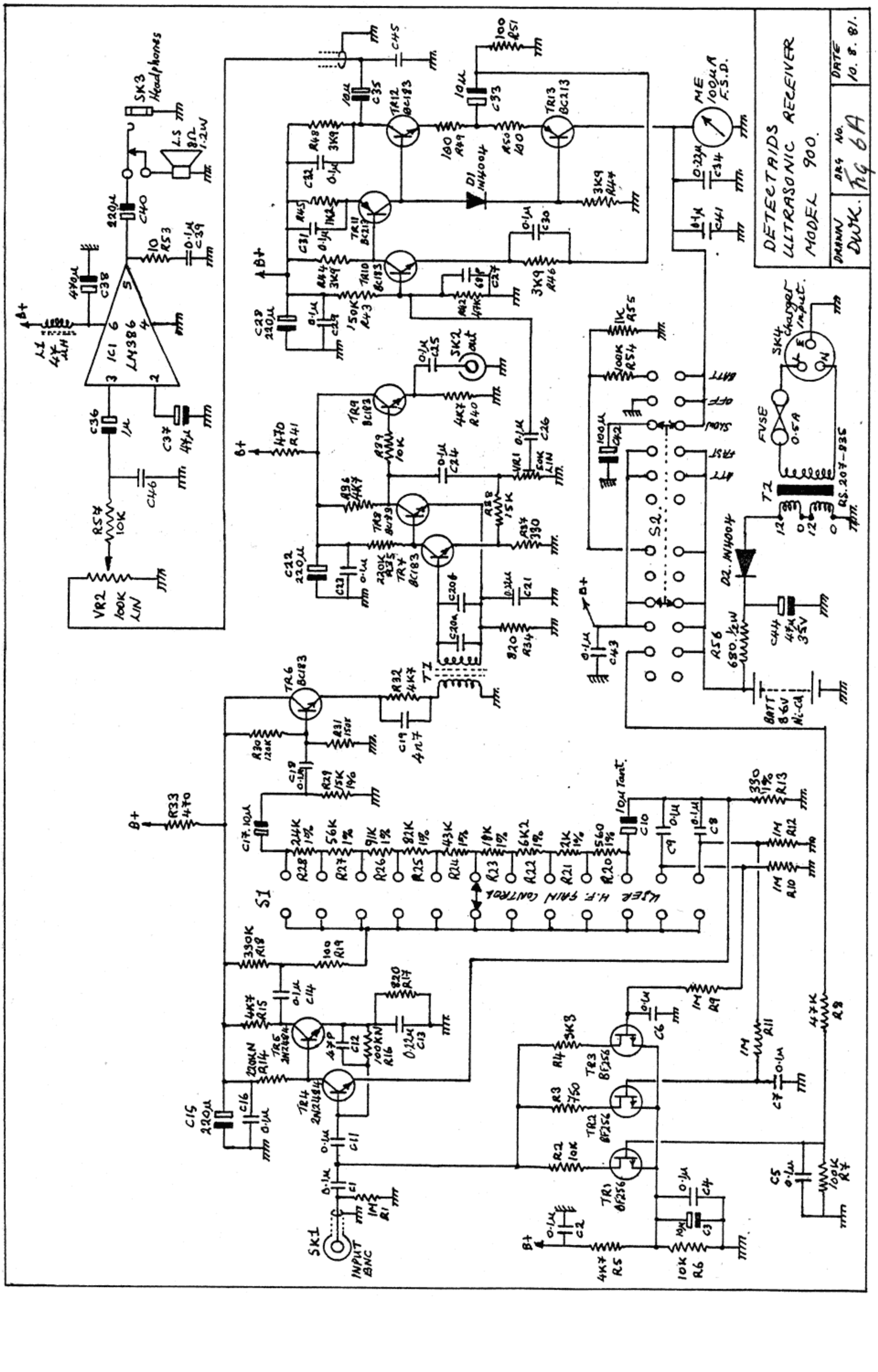
$$\text{expansion ratio} = \frac{R51 + R52}{R52}$$

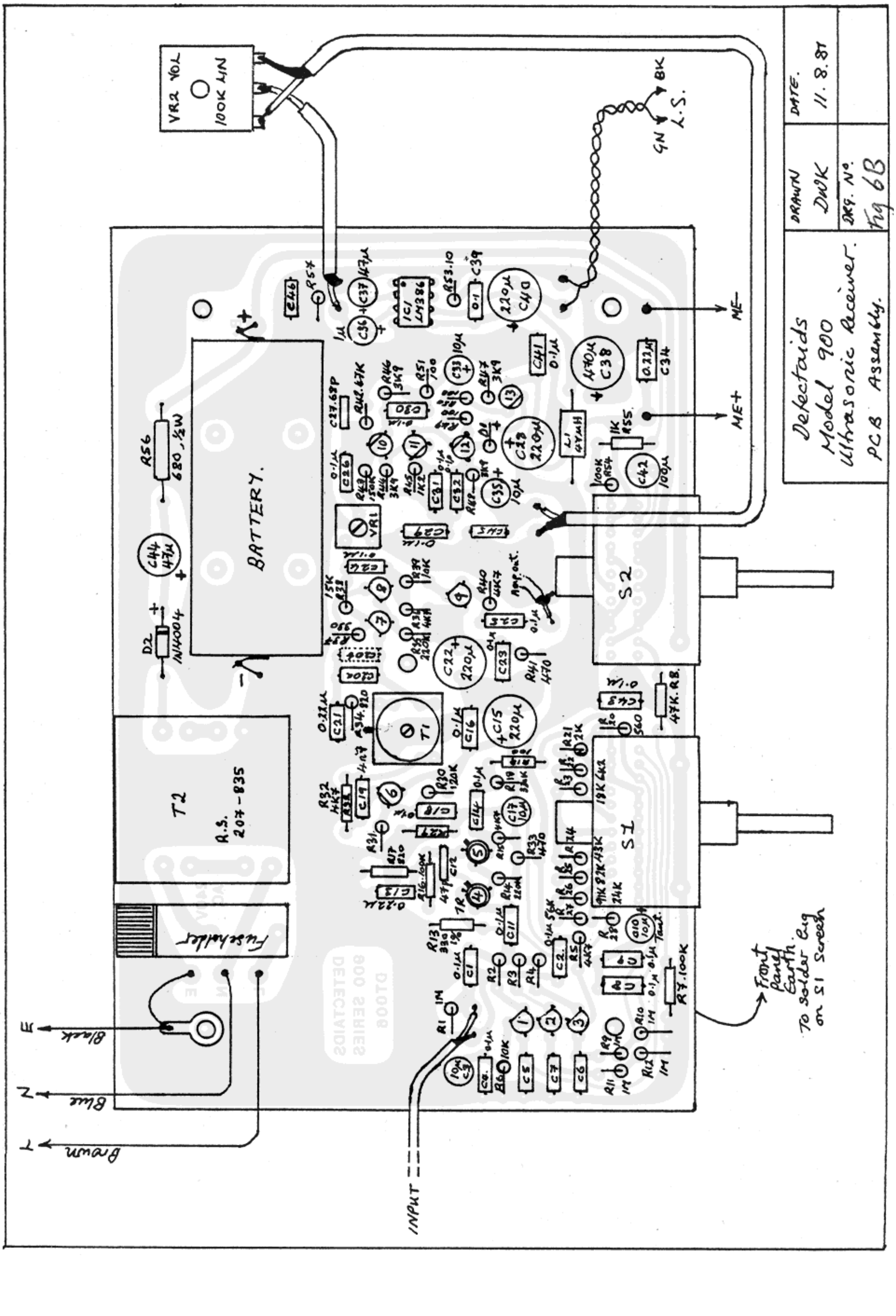
$$= 20 \log_{10}\left(\frac{R51 + R52}{R52}\right) \text{ dB}$$

3.9 A.F. bandwidth

A.F. bandwidth may be tailored by choice of C45, C46 (model 900), C36, C45 (model 910) and C34, C37 (model 950).

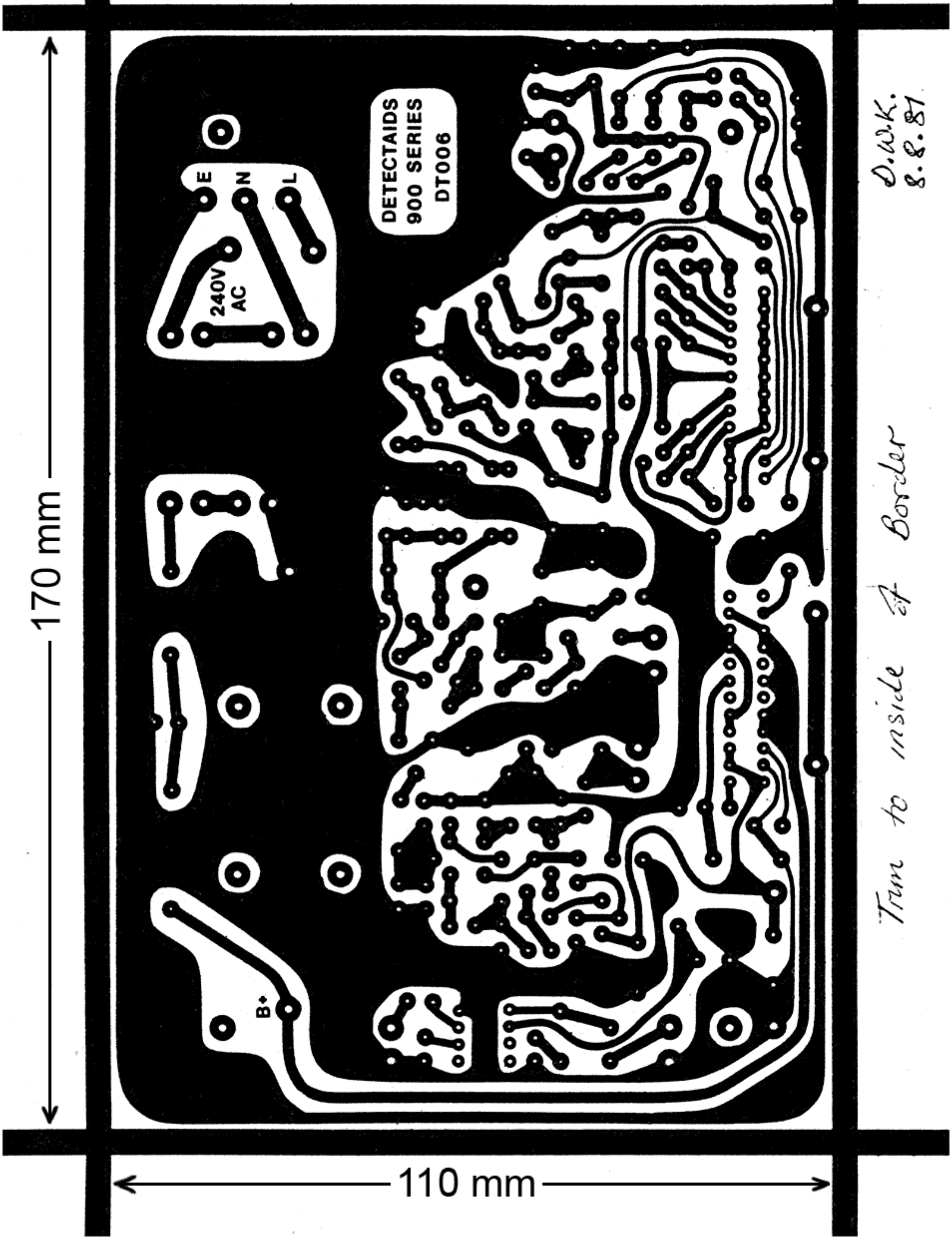
Note that the only reason for including these capacitors is to make the 900 series receivers sound like the old 800 series receiver, which had very poor audio quality.

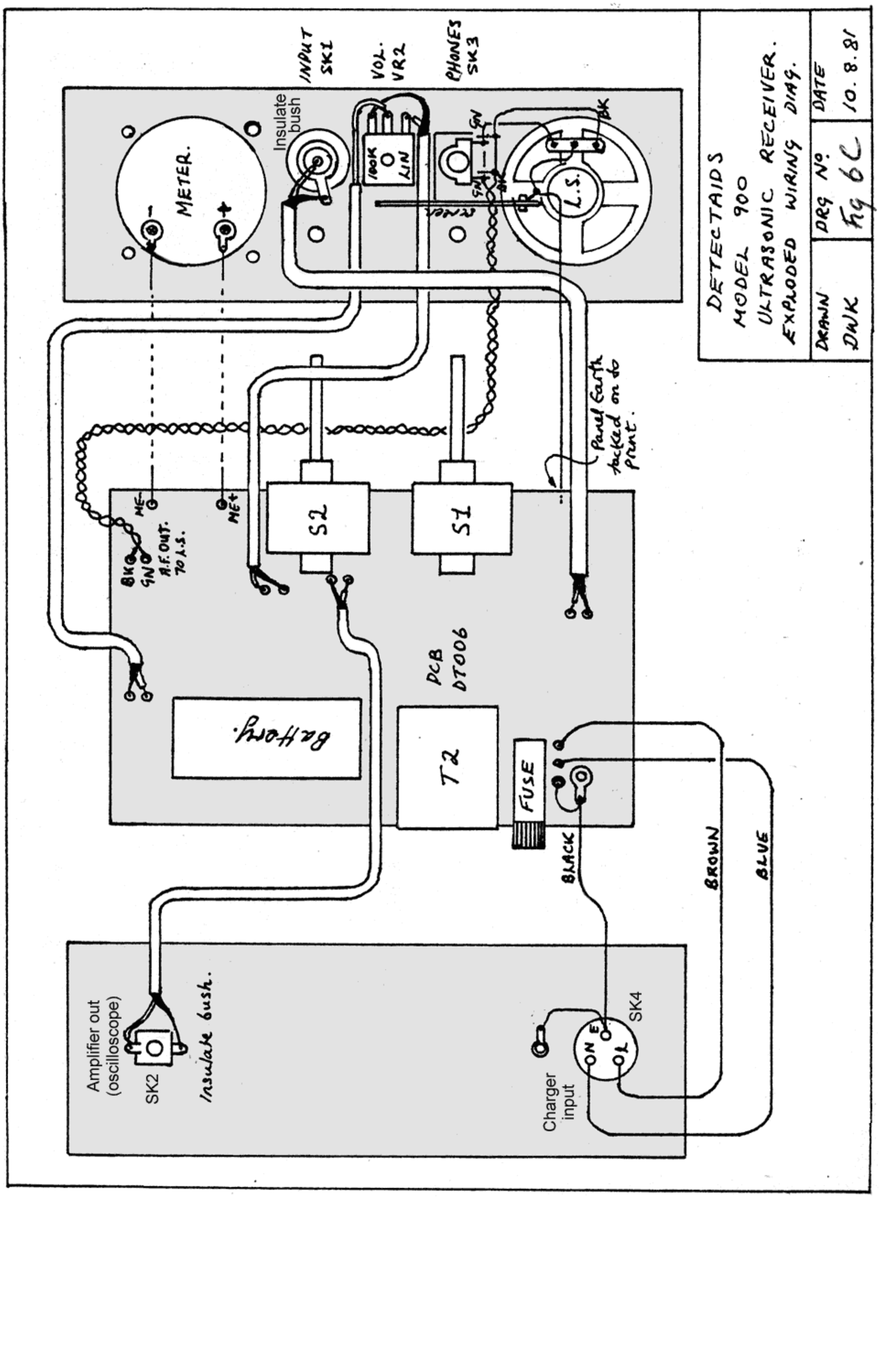

DETECTAIDS
ULTRASONIC RECEIVER
MODEL 900.
DRAWN
DWK.
DRG No.
Fig 6A
DATE
10. 8. 81.
SK1
INPUT BNC
SK2
out
SK3
Headphones
SK4
Charger input.
IC1
LM386
VR2
100K
LIN
ME
100μA
F.S.D.
FUSE
0·5A
RS.207-835
BATT
8·6V
Ni-Cd
S1
USER H.F. GAIN CONTROL
S2
BATT
OFF
SLOW
FAST
ATT


Detectoids
Model 900
Ultrasonic Receiver.
PCB Assembly.
DRAWN
DWK
DATE.
11.8.81
DRG. No.
Fig 6B
BATTERY.
INPUT
S1
S2
T2
R.S. 207-835
Fuseholder
VR2 VOL
100K LIN
L.S.
ME+
ME-
Front panel Earth lug
To solder lug on S1 Screen

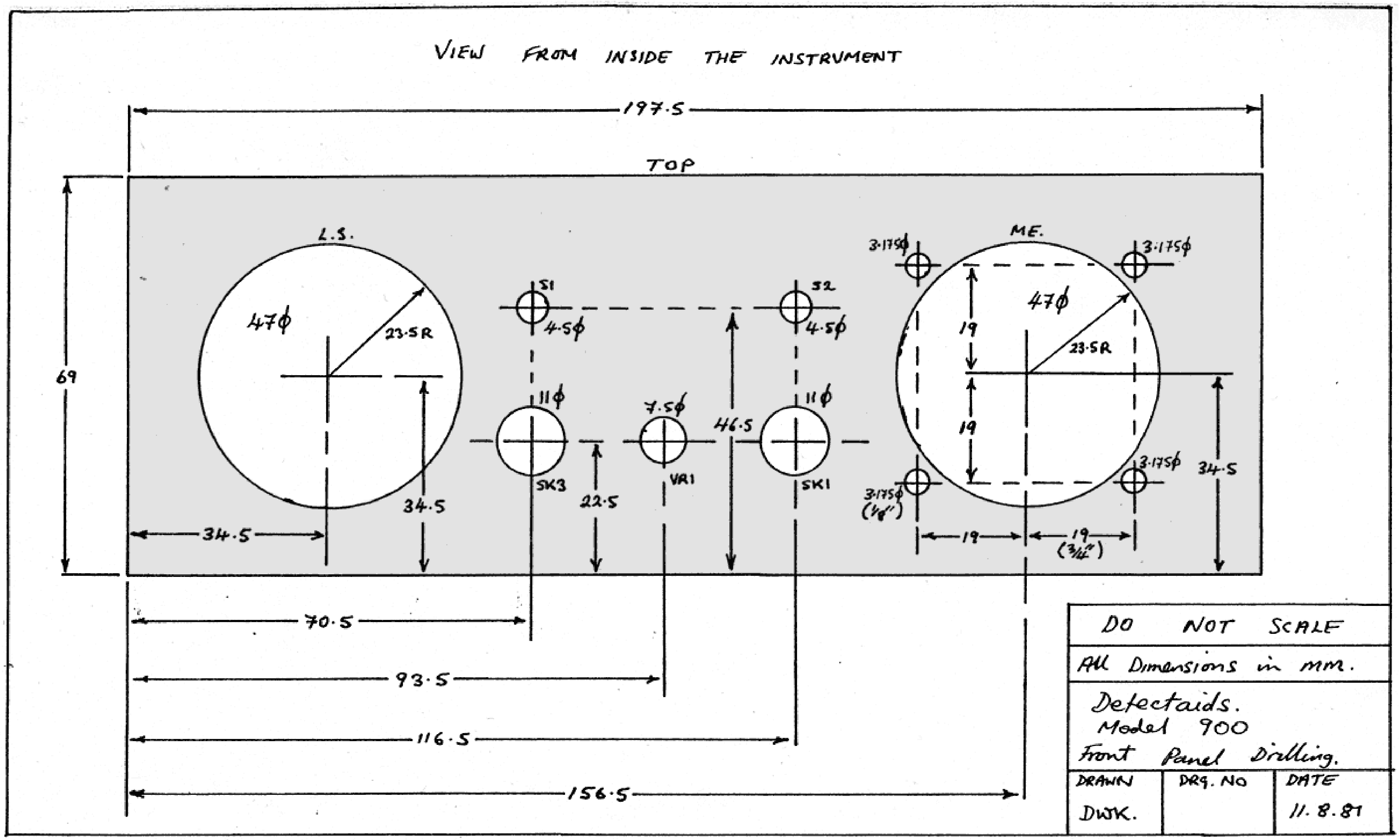
PCB Artwork. Detectaids Model 900

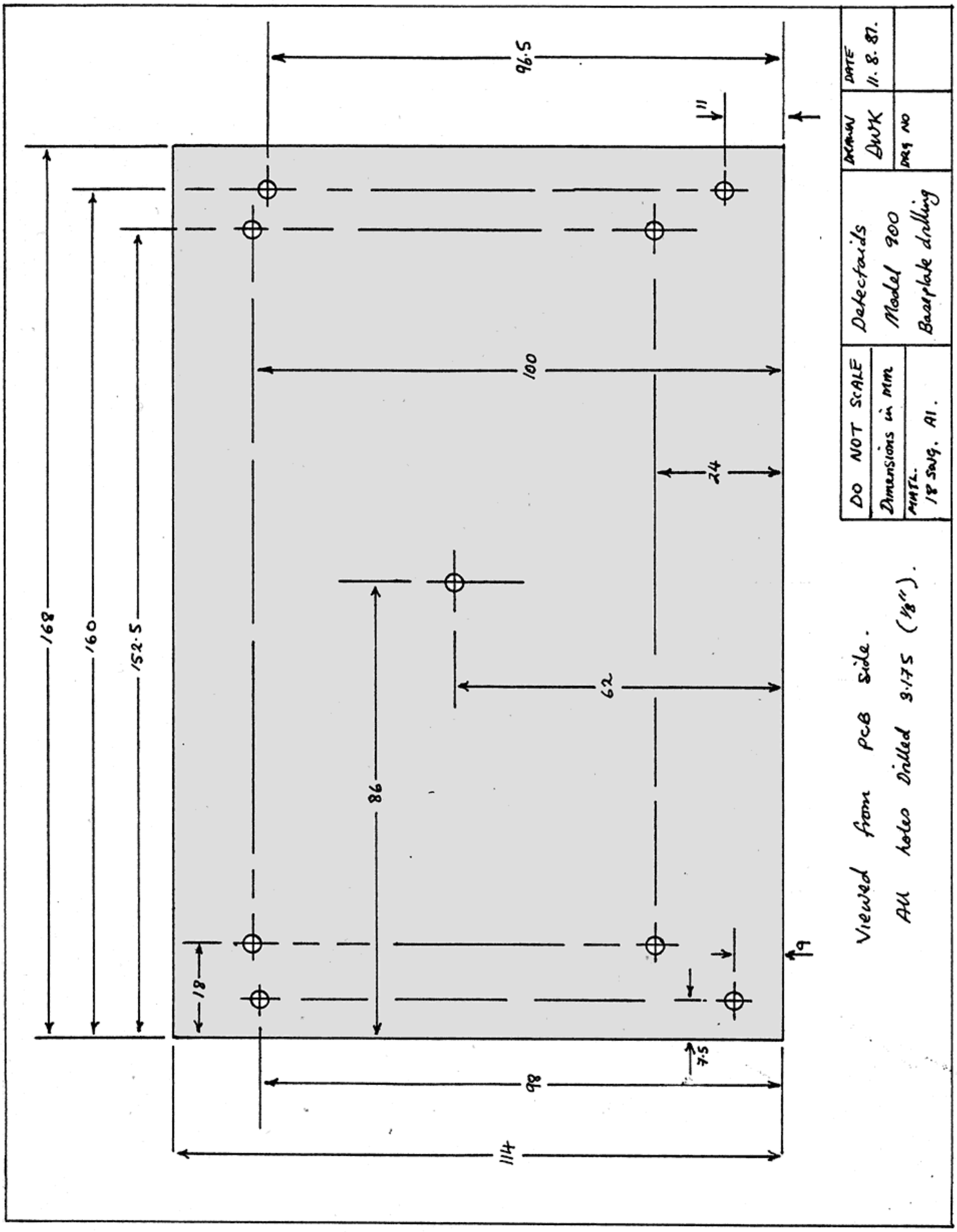
Expose from this side.

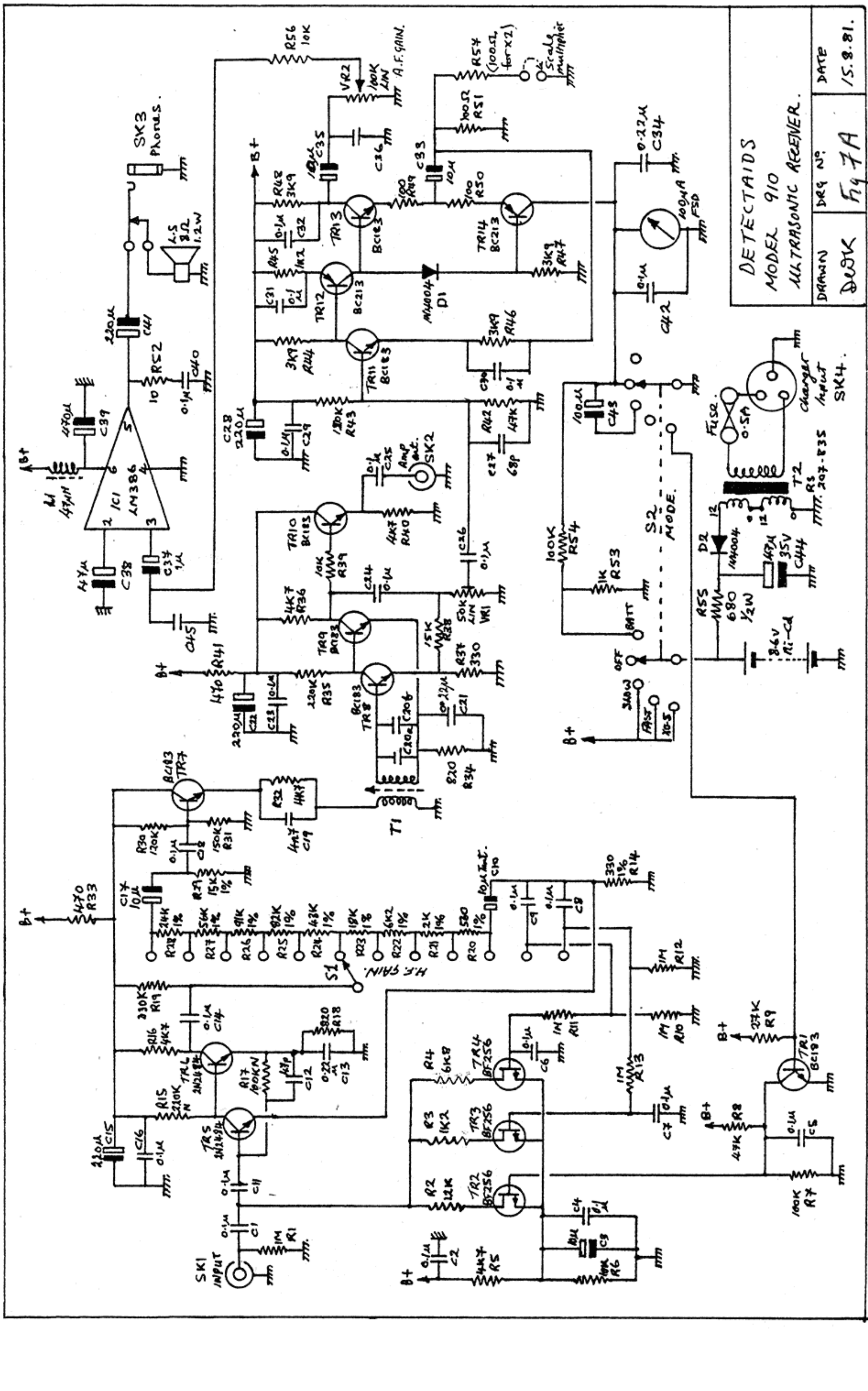


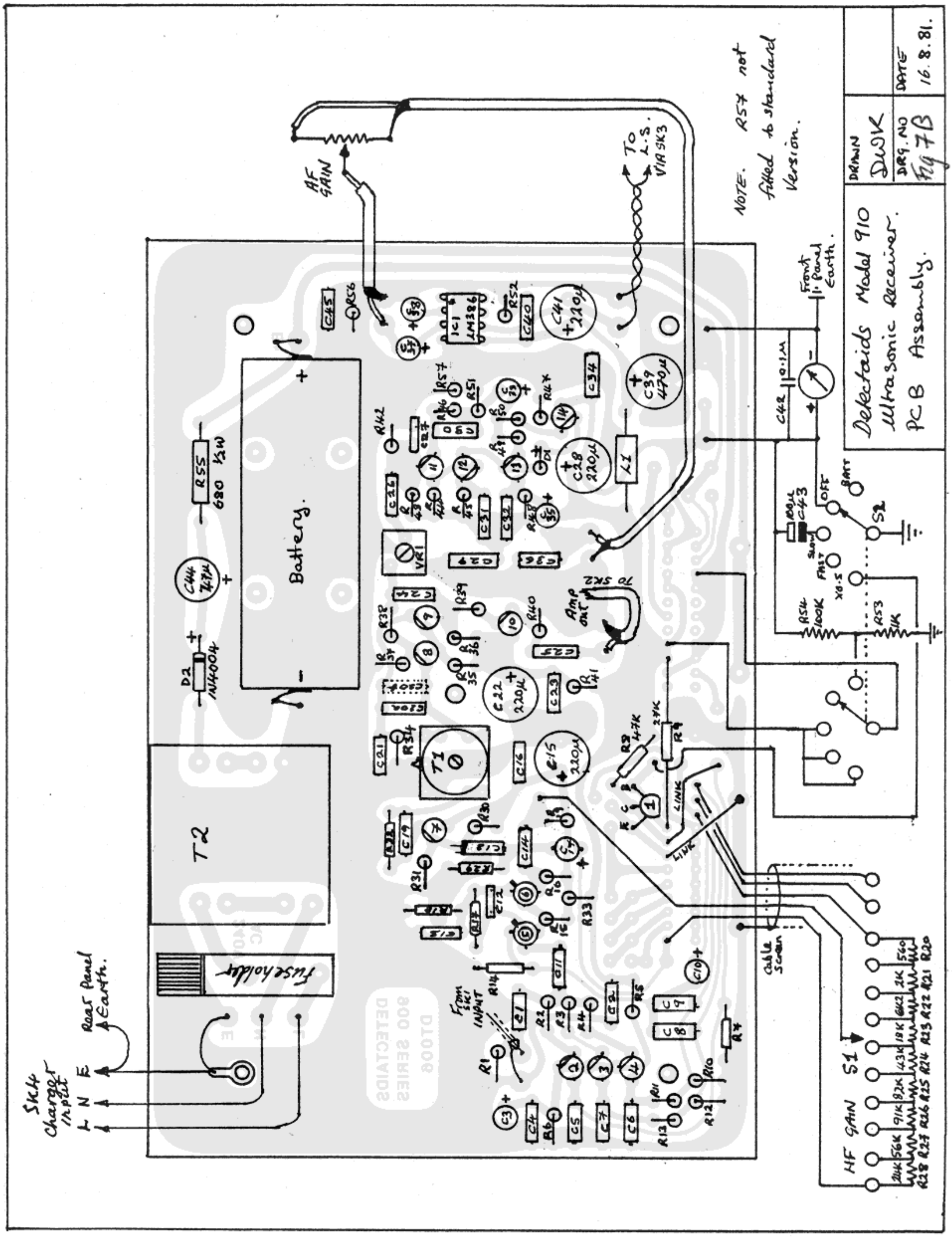


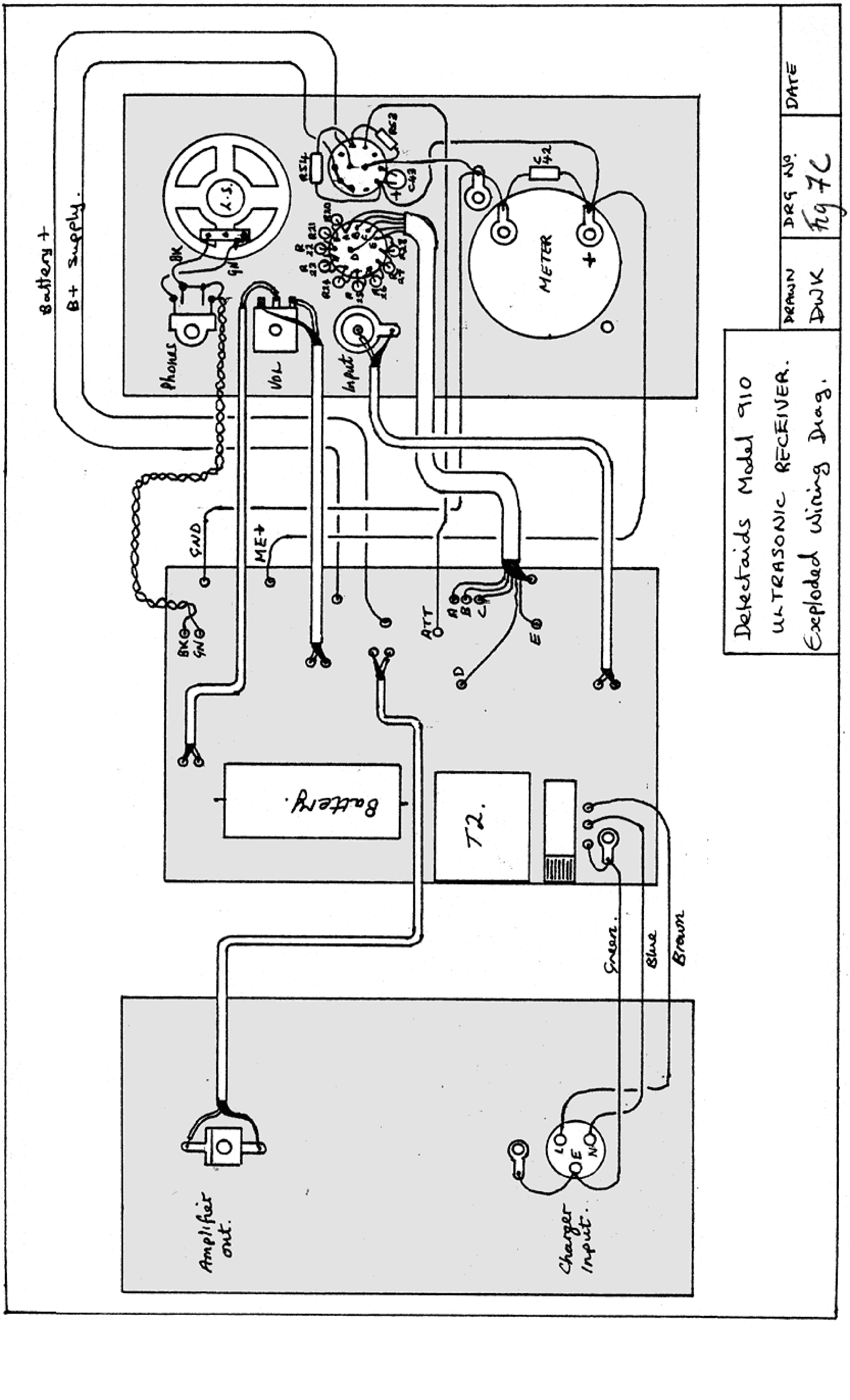
METER.
INPUT SK1
VOL. VR2
PHONES SK3
Insulate bush
100K LIN
L.S.
BK
GN
SW1
DETECTAIDS
MODEL 900
ULTRASONIC RECEIVER.
EXPLODED WIRING DIAG.
DRAWN
DWK
DRG No
Fig 6C
DATE
10.8.81
Panel Earth tacked on to Print.
S2
S1
ME-
ME+
BK
GN
A.F. OUT. TO L.S.
PCB DT006
Battery.
T2
FUSE
BLACK
BROWN
BLUE
Amplifier out (oscilloscope)
SK2
Insulate bush.
Charger input
SK4
N
E
L

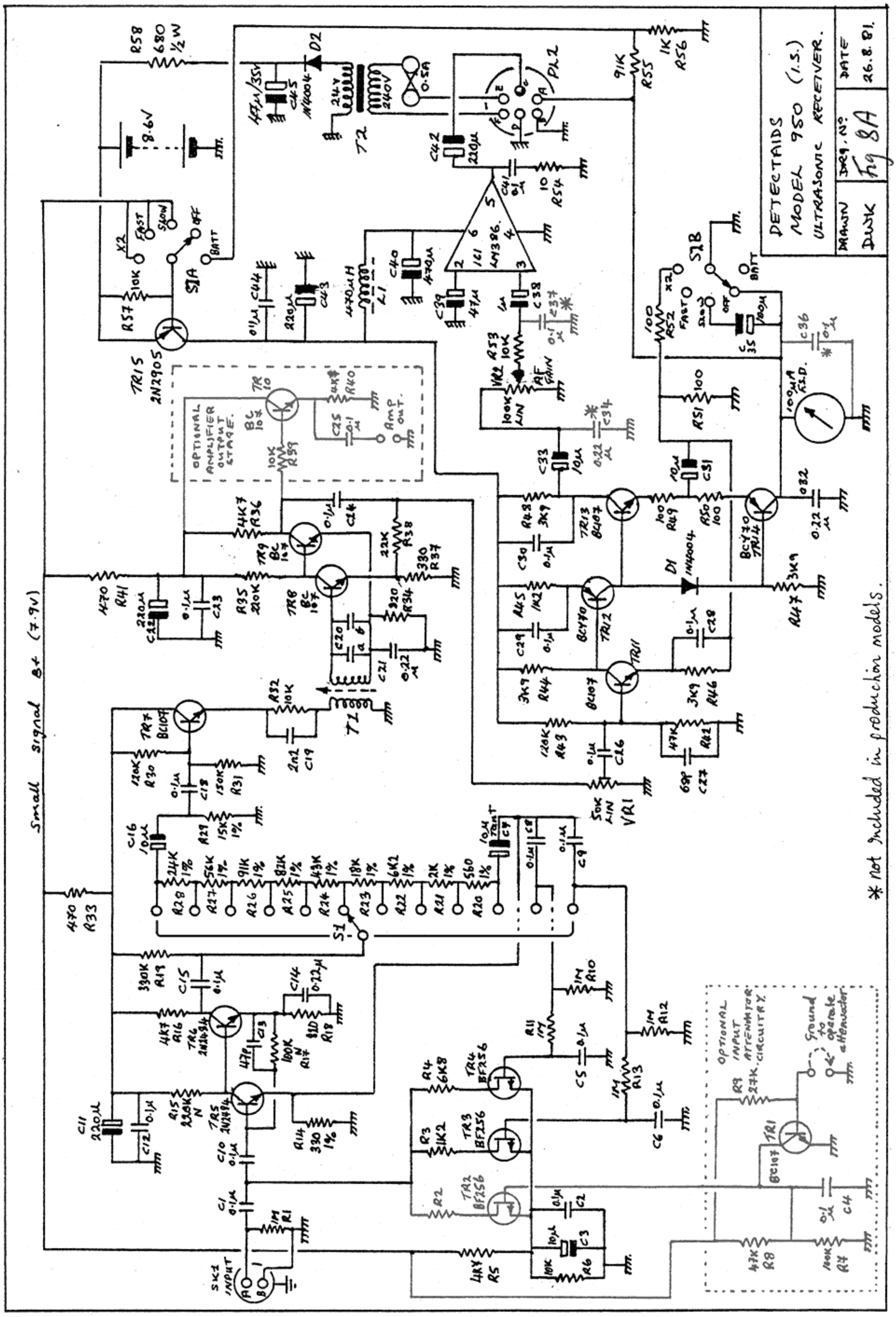

VIEW FROM INSIDE THE INSTRUMENT
197·5
TOP
69
L.S.
47φ
23·5R
34·5
34·5
S1
4·5φ
11φ
SK3
22·5
7·5φ
VR1
46·5
S2
4·5φ
11φ
SK1
ME.
3·175φ
3·175φ
3·175φ
3·175φ (1/8")
47φ
23·5R
19
19
19
19 (3/4")
34·5
70·5
93·5
116·5
156·5
DO NOT SCALE
All Dimensions in mm.
Detectaids.
Model 900
Front Panel Drilling.
DRAWN
DWK.
DRG. NO
DATE
11. 8. 81

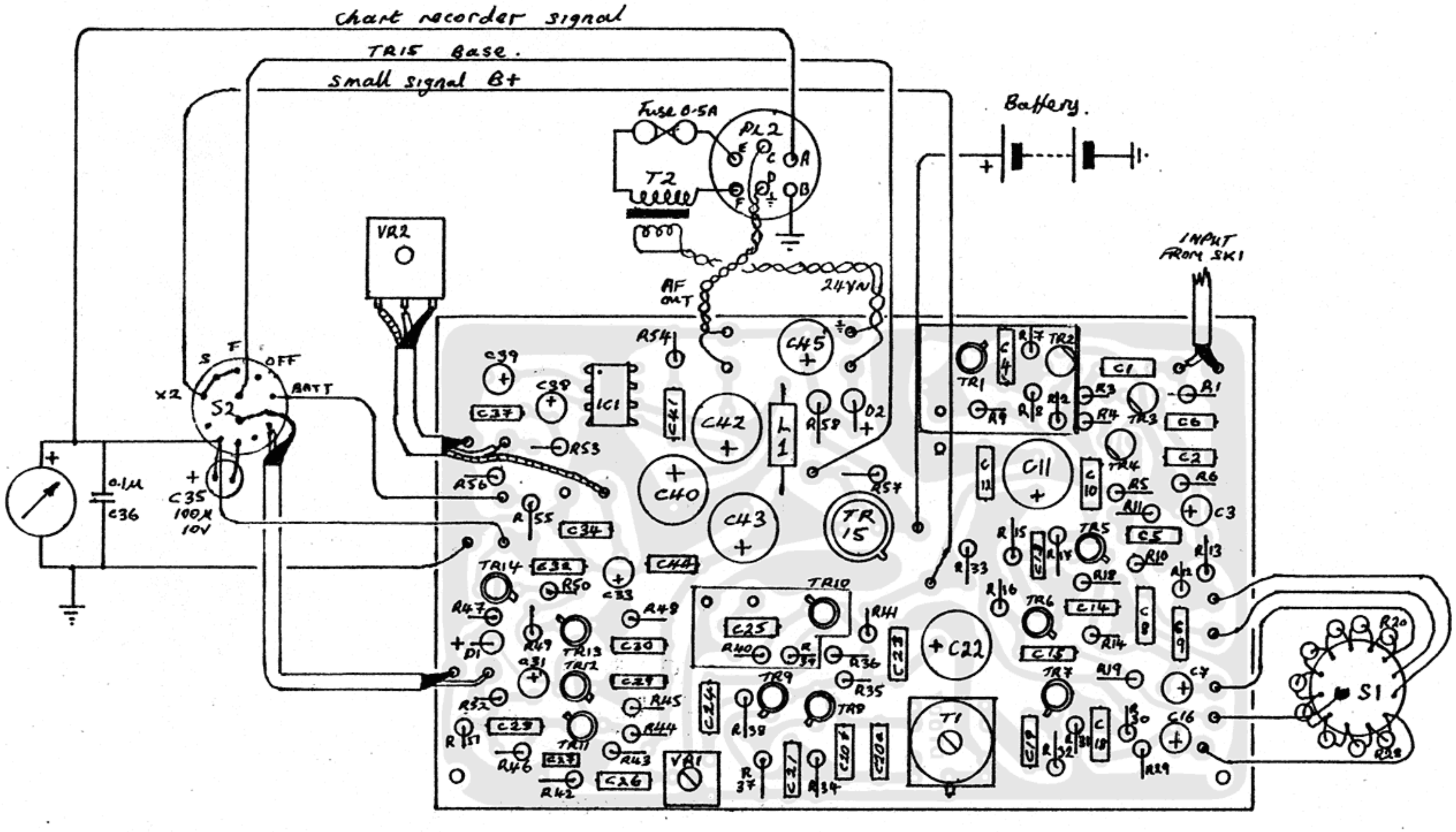

96.5
11
6
100
24
62
86
168
160
152.5
18
7.5
98
114
DO NOT SCALE
Dimensions in mm
MATL. 18 SWG. Al.
Model 900
Baseplate drilling
DRAWN DWK
DATE 11.8.87.
DRG NO
Viewed from PCB side.
All holes Drilled 3.175 (1/8").


DETECTAIDS
MODEL 910
ULTRASONIC RECEIVER
DRAWN
DRG No.
Fig 7A
DATE
15.8.81.


NOTE. R57 not fitted to standard Version.
Detectaids Model 910 Ultrasonic Receiver. PCB Assembly.
DRAWN DWK
DRG. NO Fig 7B
DATE 16.8.81.
AF GAIN
TO L.S. VIA SK3
Battery.
T2
Fuseholder
SK4 Charger Input
Rear Panel Earth.
Front Panel Earth.
HF GAIN S1
S2
Amp out
TO SK2
cable screen
LINK


Battery +
B+ Supply.
L.S.
BK
GN
Phones
VOL
Input
R54
R53
C43
C42
METER
+
GND
ME+
ATT
A
B
C
D
E
BK
GN
Battery.
T2.
Green.
Blue
Brown
Amplifier out.
Charger Input.
LO
E
N
Detectaids Model 910
ULTRASONIC RECEIVER.
Exploded Wiring Diag.
DRAWN
DWK
DRG No
Fig 7C
DATE

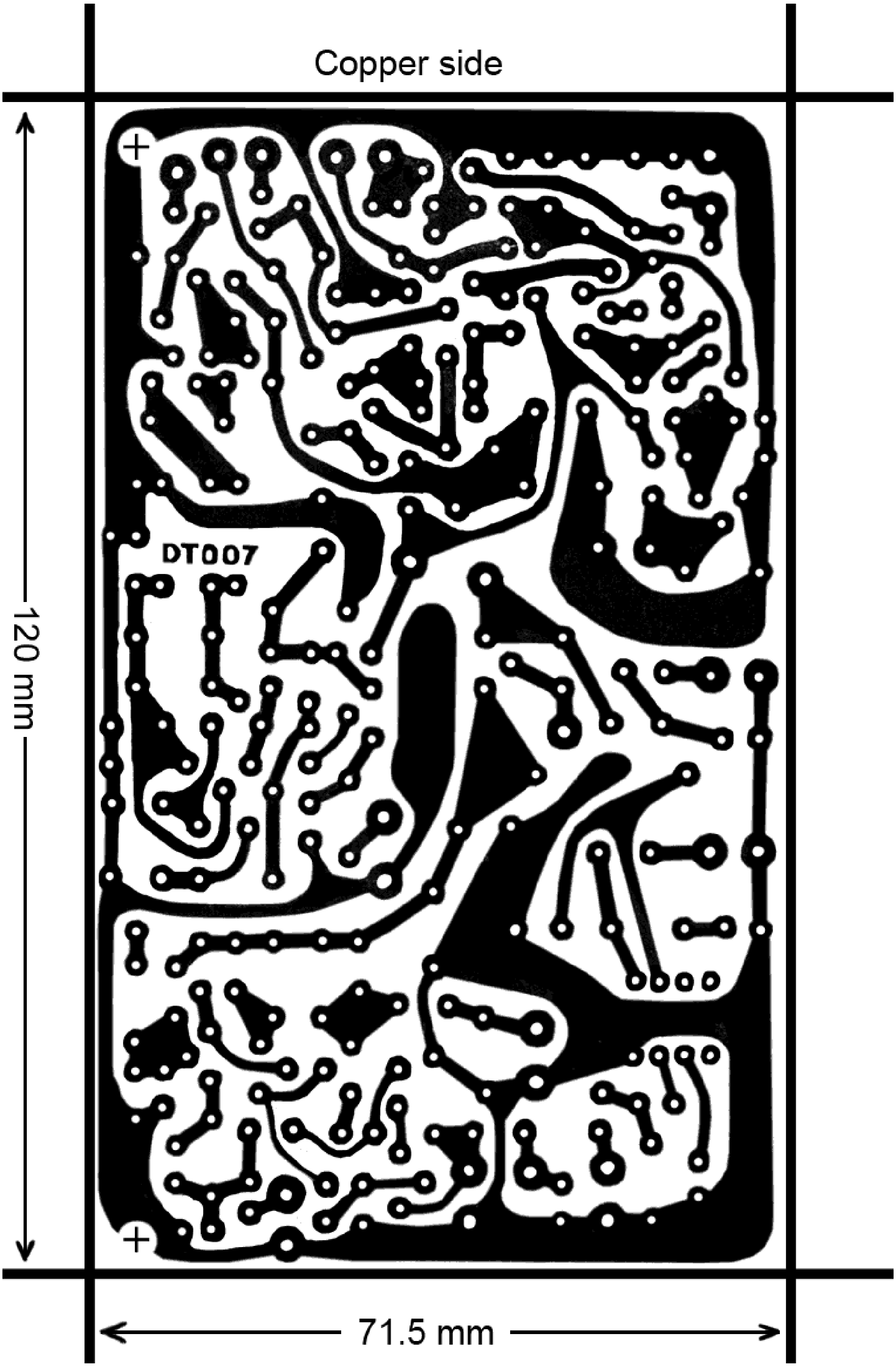

DETECTAIDS
MODEL 950 (I.S.)
ULTRASONIC RECEIVER.
DRAWN DWK
DRG. No. Fig 8A
DATE 26.8.81.
Small signal B+ (7.9V)
OPTIONAL AMPLIFIER OUTPUT STAGE.
OPTIONAL INPUT ATTENUATOR CIRCUITRY.
* not included in production models.



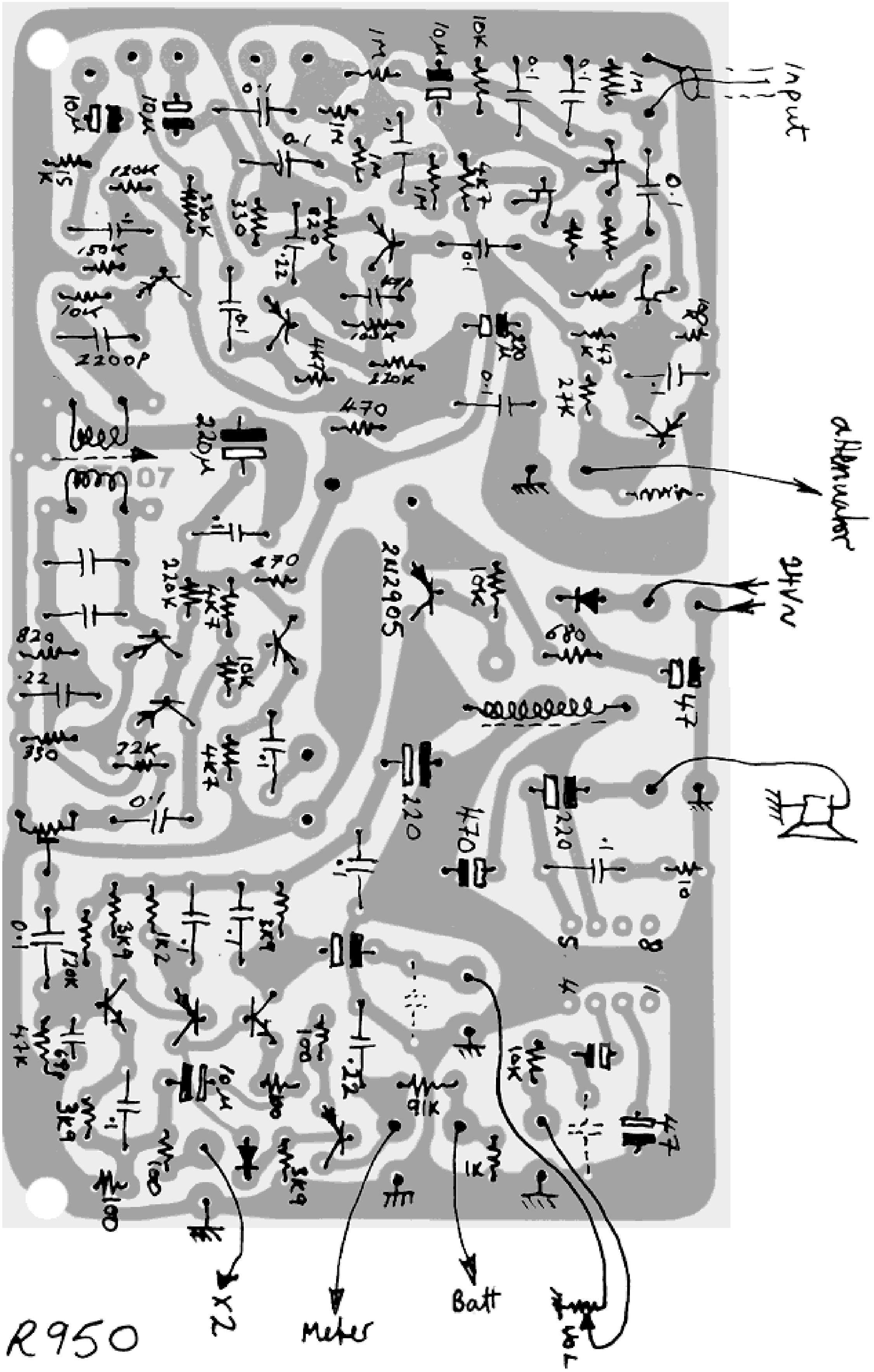
Notes. 1) TR1, 2, 10, R2, 7, 8, 9, 39, 40 and C4, 25
Belong to optional circuitry and need not be included on standard versions.

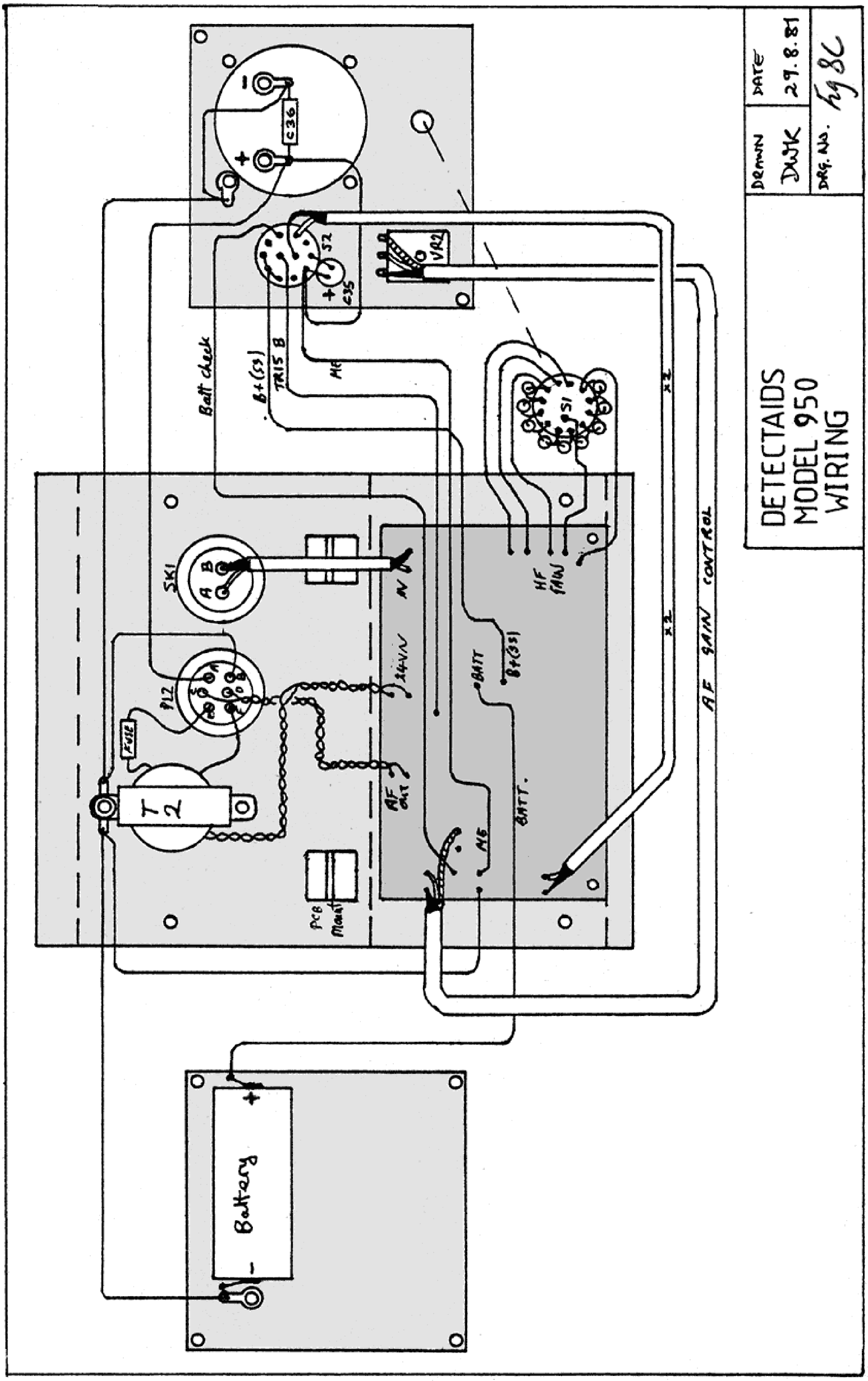
2) R55 and R56 may be mounted on S2 if desired, thus eliminating one wire.

DETECTAIDS MODEL 950 PCB LAYOUT		
DRAWN	DRG No	DATE.
DWK	Fig 8B	27.8.81



Detectaids model 950 IS. Printed circuit board


Input
attenuator
24V
2N2905
Batt
Meter
X2
Vol
R950


DETECTAIDS
MODEL 950
WIRING
DRAWN
DWK
DATE
29.8.81
DRG. No.
fig 8C
C36
C35
S2
VR2
S1
Batt check
B+(S3)
TRIS B
SK1
PL2
FUSE
T 2
AF OUT
24V/N
BATT
B+(S3)
HF FAW
AF GAIN CONTROL
x2
BATT.
PCB Mount
Battery

Detectaids 900 Series Supplement.

Model R950AF Ultrasonic Receiver with Audio Filter.

Models T800 mkII, T900X and T900V Ultrasonic Transmitters.

Design and Pre-production Development.

Introduction

This document describes additions to the 900 series range of instruments, designed to improve performance in situations where receiver - transmitter pairs are operated in conditions of high ambient noise. This improvement is obtained by equipping the receiver with an audio filter, designed to select the modulation component of the transmitted signal. Modifications to the transmitter are also required because the existing 800 series transmitter does not have an accurately defined or stable modulation frequency.

Note on Nomenclature.

This document introduces transmitters into the 900 series range. Consequently, the letters R and T have been added to the model designations. The basic receiver thus becomes the R900 and the basic transmitter the T900. The new receiver is identical to the earlier R950 except for the inclusion of an audio filter module. It is therefore designated R950AF. The new transmitters are the T800 mkII the T900X and the T900V. The T800 mkII is electrically identical to the earlier model 800 but incorporates production changes. For the T900, X and V refer to versions having either crystal controlled or variable modulation frequency.

Model R950AF

This model retains all of the earlier capabilities of the R950, but provides the user with an additional, switch selectable, audio filter and VU (volume unit) meter. When the filter option is selected, audio from the detector is re-routed through a bandpass filter before being presented to the volume control and audio output stage. The filter is tuned to the transmitter modulating frequency (977Hz in the prototype) and therefore helps to isolate the transmitter signal from background noise. The panel meter is also switched to monitor audio level at the output of the filter, the signal being taken from a point which is independent of the volume control setting. The meter sensitivity is thus controlled, from the user's viewpoint, by the setting of the HF gain control as before.

Bandpass Filter

The filter used is a second-order multiple feedback path filter (MFPF) based around an operational amplifier. This type of filter uses no inductors and is therefore inexpensive to manufacture. Appropriate design equations are given in figure A1. For these equations to hold true, the op-amp must have a high open loop gain and a very high input impedance. Most bipolar op-amps (eg. 741) are therefore not suitable. The circuit performance will also be degraded by poor component tolerances and it is recommended that 1% metal film resistors and 2.5% polystyrene film capacitors are used.

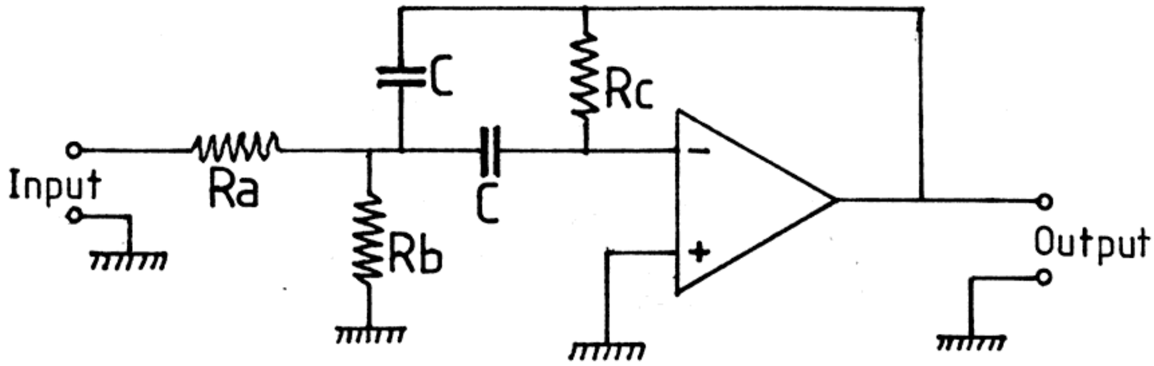
Note that in the process of filtration, the signal is first attenuated by the resistor network comprising Ra and Rb and then re-amplified by the op-amp. Inspection of the design equations reveals that this attenuation-reamplification process becomes more extreme as Q is increased. This can easily lead to a situation in which the op-amp is operated with very high gain, giving rise to attendant noise and instability problems, gain-bandwidth product restrictions, and the additional difficulty that the design equations break down when the amplification required becomes comparable to the open loop gain of the amplifier. For this reason, very high values of Q are not practicable. The open-loop gain of the op-amp, at the desired frequency, should always be significantly greater than $20Q^2$.

For the prototype, a centre frequency (fo) of 977 Hz was chosen. This corresponds to 1MHz/1024 and is therefore a convenient transmitter modulating frequency. A gain of unity (A=1) was selected since no additional audio amplification is required. A number of calculations were then performed as follows:

(all filters have fo=977Hz, A=1)

	Q	C	Ra	Rb	Rc
a)	20	10 nF	325949	408	651899
b)	20	4.7 nF	693509	868	1387018
c)	12	10 nF	195570	681	391139
d)	12	4.7 nF	416606	1450	832211

Single stage Multiple Feedback Bandpass Filter.



Design Procedure:

1) Choose centre frequency f_o and overall gain A.
2) Select Q from $Q = f_o / w$, where w is the bandwidth at the half power (-3dB voltage) points.
3) Choose C on the basis of availability of high-tolerance capacitors (eg., 2·5% polystyrene).
4) Calculate resistor values as follows:

$$Ra = \frac{Q}{2\pi f_o C A}$$

$$Rb = \frac{Q}{2\pi f_o C (2Q^2 - A)} = \frac{A\,Ra}{2Q^2 - A}$$

$$Rc = \frac{Q}{\pi f_o C} = 2A\,Ra$$

To discover the parameters of a given filter, use the following equations:

$$f_o = \frac{1}{2\pi C}\sqrt{\frac{Ra + Rb}{Ra\,Rb\,Rc}} \qquad A = \frac{Rc}{2\,Ra}$$

$$Q = \frac{1}{2}\sqrt{\frac{(Ra + Rb)Rc}{Ra\,Rb}}$$

Fig A1

The Q=20 filters have an input attenuation factor of 800 (58dB). They will be noisy and prone to instability in practice. The Q=12 filters have an input attenuation factor of 288 (49dB), which is still large but tolerable. Filter (c) was therefore selected for development since its Q is adequate for the present purpose and it has a smaller feedback resistor than filter (d).

Audio Filter and VU Meter Module.

Figure A2 shows the circuit of the new module incorporated into the R950AF. It consists of a practical implementation of the bandpass filter described in the previous section, feeding into an audio level meter circuit. Note, in the practical filter, that the virtual earth at the non-inverting input of the op-amp is taken to a point at half the supply voltage.

The op-amp used must not only fulfil the filter requirements of high gain and high input impedance, it must also be capable of operating from the available supply rail of (nominally) 8V. The choice of op-amp is therefore critical. The original circuit was built using a CA3130E CMOS op-amp, but was not unconditionally stable. Four integrated circuits, from two production batches, were tried and all exhibited momentary instability at switch off. This instability was apparently layout independent and could not be cured without degrading filter performance. The problem was completely cured by using a CA3140E op-amp. Three ICs from two batches were tried and all performed correctly. Other candidate op-amps were the LF355 (2 ICs, 2 batches), ADOP07 (2 ICs, 2 batches) and the TL091CP (3 ICs, 2 batches). The LF355 and the ADOP07 showed a tendency towards momentary instability at switch on. The TL091CP performed correctly in all cases. Of the above devices, the CA3140E is the least expensive.

The VU meter circuit consists of a one-transistor amplifier (TR16) feeding into a voltage doubler or diode pump (D4,D5). Meter sensitivity adjustment is provided by VR3. TR17 and its associated base network is provided in order to prevent C54 charging current, at switch on, from flowing in the meter. At switch on, C55 charging current turns TR17 on momentarily, thus bypassing C54 charging current to ground.

R950AF Production Details.

The complete circuit of the R950AF is given in figure A3. Physical wiring and PCB layout is given in figure A4. Note that the existing R950 component designations are unchanged. New components begin at R59, C46, D3 and TR16. C36, which was originally wired directly accross the meter terminals, is now moved onto the filter board and increased to 1uF. This capacitor was deleted from the production R950, but is now essential to the proper operation of the VU meter circuit.

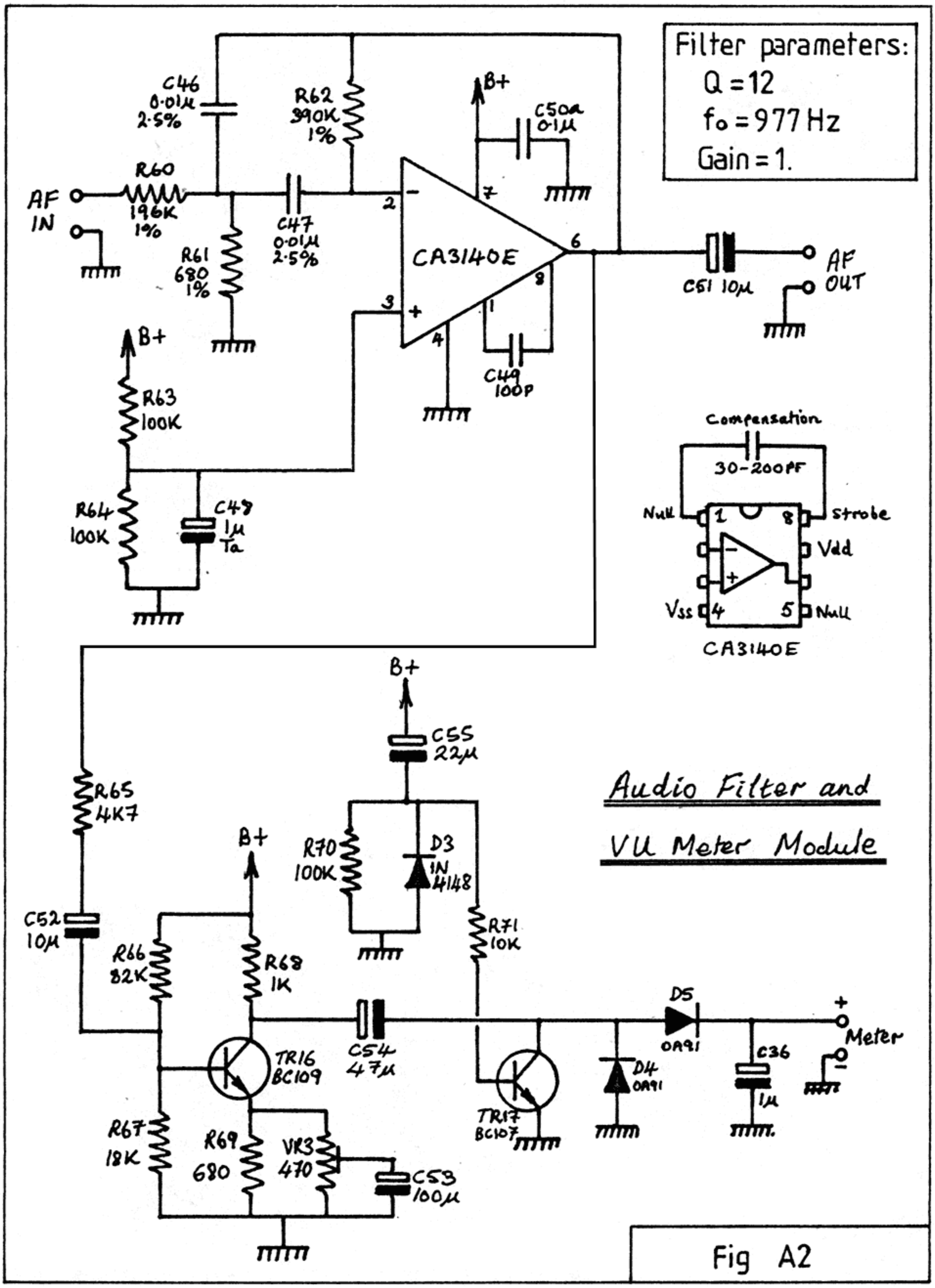

Filter parameters:
Q = 12
fo = 977 Hz
Gain = 1.
AF IN
R60 196K 1%
C46 0·01μ 2·5%
R62 390K 1%
C47 0·01μ 2·5%
R61 680 1%
B+
C50a 0·1μ
CA3140E
C49 100P
C51 10μ
AF OUT
R63 100K
R64 100K
C48 1μ Ta
Compensation
30-200PF
Null
Strobe
Vdd
Vss
Null
CA3140E
R65 4K7
C52 10μ
C55 22μ
R70 100K
D3 1N4148
R71 10K
Audio Filter and
VU Meter Module
R66 82K
R68 1K
C54 47μ
TR16 BC109
TR17 BC107
D5 OA91
D4 OA91
C36 1μ
Meter
R67 18K
R69 680
VR3 470
C53 100μ

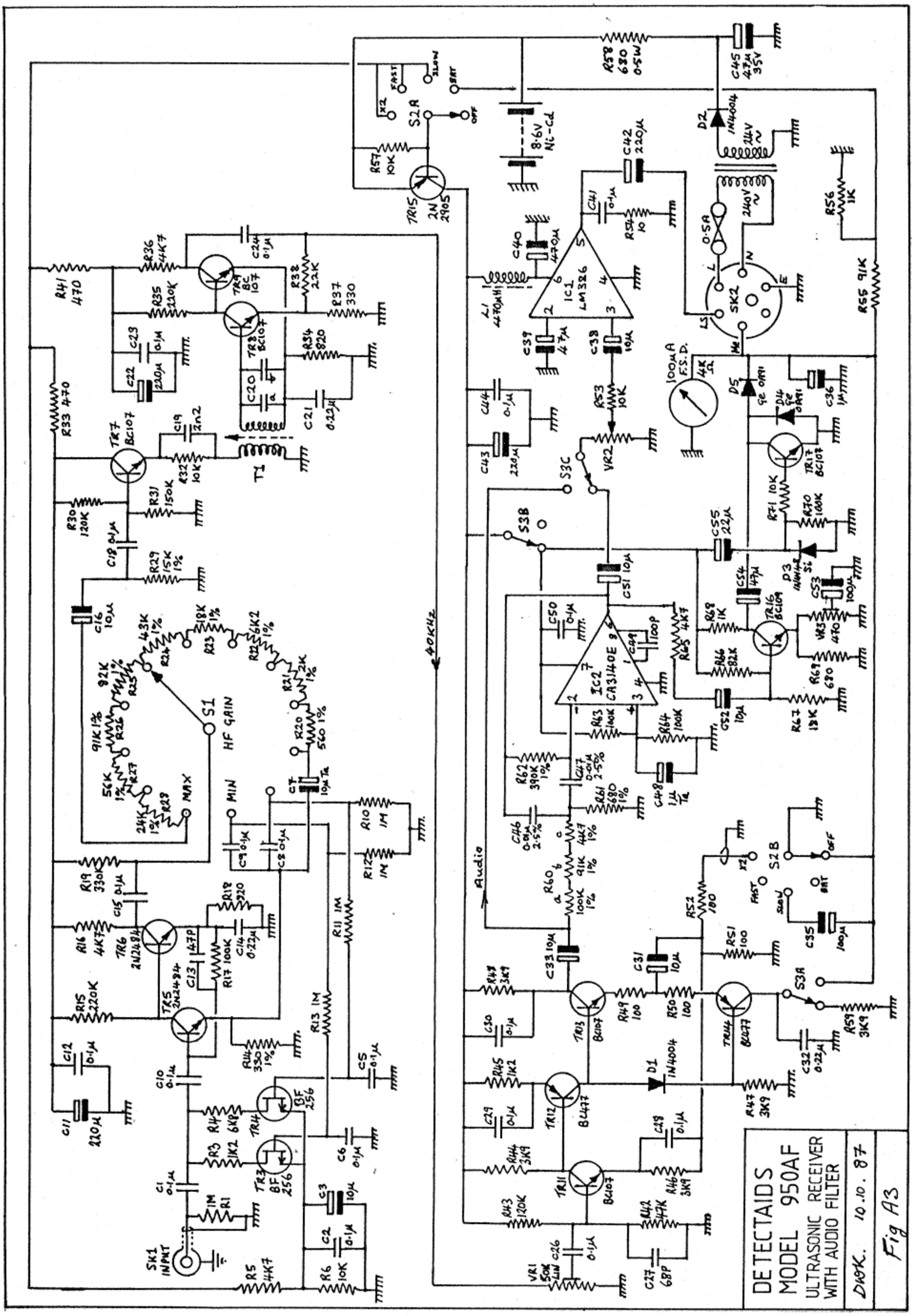
Fig A2

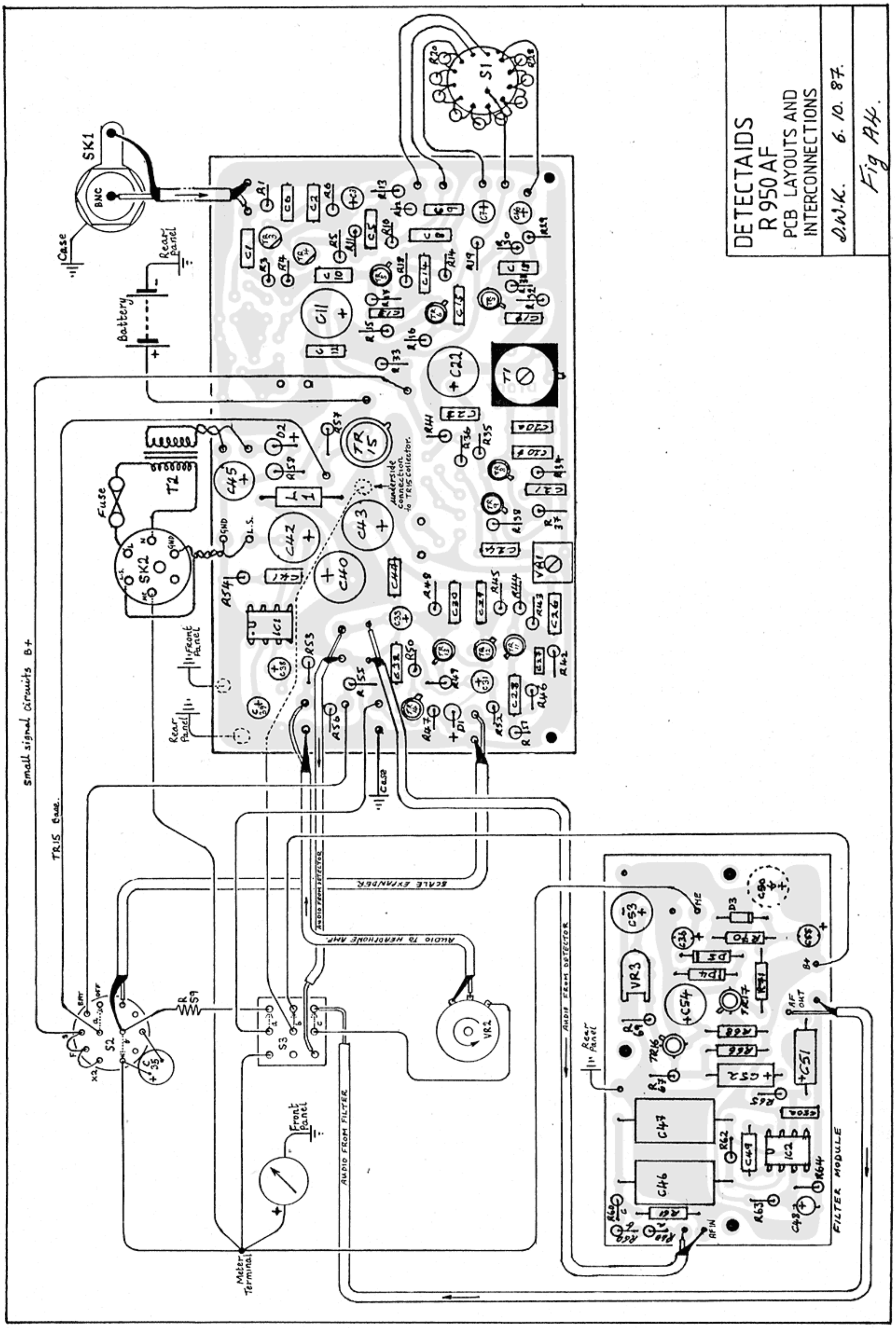
Switching the filter into operation is accomplished by means of a three-pole two-way switch S3. S3a terminates the detector circuit in R59. S3b powers the filter and VU meter circuits. S3c diverts the audio take-off point to the output of the filter. There is no need to switch the filter input. There is also no need to switch the VU circuit output, since this circuit has no effect unless it is powered.

The audio feed to the filter is taken from a patch, on the main circuit board, which was originally provided for C34 (deleted). Note also that there are two patches for supply decoupling on the filter board. These patches, designated C50a and C50b allow the board to be used in situations where there is not sufficient off-board supply decoupling. In the R950AF application, only C50a (0.1uF) is required. A larger decoupling capacitor is undesirable in this case, because it may give rise to a spark, on closure of S3b, and the instrument will then no longer be intrinsically safe.

Using the component values shown, the filter has a centre frequency of 977 Hz and a Q of 12. This corresponds to a bandwidth, at the half power points, of 81 Hz. Should requirements change, the filter can be re-calculated using the equations in figure A1. Note that R60 is made up of three resistors in series in this version. Additional patches are provided on the board such that R61 may be made up from two resistors in series if necessary. It is not normal practice to adjust this type of filter, it is better to adjust the transmitter modulation frequency. If filter adjustment becomes necessary, this can best be accomplished by padding C46 and C47 by equal amounts. The bandwidth of the filter however, is large enough that adjustment should be unnecessary even when the transmitter modulation frequency is fixed, as in the T900X.

The only production adjustment on the filter board is that of VR3 - VU meter sensitivity. This adjustment should be carried out by applying a steady modulated signal to the input of the receiver (Sk1) and rotating VR3 until the meter reading remains unchanged on switching the filter in and out. Such agreement however cannot be obtained at all points on the scale because the VU meter is non-linear and suffers from cramping at the bottom end. VR3 should therefore be set so that agreement is achieved at about 70 - 80% of full-scale deflection. Note incidentally, that when the filter is switched in, there is a short delay before the meter registers. This is due to the clamping effect of TR17. The delay period may be altered by changing C55, but if the value of this capacitor is made too small, the meter needle will be flicked onto the full-scale end stop as C54 charges.


DETECTAIDS
MODEL 950AF
ULTRASONIC RECEIVER
WITH AUDIO FILTER
DWG K. 10.10.87
Fig A3
HF GAIN
40KHz
Audio
SK1 INPUT
8.6V Ni-Cd
LM326
CA3140E


DETECTAIDS
R 950 AF
PCB LAYOUTS AND
INTERCONNECTIONS
D.W.K. 6. 10. 87.
Fig A4.
SK1
BNC
Case
Rear panel
Battery
T2
Fuse
SK2
Front Panel
small signal circuits B+
TR15 Base.
underside connection to TR15 collector.
SCALE EXPANDER
AUDIO FROM DETECTOR
AUDIO TO HEADPHONE AMP.
AUDIO FROM FILTER
FILTER MODULE
S1
S2
S3
VR2
Meter Terminal

Models T800 mkII, T900X and T900V.

The three new transmitters can all be built using the same circuit board. The T800 mkII is electrically identical to the original T800 but, by implementing its circuitry on the T900 board, the need to restock two types of transmitter board is eliminated. The new board is also correctly drilled for the current issue of transmitter case. Production details of the T800 mkII are given in figure A5.

The T900X and V retain the ultrasonic oscillator circuit of the T800, but use a different modulation system. The modulating audio signal is derived using a CMOS 4060B oscillator and frequency divider IC. This gives greatly improved modulation frequency stability and a very wide range of possible modulating frequencies.

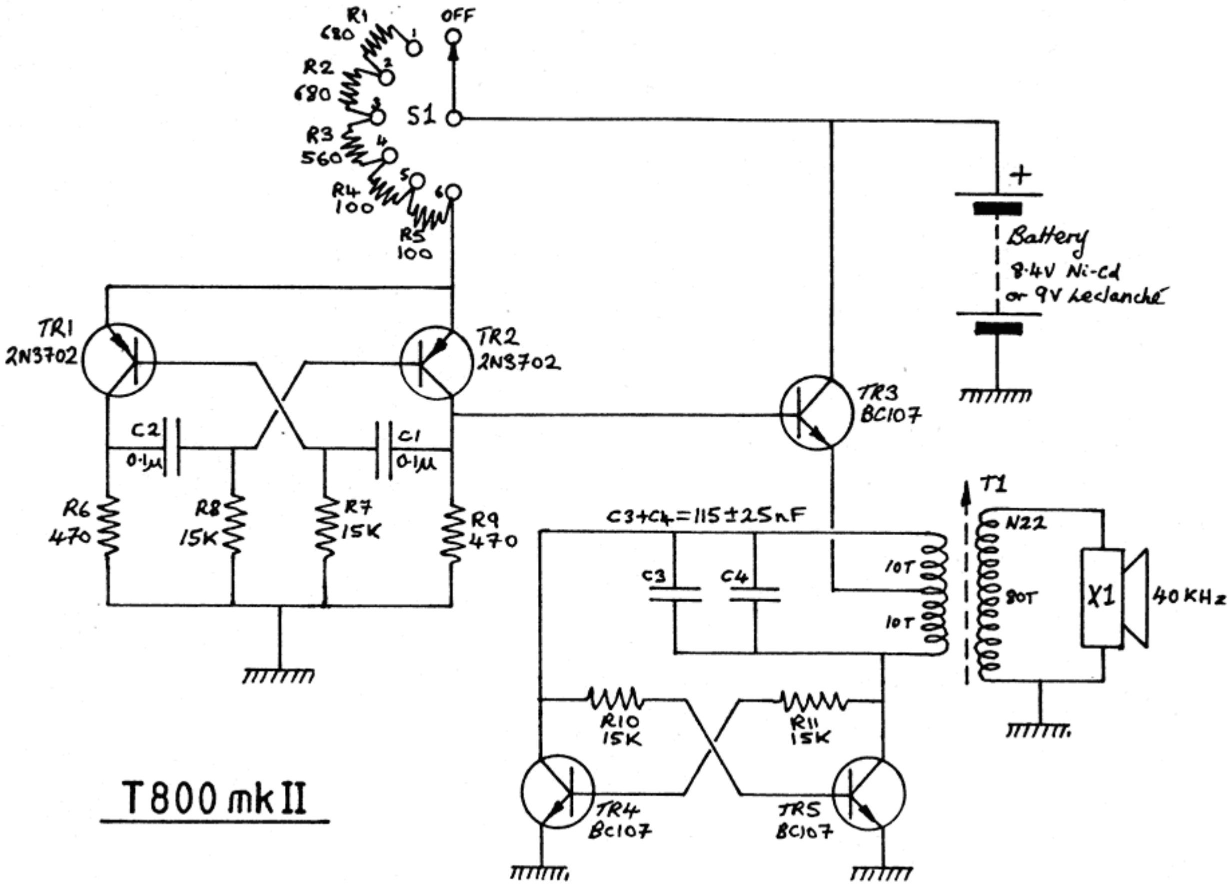
The T900X has its modulation frequency controlled by a quartz crystal. This gives a highly stable audio signal which is independent of supply voltage over a wide range. The unit may therefore be used with a conventional (Zn-MnO2) disposable battery and its modulation frequency will not change when high power is selected or when the battery is nearly exhausted. The disadvantage of the T900X is that its modulation frequency cannot easily be altered.

Production details of the T900X are given in figure A6. Patches are provided for wire-ended HC33/u crystals (0.5 inch pitch) and wire-ended HC18/u crystals (0.2 inch pitch). Crystals of less than 2MHz may be used, subject to size constraints, and the appropriate division ratio selected by tapping into the 4060 divider chain. In the version shown, a modulation frequency of 977Hz is achieved by using a 1MHz crystal and dividing by 1024. For other crystals, it may be necessary to alter C6 and C7 in order to obtain satisfactory oscillation. To change the division ratio, simply move R14 so that it connects to a different pin on the 4060. All divider outputs are provided with patches on the PCB. See figure A9 for details of the division ratios available.

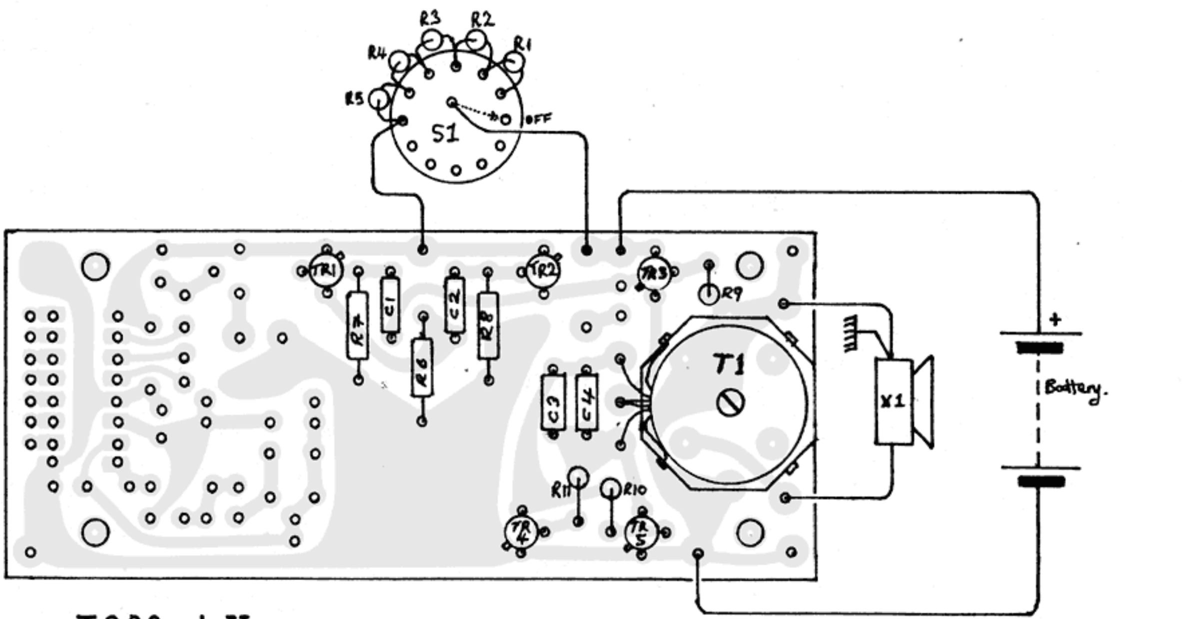
The T900V has its modulation frequency controlled by a free-running RC oscillator. This allows the frequency of the audio signal to be adjusted simply by means of a variable resistor, and allows the transmitter to be tuned exactly on to the peak in the receiver audio filter response. The disadvantage of this arrangement is that the audio frequency is affected by severe supply voltage fluctuations. For this reason, if the transmitter is to be used in conjunction with an audio filter, it is recommended that only Ni-Cd batteries are fitted.

Production details of the T900V are given in figure A7. In the version shown, a modulation frequency of 977Hz is achieved by setting the RC oscillator to 31.264KHz and dividing by 32. The nominal value for R17+VR1 is 9.69K in this case, and VR1 permits the modulation frequency to be varied from 650Hz to 2KHz. Design criteria for the RC oscillator are given in figure A9. It is recommended that R17 is not reduced below 4.7K. Low values of R17+VR1 increase the susceptibility to supply voltage variations.

General information on the CMOS 4060 IC is given in figures A8 and A9.



T800 mkII



T800 mkII

Fig A5

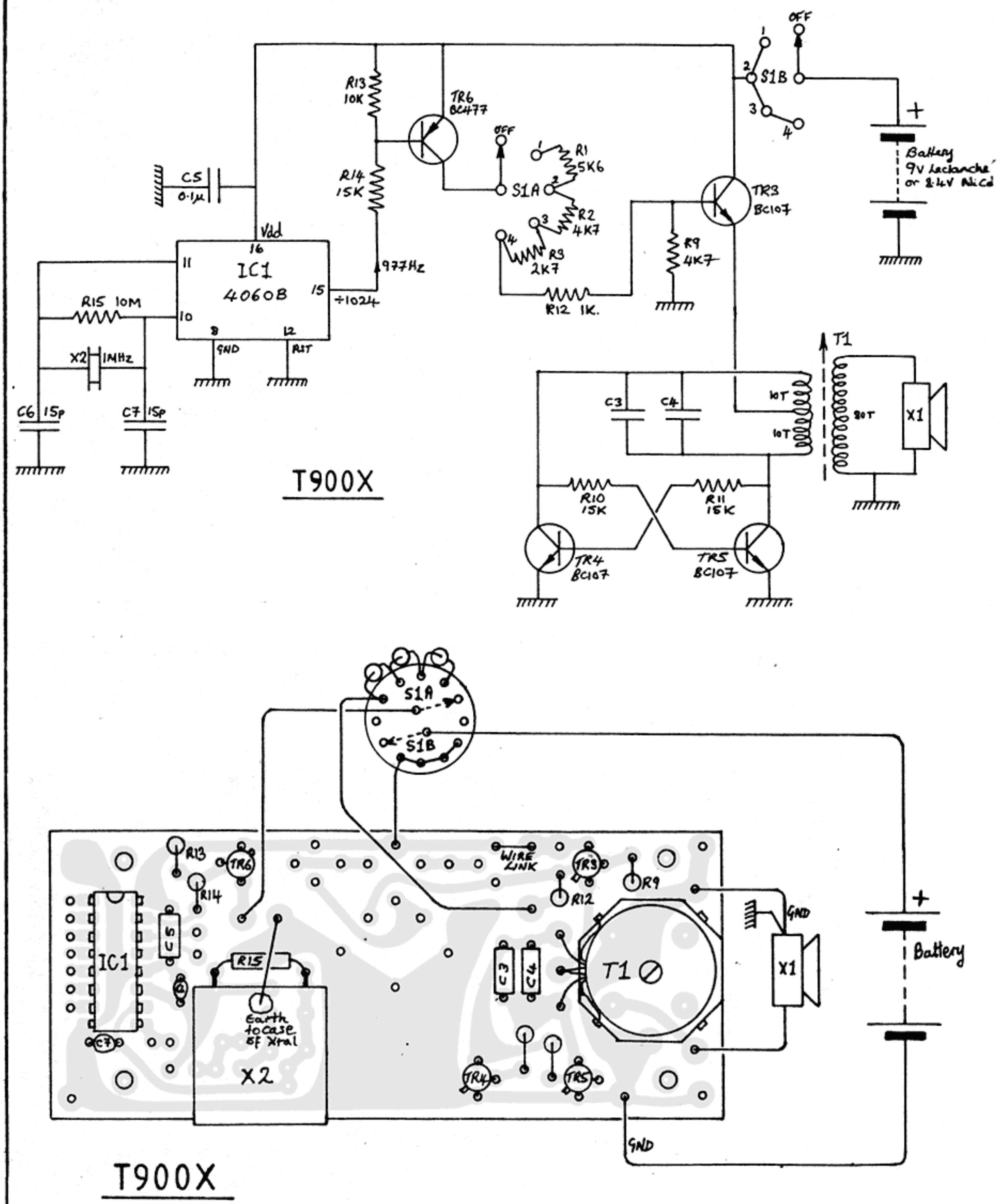
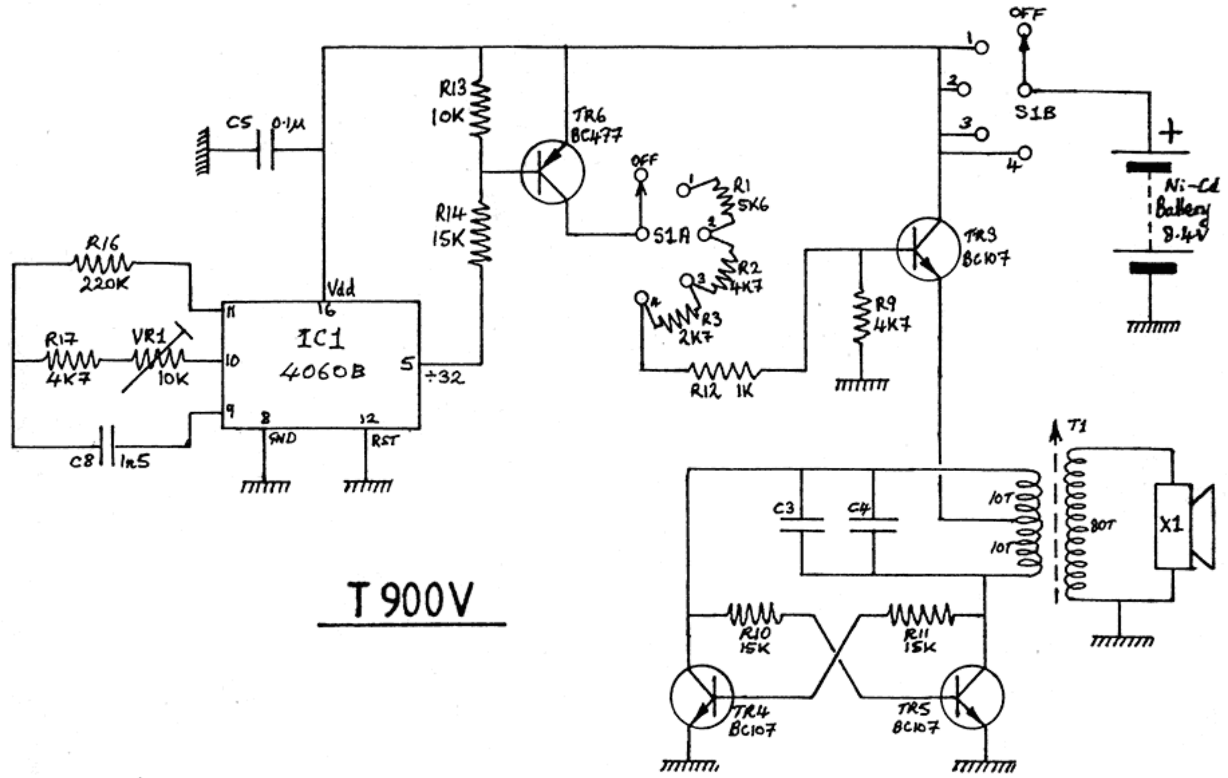

OFF
S1B
Battery 9V Leclanche' or 8.4V NiCd
R13 10K
TR6 BC477
R14 15K
C5 0.1µ
S1A
R1 5K6
R2 4K7
R3 2K7
R12 1K.
TR3 BC107
R9 4K7
Vdd
IC1 4060B
977Hz
÷1024
R15 10M
X2 1MHz
GND
RST
C6 15p
C7 15p
T1
10T
80T
X1
C3
C4
R10 15K
R11 15K
TR4 BC107
TR5 BC107
T900X
WIRE LINK
Earth to case of Xtal
Battery
GND
T900X

Fig A6


R13
10K
TR6
BC477
C5
0.1μ
R14
15K
OFF
S1A
R1
5K6
R2
4K7
R3
2K7
R12 1K
R9
4K7
TR3
BC107
S1B
OFF
Ni-Cd
Battery
8.4V
R16
220K
R17
4K7
VR1
10K
C8
1n5
Vdd
IC1
4060B
÷32
GND
RST
T1
10T
10T
80T
X1
C3
C4
R10
15K
R11
15K
TR4
BC107
TR5
BC107
T900V

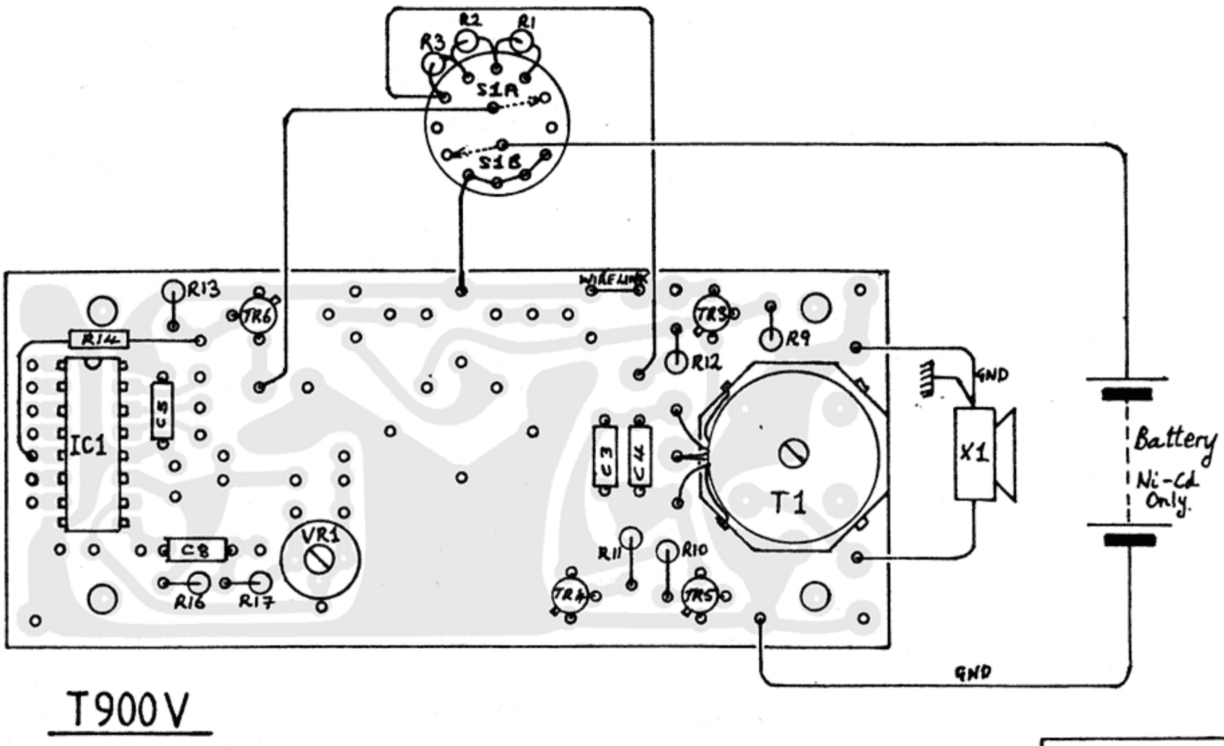

R3
R2
R1
S1A
S1B
R13
TR6
R14
IC1
C5
C8
VR1
R16
R17
WIRE LINK
TR3
R9
R12
C3
C4
T1
R11
R10
TR4
TR5
GND
X1
Battery
Ni-Cd
Only.
GND
T900V

Fig A7

4060 14-Stage (1 / 16384) binary ripple counter with internal oscillator

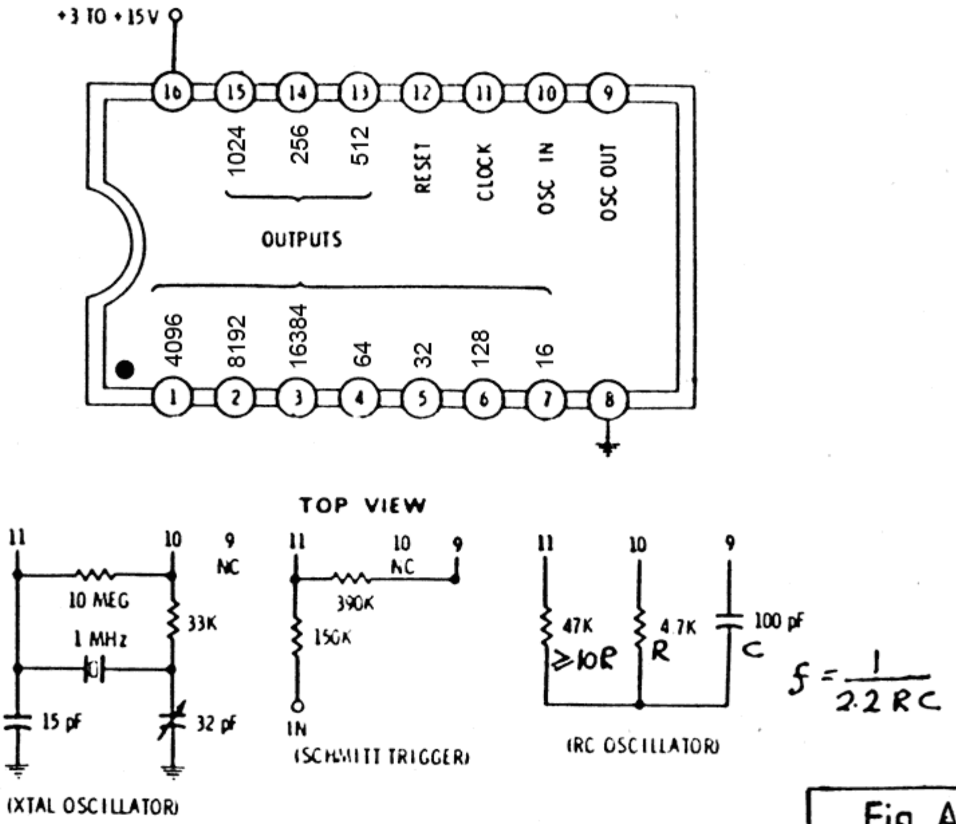
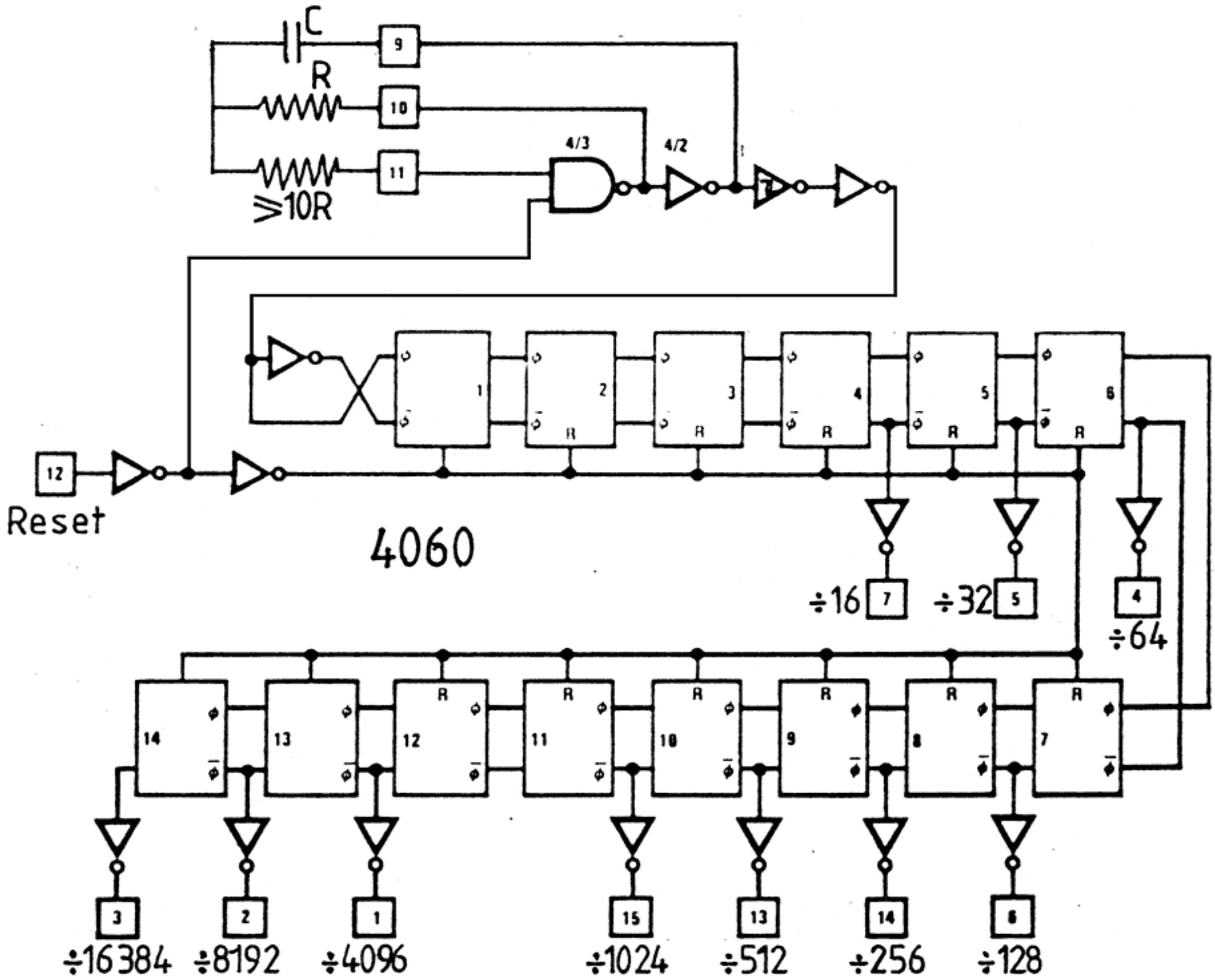
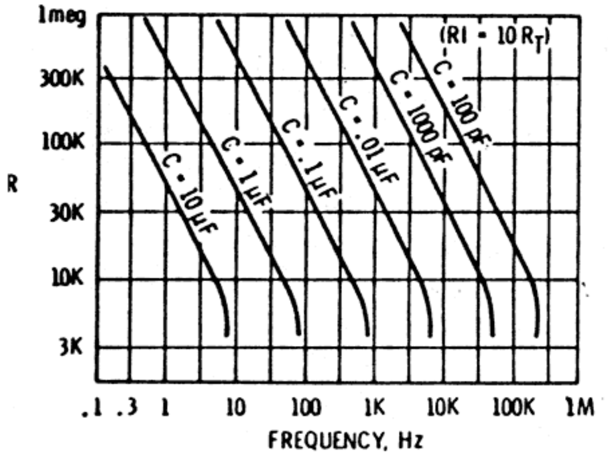


Fig A8

Note:

The CMOS 4060 counter has no outputs for divide by 2, 4, 8, and 2048.





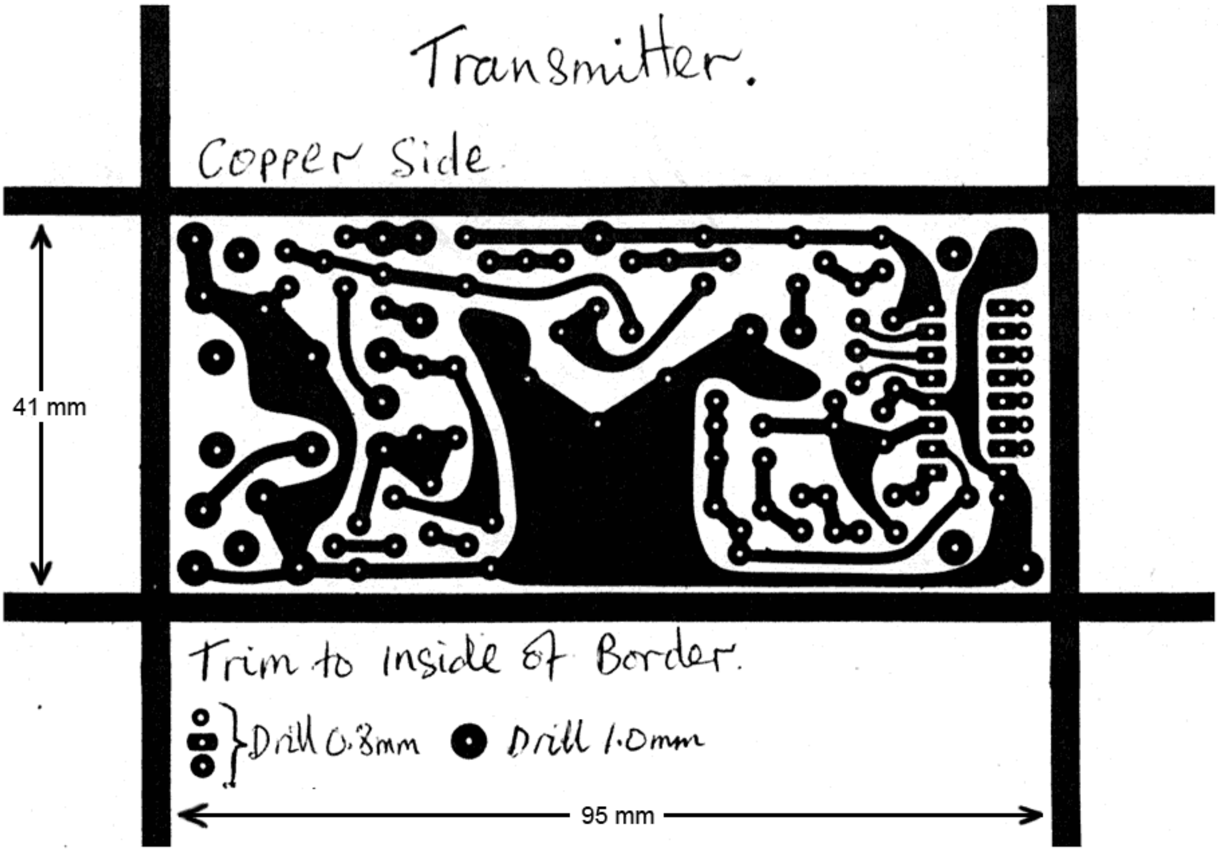
$$\text{FREQUENCY} = \frac{1}{2.2RC}$$

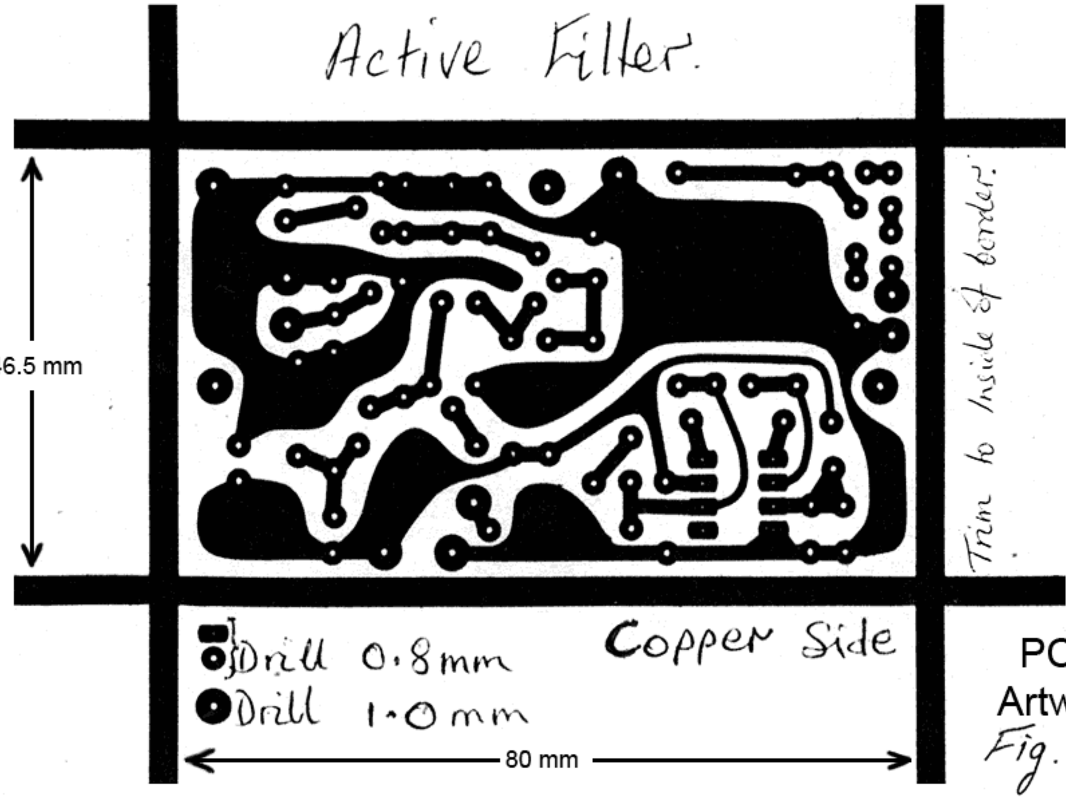
FOR C > 1000 pF

CMOS 4060

Internal Circuitry and external components for astable oscillator.

Fig A9





PCB
Artworks
Fig. A10

Filter Module Components List.

Part No.	Description		RS Stock No./Notes.
R60a	100K	1% MF	R60 = 196K 1%
b	91K	1% MF	
c	4K7	1% MF	
R61	680	1% MF	
R62	390K	1% MF	
R63	100K		
R64	100K		
R65	4K7		
R66	82K		
R67	18K		
R68	1K		
R69	680		
R70	100K		
R71	10K		
VR3	470		186-700 or equiv.
C36	1uF	63V rad.	104-051
C46	10nF	2.5% PS	113-409
C47	10nF	2.5% PS	113-409
C48	1uF	Ta	
C49	100pF	PS	113-263
C50a	0.1uF	Cer.	124-178
C50b	-		
C51	10uF	25V ax.	104-938
C52	10uF	25V ax.	104-938
C53	100uF	6.3V rad.	103-929
C54	47uF	10V rad.	103-941
C55	22uF	16V rad.	103-979
IC2	CA3140E		
TR16	BC109C		
TR17	BC107		
D3	1N4148 / 1N914 etc.		gen. purpose Si.
D4	OA90 / AA143 etc.		gen. purpose Ge.
D5	OA90 / AA143 etc.		

Transmitter Components List.

	T800	T900X	T900V	Notes.
R1	680	5K6	5K6	R1 - R5 may be replaced by a variable resistor with an on-off switch.
R2	680	4K7	4K7	
R3	560	2K7	2K7	
R4	100	-	-	
R5	100	-	-	
R6	470	-	-	
R7	15K	-	-	
R8	15K	-	-	
R9	470	4K7	4K7	
R10	15k	15K	15K	
R11	15K	15K	15K	
R12	-	1K	1K	
R13	-	10K	10K	
R14	-	15K	15K	
R15	-	10M	-	
R16	-	-	220K	
R17	-	-	4K7	
VR1	-	-	10K Lin.	
C1	0.1uF	-	-	
C2	0.1uF	-	-	
C3				Select on test.
C4				Select on test.
C5	-	0.1uF	0.1uF	
C6	-	15pF	-	
C7	-	15pF	-	
C8	-	-	1500pF	
TR1	2N3702	-	-	
TR2	2N3702	-	-	
TR3	BC107	BC107	BC107	
TR4	BC107	BC107	BC107	
TR5	BC107	BC107	BC107	
TR6	-	BC477	BC477	
IC1	-	CD4060B	CD4060B	16 Pin DIL socket.
X1				40KHz Transducer.
X2	-	1MHz	-	

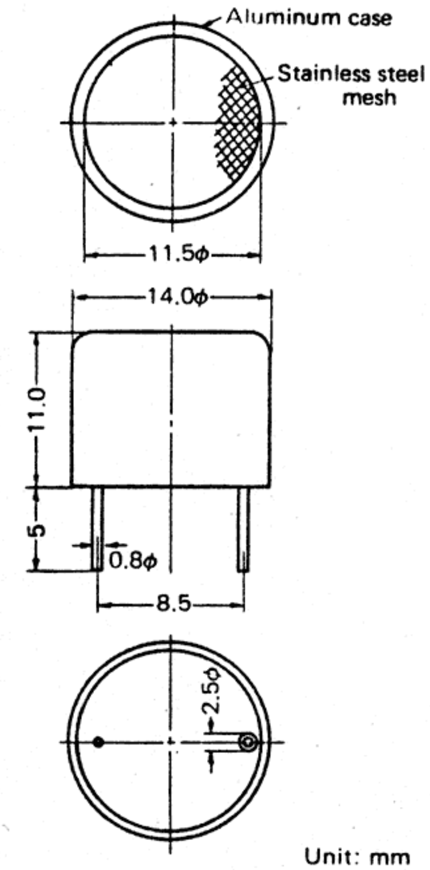
ULTRASONIC SENSOR
SE05B-40T, SE05B-40R

High-sensitivity ultrasonic transmitter SE05B-40T and receiver SE05B-40R are designed for sending and receiving ultrasonic sound through the air in the form of continuous wave or pulses.

Applications

Burglar alarm systems
Proximity switches
Liquid level meters
Anti-collision devices
Counters for moving objects
TV remote control systems

Shape and dimensions

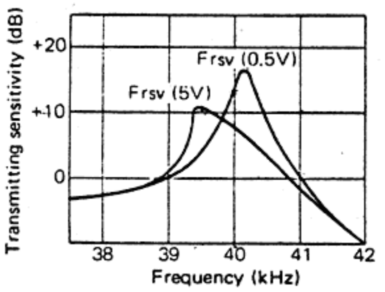


Characteristics

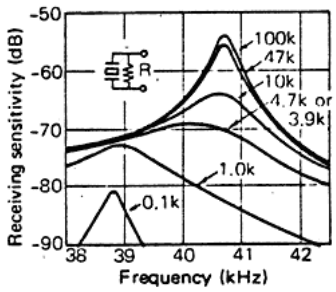
Item		Unit	-40T	-40R
Transmitting sensitivity	Sv	dB*1	17±6	–
Receiving sensitivity	Mv	dB*2	–	-56±6
Resonant frequency (transmitting)	Frsv	kHz*3	40±1	–
Resonant frequency (receiving)	Frmv	kHz*4	–	40±1
Directional angle	θ ½	°	Approx. 20	
Maximum input voltage		Vrms	7	–
Impedance		Ω	Approx. 200	Approx. 70K
Capacitance		pF	1400±20%	
Pulse rise time		msec.	2.0	0.5
Maximum input voltage for pulse operation		Vp-p	60	–
Temperature range		°C	-15 to +65	
Transmitting selectivity	Qsv		Approx. 70	–
Receiving selectivity	Qmv			Approx. 60

*1 OdB = 1μBar/V/meter
*2 OdB = 1V/μBar, measured with a shunt resistance of 47 KΩ.
*3 Frequency where transmitting sensitivity is maximum.
*4 Frequency where receiving sensitivity is maximum.

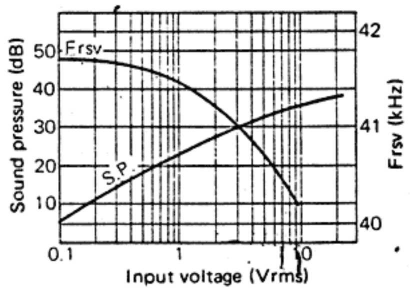
Frequency response (transmitting)
SE05B-40T



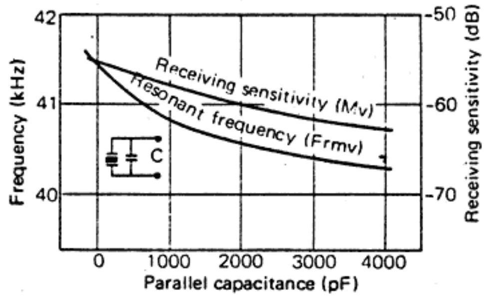
Frequency reponse (receiving) vs. shunt resistance
SE05B-40R



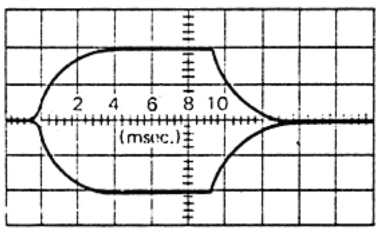
Sound pressure, resonant frequency vs. input voltage
SE05B-40T



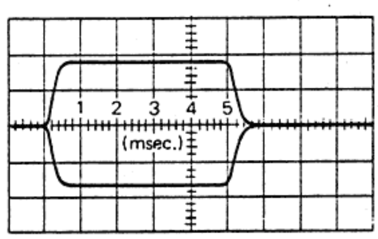
Effect of parallel capacitance
SE05B-40R



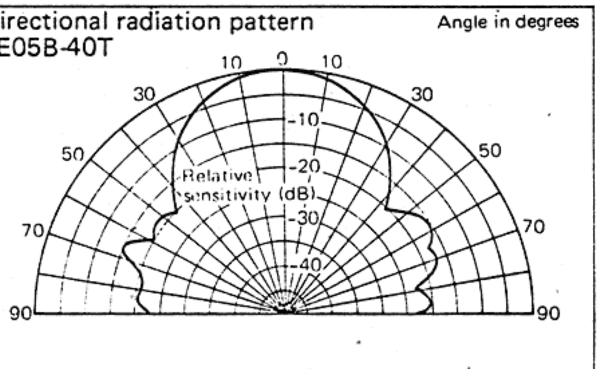
Pulse response (transmitting)
SE05B-40T



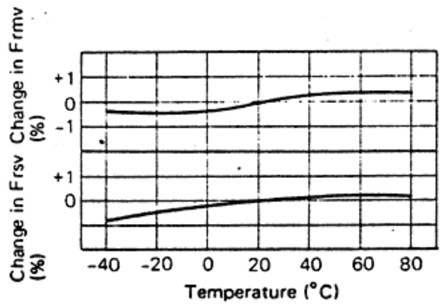
Pulse response (receiving)
SE05B-40R



Directional radiation pattern
SE05B-40T



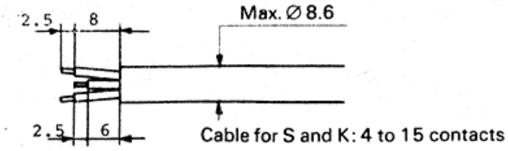
Temperature characteristics
SE05B-40R, 40T

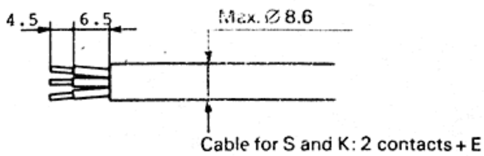


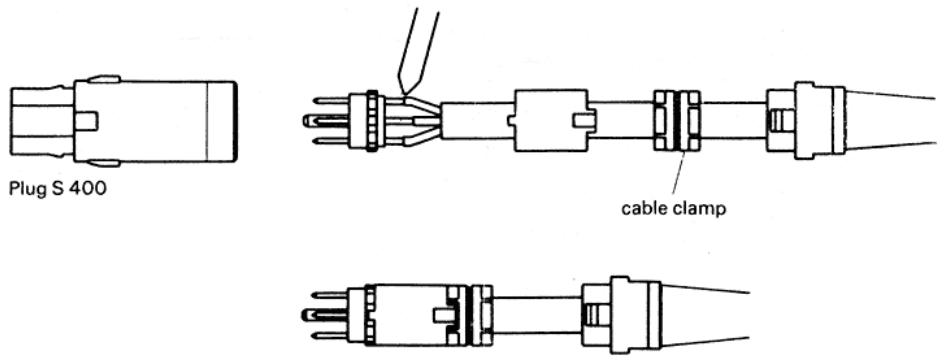
Headset plug S 400 A00 6-3
Socket D400 A00 6

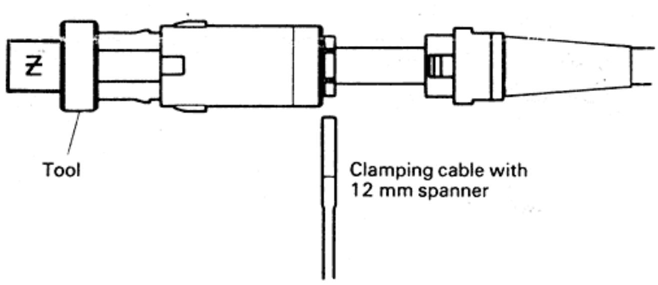
Plastic connector - 400 Series - Assembly instructions

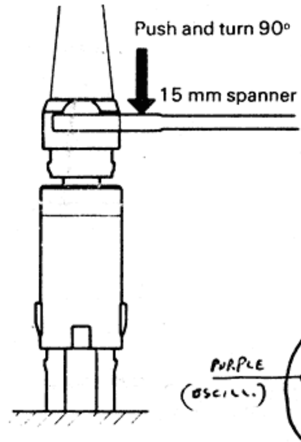
Preparation of cables

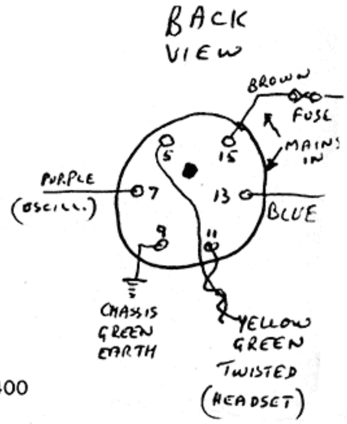


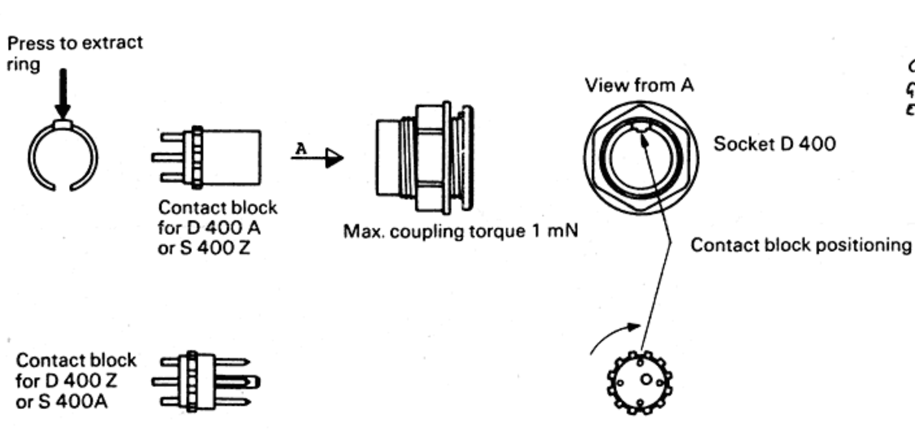












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