Chapter 4

PROTOTYPE DEVELOPMENT OF RF BANDWIDTH SWITCH

This chapter sets out the specifications of an RF bandwidth switch to be used in a T.B.T. system model, as well as the development of the electronics.

The prototype circuitry will first be specified in terms of its expected electrical inputs and outputs, where after each section will be specified and dealt with separately during the development of the electronics, namely the *combiner*, *the selector* and *the modulator* respectively.

The chapter is concluded with setup and test procedures.

4.1 SPECIFICATION OF RF BANDWIDTH SWITCH

4.1.1 Block diagram of RF bandwidth switch

Figure 4.1 shows the block diagram of the bandwidth switch. All electrical inputs and outputs are indicated. A separate ganged switching feature for composite video and audio is included.

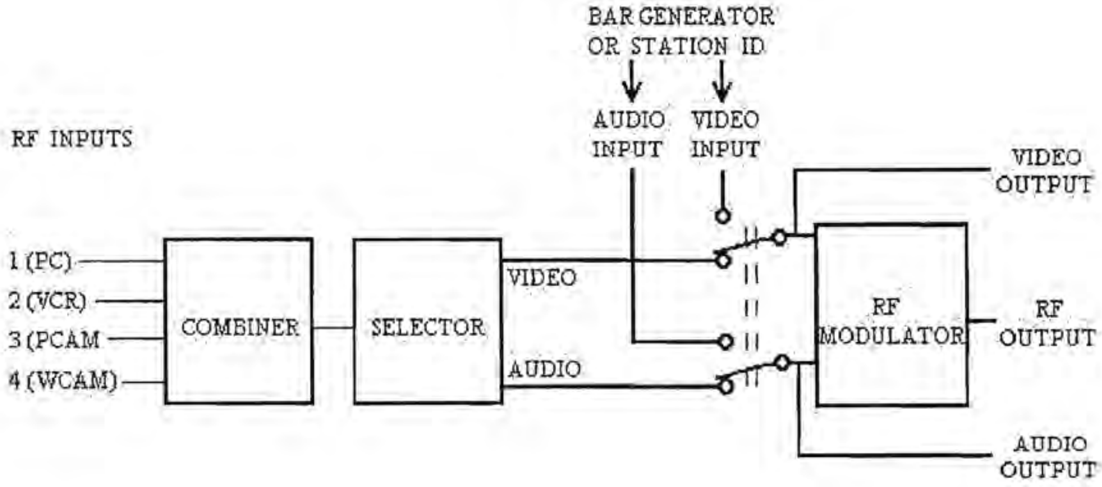


Figure 4.1. Block diagram of RF bandwidth switch.

The ganged switching arrangement provides for an additional video/audio input such as colour bar generator, or station ID (identification) screen, that may be generated by a PC, electronic generator or VCR. This feature is required to incorporate a method of transmission activation or deactivation, while maintaining warm standby condition at all times. For the purpose of the thesis it shall be known as the activation function switch.

4.1.2 Electrical inputs and outputs

Inputs and outputs for the prototype are specified in corresponding sections for video, audio and RF.

4.1.2.1 Video

Input voltage:	0,5 - 2,0 Vp-p
Input impedance:	75 ohm
Output voltage:	1 Vp-p, typically
Output impedance:	75 ohm

4.1.2.2 Audio

Input level:	-10 dBm
Output level:	-5 dBm
Output impedance:	High, typically 600 ohm
Frequency response:	50Hz - 10kHz

4.1.2.3 RF (Radio Frequency)

Input RF bandwidth:	174-254 MHz and 470-854 MHz
Input channel bandwidth:	6MHz
Input RF level:	60 dB/μV to 90 dB/μV
Input impedance:	75 ohm
Number of inputs:	Minimum 4
Output RF bandwidth:	Ch 30 - Ch 39. Typ Ch 36
Output RF level:	Typically 75 dB/μV
Output impedance:	75 ohm
Overall S/N:	better than 40 dB
Noise figure:	better than 3 dB

4.1.3 **Combiner**

The function of the combiner is to accept a minimum of four inputs of RF bandwidth at separate channel frequencies, and combine them to a common broadband output.

The four input channels are allocated for most common sources of media; a video cassette recorder (VCR), a personal computer (PC) and two camera inputs (CAM1/2). An auxillary channel for electronic whiteboard input is nice to have, as well as a commercial television input. Specifications are as follows:

Type:	broadband transformer type
Frequency range:	174 MHz to 854 MHz
Characteristic impedance:	75 ohms
Insertion loss:	less than 14dB across the band.

4.1.4 Selector

The function of the selector is to pre-select any one of a minimum number of four RF inputs, and to extract the modulated video and audio information. To achieve high selectivity and sufficient adjacent channel rejection, combined with relative broadband operational capabilities, the selector requires two integrated RF superheterodyne receiver sections in a single tuner section for signal selection. A common intermediate frequency (IF) section will extract the vision intermediate frequency (VIF) and the sound intermediate frequency (SIF), and consequently the composite video and audio components of the signal. The selector will require AFC (automatic frequency control) and AGC (automatic gain control) circuitry, as well as a sophisticated IF filter to keep the IF stable and at a constant level. This is a crucial requirement to ensure satisfactory operation.

4.1.5 RF modulator

A wide range of modulators designed for use in AV and cable TV applications is available in the industry. It is believed that a suitable unit may be easily obtained which requires little or no modification to be utilised in the prototype.

4.1.6 Power supply requirements

The RF bandwidth switch is specified to operate from a single 12VDC power source. The reason for this is twofold: the latest manufacturers' trend is to design electronic equipment to run from external DC supplies, and supporting this trend, renders manufactured equipment to be world standardised even though AC mains voltages differ between countries.

4.2 DEVELOPMENT OF THE ELECTRONICS

The combiner is regarded as the nerve centre of the RF bandwidth switch. In the development of the electronics for the RF bandwidth switch, this study will therefore be focused upon the combiner, selector and modulator in the descending order of their importance.

For the purpose of explanation, it was decided to extract most relevant information from a literature study conducted by the student, to be presented in Annexure A to this dissertation.

4.2.1 Hybrid combiner for RF bandwidth switch

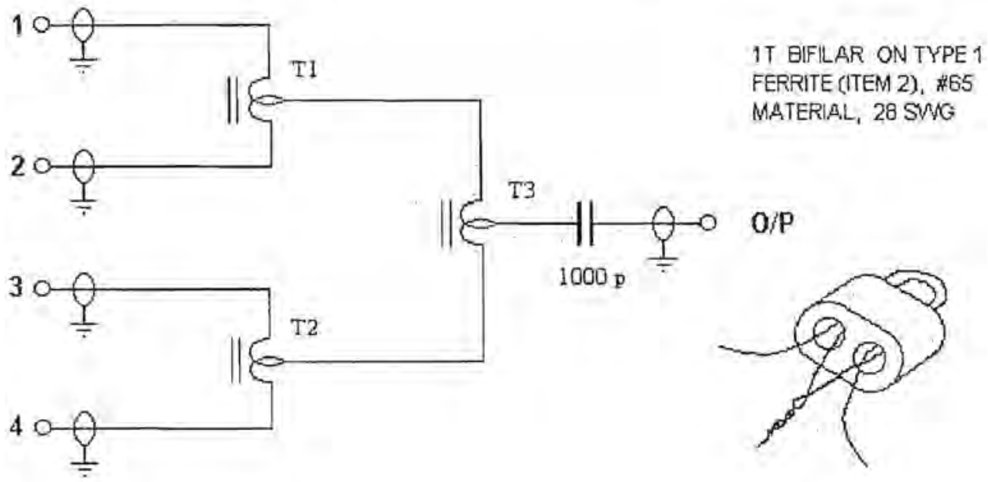


Figure 4.2 Cascaded hybrid for RF bandwidth switch.

The circuit initially considered, as depicted in Figure 4.2, provides for three broadband transmission line transformers, accepting four PAL I RF inputs in the frequency range as specified. Inputs are evenly spaced across the RF bandwidth, spaced at two channel intervals.

The output is via a 1000pF capacitor to eliminate unwanted DC components that may be present on the input signal. The minimum recommended RF signal level for PAL system I is 60-dB/μV. Most commercial modulators have their RF output levels set to around 80-dB/μV.

Four commercial modulators were obtained for test purposes, and their outputs were measured at 80, 76, 74 and 81 -dB/μV respectively, which yields an average of around 78-dB/μV. The maximum acceptable insertion loss across the combiner circuitry would therefore be around 18-dB.

The inductance of the single broadband transformer had to be determined at the low end of the operating frequency band, as well as at the high end of the same. To determine the inductance at the low end of the band, the inductance was measured with a Hewlett-Packard 4260A universal LCR bridge. The measurement indicated an inductance of 0,2-uH, expected accuracy within 20%. The inductive reactance X_ℓ may now be calculated for the low end of the operating band (170 MHz), where $X_\ell = 2.\pi.f.\ell$, which yields 213,6-ohm.

At high frequency, and since the effects of the ferrite vanish electrically at high frequency, the inductance was first calculated using the formula [12]:

$$L(\mu H) = \frac{d^2 n^2}{18d + 40\ell} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots (1)$$

where L = inductance in microhenrys, ℓ = coil length in inches, n = number of turns, and d = coil diameter in inches. Substituting with $d = 0{,}16"$, $\ell = 0{,}05"$, and $n = 2$, then $L = 0{,}021$-μH.

The calculation was confirmed by connecting a known value silver mica capacitor (0,47-pF 5%) across the inductor, and measuring the resonant frequency (519-MHZ) of the tuned circuit with a grid dip meter. Since $f_r = \frac{1}{2\pi\sqrt{LC}}$, the value of L = 0,02-μH. Accuracy of this measurement is subject to the tolerance of the known value capacitor, rated at 5%. This value for L equates to an inductive reactance of 88-ohms at the high end of the band.

To determine the effects of insertion loss across the combiner, two measurements were conducted. The insertion loss characteristics vs. frequency for a single transformer was firstly determined, where after the insertion loss vs frequency across the cascaded transformers, i.e. complete combiner was measured.

4.2.1.1 Insertion loss across a single broadband hybrid.

The transmission characteristics for a single hybrid is depicted in Fig. 4.3.

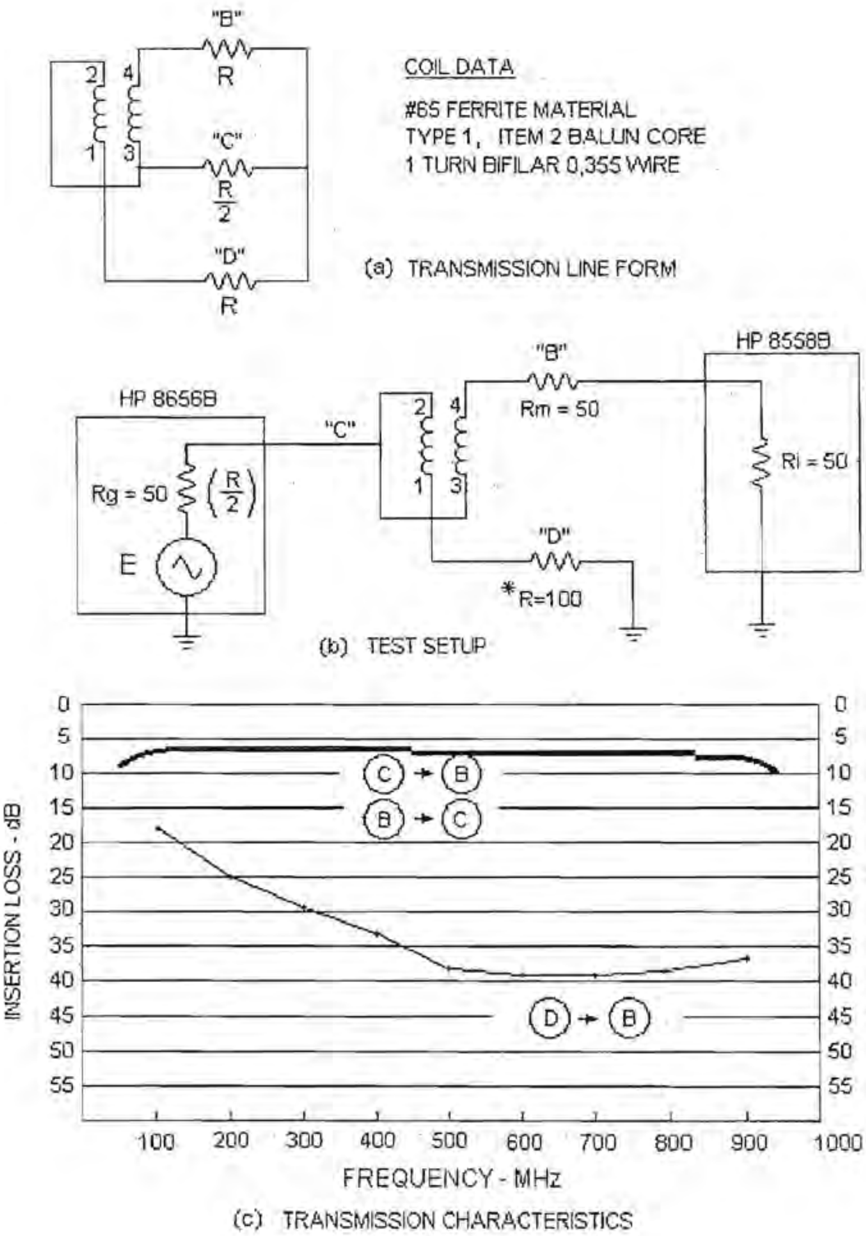


Figure 4.3. Single hybrid combiner : Insertion loss vs. frequency.

The response of the hybrid in Fig. 4.3.a is depicted in Fig. 4.3.c. For this measurement R = 100-ohms. In order to measure the hybrid in a 50-ohm circuit, arms B and D were measured with a 50-ohm resistor in series with the 50-ohm measuring gear. This accounts for 3 dB additional loss observed. The return loss between arms C and B is tabulated below:

Transmission characteristics C to B for single hybrid with arm D terminated in 100-ohms.									
Insertion loss (dB)	-7	-6	-6	-6	-6,2	-6,2	-6,5	-7	-8
Frequency (MHz)	100	170	340	430	510	600	680	840	920

Table 4.1

The test set-up was calibrated by setting the HP8656B generator output to 0dBm at 100-MHz, and the reading observed as 0dBm on the HP8558B measuring gear. This reading was confirmed by measuring the true RMS voltage of the generator output with a RACAL-DANA 9302 RF-millivolt meter, which yielded 223 mV-RMS.

Set at 170-MHz, the 50-ohm series resistor R_m was installed at the input port of the HP8558B, and the reading observed as -3dBm. Next, the generator output was fed to port C and the response observed as -6bBm at port B, with port D terminated in 100-ohms.

Measurements made at 170-MHz, 340-MHz and 430-MHz all yielded -6dBm. At 510-MHz the reading is tabulated as -6,2dBm, also at 600-MHz. The response falls to -6,5dBm at 680-MHz, -7dBm at 840-MHz, and to -8dBm at 920-MHz. The measurements were repeated with arm D open circuit:

Transmission characteristics C to B for single hybrid with arm D open circuit.									
Insertion loss (dB)	-6	-6,5	-7,5	-7	-6,5	-6	-6,5	-7	-7,5
Frequency (MHz)	100	170	340	430	510	600	680	840	920

Table 4.2

Isolation between arms B and D were measured with arm C terminated in 50-ohms. For this measurement a 50-ohm resistor was connected in series with the 50-ohm generator as well, as shown in Fig 4.3.d, similar to the series resistor used with the HP8558B in Fig. 4.3.

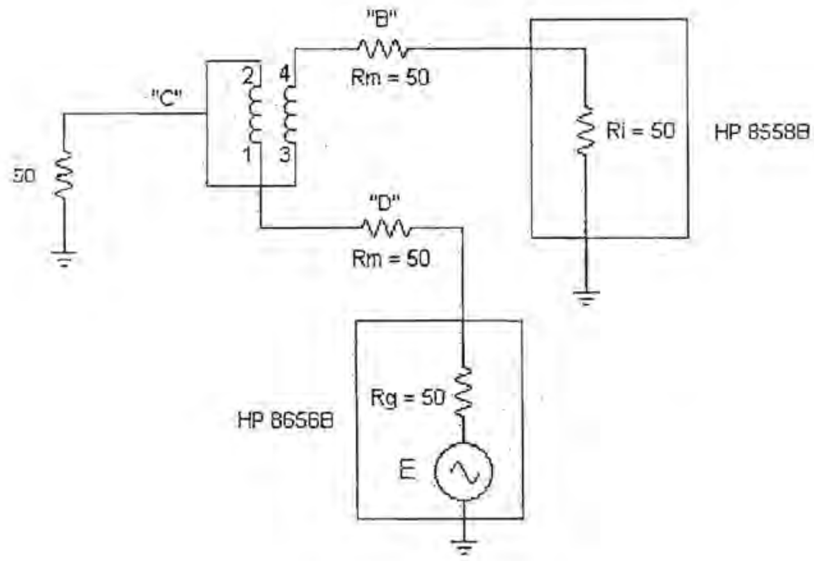


Fig. 4.3.d. Test setup for isolation measurement

The results are tabulated below:

Isolation characteristics between B and D for single hybrid with arm D = 50-ohms.									
Isolation (dB)	18	25	29	34	38	39	39	38	37
Frequency (MHz)	100	200	300	400	500	600	700	800	900

Table 4.3

It was noted that the 50-ohm termination resistor's value at port C is critical to obtain optimum isolation between ports B and D. Adjusting this value by 50% reduces the isolation by approximately 15 dB in the high frequency operating range between 400 and 600-MHz.

4.2.1.2 Insertion loss across cascaded broadband hybrid.

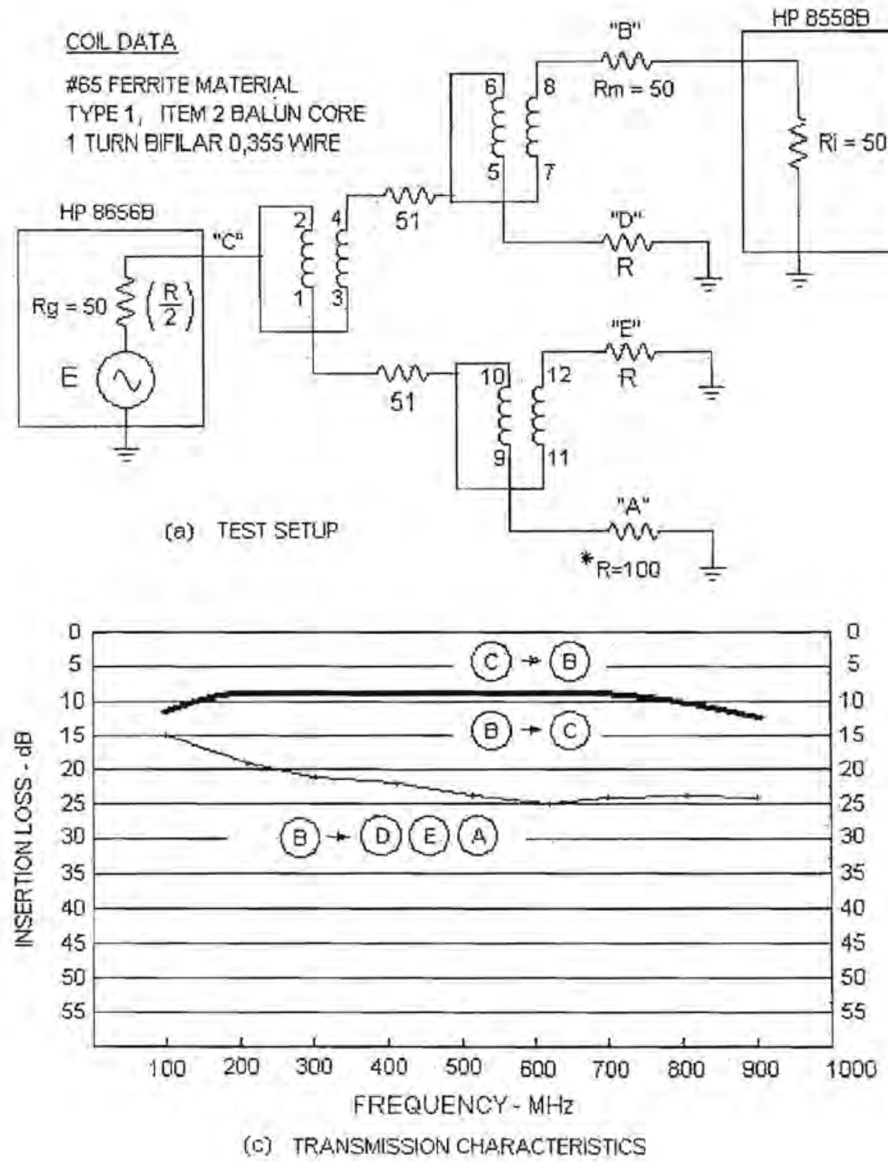


Figure 4.4 Cascaded hybrid combiner

As shown in Figure 4.4, the measurements were conducted with the D, E and A arms terminated in 100-ohms, and repeated with the inputs open circuit. Series resistors (51-ohm chip resistors) were inserted between cascaded sections to prevent mismatched operation. The resistors were selected for their low noise characteristics, but were unfortunately not available in 50-ohm.

Transmission characteristics C to B for cascaded hybrid with arm D, E and A = 100-ohms.									
Insertion loss (dB)	11	9	9	9	9	9	9	10	12
Frequency (MHz)	100	200	300	400	500	600	700	800	900

Table 4.4

The measurements were repeated with arms D, E and A open circuit:

Transmission characteristics C to B for cascaded hybrid with arms D, E and A open circuit.									
Insertion loss (dB)	9	8	7,5	8	8,5	8	8	10,5	12,5
Frequency (MHz)	100	200	300	400	500	600	700	800	900

Table 4.5

Isolation between arms B and DEA were measured with arm C terminated in 50-ohms. As with the isolation measurement for a single hybrid, a 50-ohm resistor was connected in series with the generator as well. The results are tabulated below:

Isolation characteristics between B and D for cascaded hybrid with arm C = 50-ohms.									
Isolation (dB)	15	19	21	22	24	25	24	23	24
Frequency (MHz)	100	200	300	400	500	600	700	800	900

Table 4.6

4.2.1.3 Recurring pattern mismatch

The results depicted in Fig. 4.3 appears to conform to the theory described in Annexure A to the thesis, until the transmission characteristics are measured at 10-MHz intervals. The results may be described as a near sinusoidal recurring pattern that appears to be superimposed upon the transmission characteristic curve of insertion loss vs. frequency.

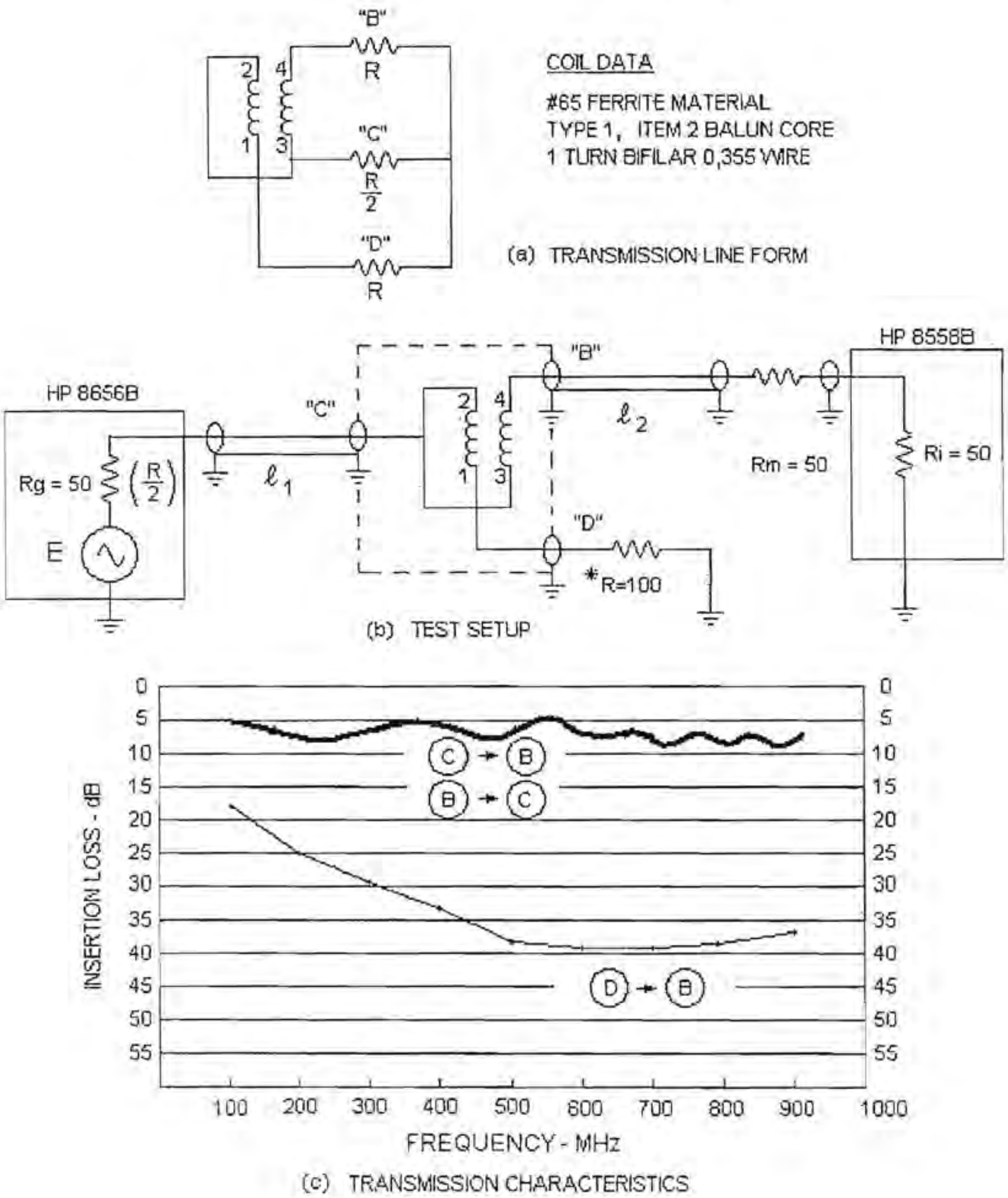
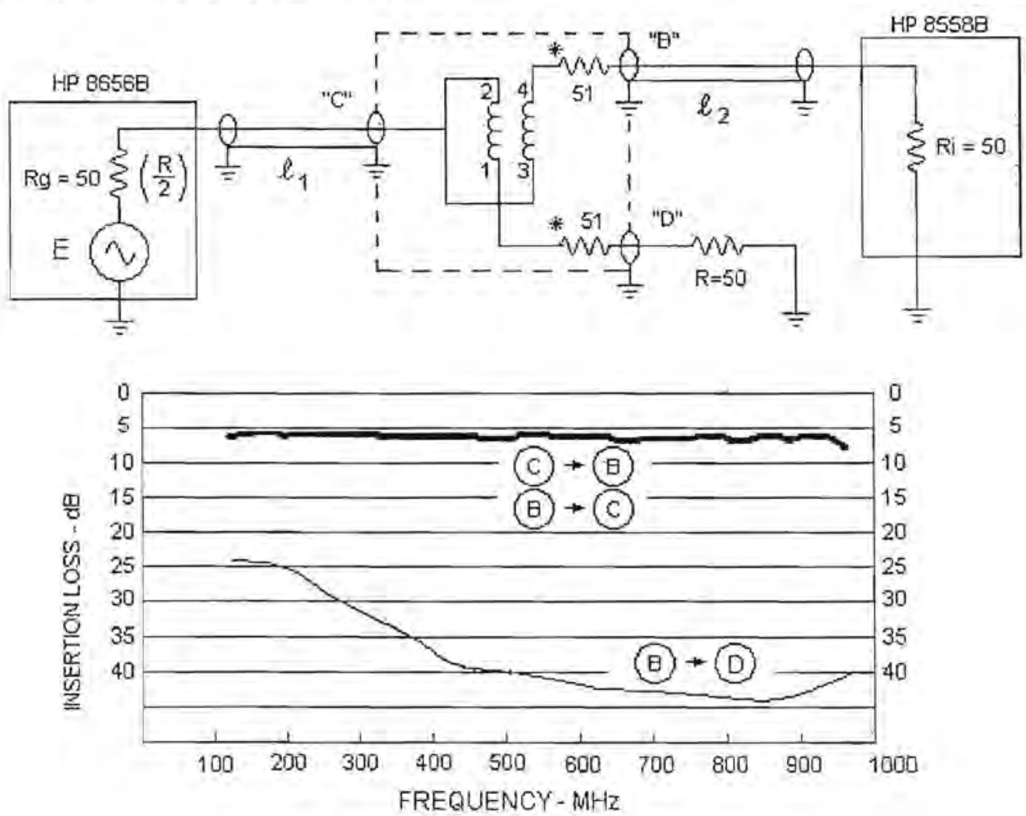


Fig.4.5 Recurring pattern mismatch

As seen in Fig. 4.5, the series resistor R_m was placed in line with the HP8558B, approximately 25-cm (the length of ℓ_2) away from the hybrid. The resistor was realised by placing a 51-ohm resistor inside a 20-mm length of brass tubing, with a male and female BNC connectors at the ends. The apparent frequency of variation of the pattern lowered by shortening ℓ_2 to 15-cm, as did the apparent deviation of the recurring pattern.

The series resistor was then moved back towards the hybrid, and finally mounted inside the hybrid enclosure, with optimum results. The resistor finally used for this purpose was a 51-ohm chip resistor, and was duplicated in the D arm of the hybrid. The device could now be measured with standard 50-ohm measuring gear, without additional matching of load to generator, as shown in Fig. 4.6 below:



(b) TRANSMISSION CHARACTERISTICS

Fig. 4.6 Matched 50-ohm hybrid

Results are tabulated under column 1 in Table 4.7. The key to the table follows at the end.

Table 4.7. Transmission characteristics

f (MHz)	1	2	3	4	5	6	7
130	-5,5	-5	-24	-12	-10	-27	-5,5
140	-6	-5	-25	-12	-10,5	-27	-6
150	-6	-5,5	-25,5	-12	-10,5	-28	-6
160	-6	-6	-26	-12	-11	-28	-6,5
170	-6	-6	-27	-12,5	-11	-28	-7
180	-6	-6	-27,5	-12,5	-11	-28	-7
190	-5,5	-6	-28	-12	-11	-28	-7
200	-6	-5,5	-28	-12	-10,5	-27,5	-6,5
210	-5,5	-6	-28,5	-11,5	-10,5	-27	-6,5
220	-5,5	-6	-29	-11,5	-10,5	-27,5	-6
230	-5,5	-6	-29,5	-12	-10,5	-27,5	-6
240	-5,5	-6	-30	-12	-10,5	-28	-6
250	-5	-5,5	-30	-12	-11	-27,5	-6
260	-5	-5,5	-30	-12	-10,5	-27	-6,5
270	-5	-5,5	-30,5	-11,5	-11	-27	-6,5
280	-5	-6	-31	-11,5	-11	-26,5	-7
290	-5	-6	-32	-11,5	-11,5	-27	-7
300	-5	-6	-32,5	-12	-11	-27	-7
310	-5,5	-6	-33	-12	-11	-27,5	-7
320	-5,5	-6	-33,5	-12	-10,5	-28	-7
330	-5,5	-6	-34	-12	-10,5	-28	-7
340	-6	-6	-34,5	-12,5	-11	-28	-7,5
350	-6	-6	-34,5	-12,5	-11	-28	-7,5
360	-6	-6	-35	-12,5	-12	-27,5	-8
370	-6	-6,5	-36	-12,5	-12,5	-27,5	-8,5
380	-6	-7	-37	-12,5	-12,5	-28	-8

f (MHz)	1	2	3	4	5	6	7
390	-6	-7	-37,5	-12,5	-12,5	-28	-8
400	-6	-6,5	-38	-12	-12	-28	-8
410	-6	-6,5	-38,5	-12,5	-12	-28	-8
420	-6	-6	-38,5	-12	-11,5	-27,5	-7,5
430	-6	-6	-39	-12	-11,5	-27,5	-7,5
440	-5,5	-6	-39,5	-12	-12	-27	-8
450	-6	-6	-40	-12	-12	-27	-8
460	-6	-6	-40,5	-12	-11,5	-28	-8
470	-6	-6	-40,5	-12,5	-11	-28	-7,5
480	-6	-6,5	-40,5	-12,5	-11	-28	-8
490	-6	-6	-40,5	-13	-11,5	-28	-8
500	-6,5	-6	-40,5	-13	-12	-27,5	-8,5
510	-6	-6	-40,5	-12,5	-12,5	-27	-8,5
520	-6	-6	-41	-12	-12	-26	-8,5
530	-5,5	-5,5	-42	-12	-12	-26	-8
540	-5,5	-5,5	-42,5	-11,5	-11,5	-26,5	-8
550	-5,5	-5,5	-42	-12	-11	-27	-7,5
560	-6	-5	-42	-12	-11	-27	-7,5
570	-6	-5	-41	-12,5	-12	-27	-8
580	-6	-5	-40,5	-12,5	-13	-27	-9
590	-6	-5	-40,5	-13	-13,5	-26,5	-10
600	-6	-5,5	-41	-13	-14	-26	-10
610	-6	-5,5	-41	-12,5	-14	-26	-10
620	-6	-5,5	-42	-12,5	-14	-26	-9,5
630	-6	-5	-42	-12,5	-13,5	-26,5	-9,5
640	-6	-5	-42	-13	-14	-27	-10

f (MHz)	1	2	3	4	5	6	7
650	-6,5	-5,5	-42	-13,5	-14,5	-27	-10,5
660	-6,5	-6	-41,5	-13,5	-15	-26	-11
670	-6,5	-6	-42	-13	-15,5	-26,5	-12
680	-6	-6	-42	-13	-16	-25	-12
690	-6	-6	-43	-13	-15,5	-25,5	-11,5
700	-6	-6	-43,5	-13	-15	-26	-11
710	-6	-6	-44	-13	-15	-26	-11
720	-6,5	-6,5	-44	-13,5	-15,5	-26,5	-12
730	-6,5	-6,5	-44,5	-13,5	-16	-26,5	-12
740	-7	-7	-45	-14	-16	-26	-12,5
750	-6,5	-7	-46	-13,5	-16	-25,5	-12,5
760	-6	-7	-47	-13	-15	-25,5	-12
770	-6	-6,5	-48	-12,5	-14	-26	-10
780	-6	-6,5	-48	-12,5	-13,5	-26,5	-9,5
790	-6	-6	-48	-12,5	-13,5	-27	-9,5
800	-6	-6	-47,5	-13	-14	-27,5	-10,5
810	-6,5	-6	-47	-13	-15	-27,5	-11
820	-6,5	-6	-46,5	-13,5	-15,5	-27	-12
830	-6,5	-6	-47	-13	-15,5	-26,5	-12
840	-6	-6	-48	-13	-15	-26,5	-11
850	-6	-6	-48	-12,5	-14,5	-27	-10,5
860	-6	-6	-47,5	-12,5	-14,5	-27,5	-10,5
870	-6	-5,5	-47	-13	-15	-28	-11
880	-6,5	-5	-46	-13	-16	-28	-12
890	-6,5	-5	-45	-13	-16	-27	-12,5
900	-6	-5	-44	-13	-16,5	-26,5	-12,5

f (MHz)	1	2	3	4	5	6	7
910	-6	-5	-44	-13	-16	-26,5	-12
920	-5,5	-5,5	-43,5	-12,5	-15,5	-26	-12
930	-6	-5,5	-43	-13	-15,5	-27	-11,5
940	-6,5	-6	-43	-13,5	-16	-28	-12
950	-7	-6	-43,5	-14	-17	-29	-13
960	-7,5	-6,5	-43	-14,5	-18	-29	-14
970	-7,5	-6,5	-42	-14,5	-18,5	-28	-14,5
980	-7	-6	-40	-14	-18	-27,5	-14
990	-7	-6	-38	-14	-17,5	-27	-13,5

Table key :

1. Single matched hybrid insertion loss (C-B), port D = 100-ohm

2. Single hybrid insertion loss (C-D), port D = open circuit.

3. Single hybrid isolation (B-D), C = 50-ohm.

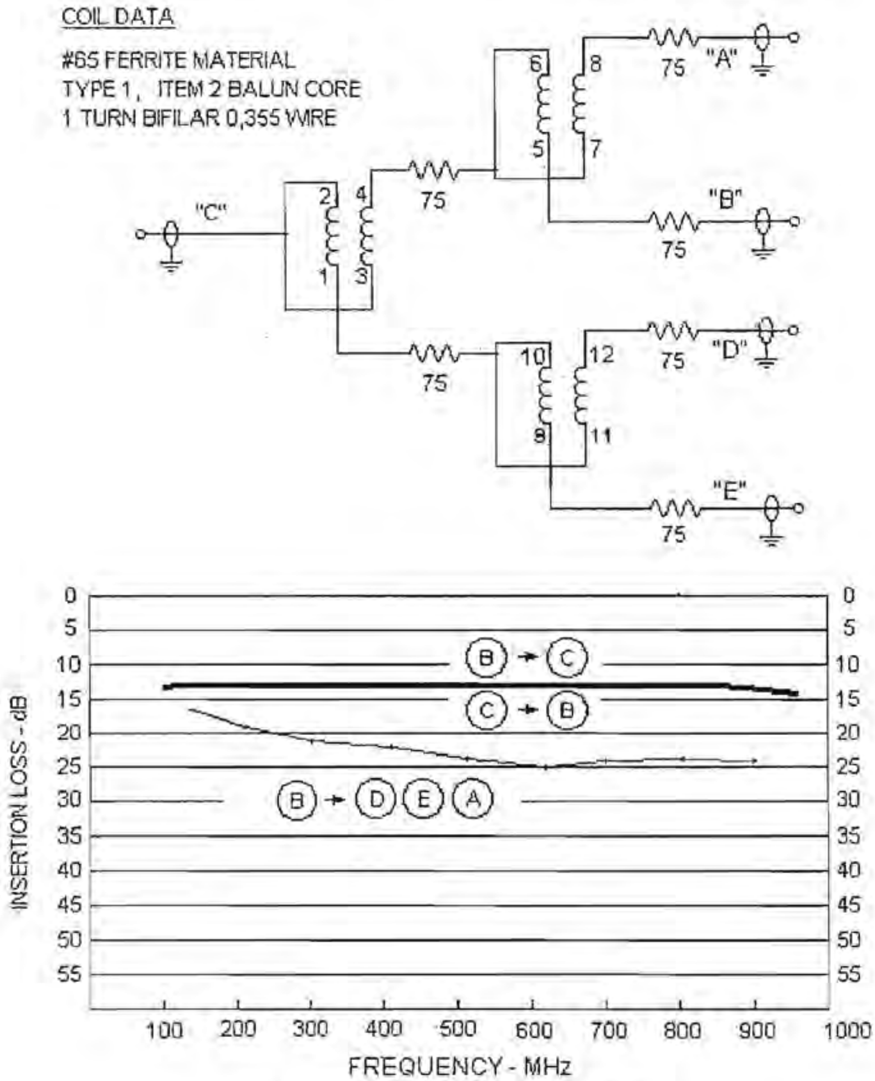
4. Cascaded matched hybrid insertion loss (C-D), D E and A = 200-ohm.

5. Cascaded hybrid insertion loss (C-B), D E and A = open circuit.

6. Cascaded hybrid isolation (B-D), E = 200-ohm, F = 200-ohm, C = 50-ohm.

7. Cascaded hybrid insertion loss, unmatched with no termination or series resistor.

4.2.1.4 Matched 75-ohm cascaded hybrids for signal combining and splitting.



(c) TRANSMISSION CHARACTERISTICS

Figure 4.7 Matched 75-ohm cascaded hybrid

The circuit depicted in Figure 4.7 is matched to 75-ohm characteristic input and output impedances. The disadvantage of inserting 75-ohm matching resistors is an increase in return loss across the circuit, to around 13-dB across the operating frequency band. Isolation characteristics and flatness are however improved when compared with an unmatched hybrid, especially at the high end of the operating band. The circuit may serve as a signal combiner at the front end of the bandwidth switch, as well as a signal divider at the output of the same, in order to obtain multiple forward connections.

4.2.2 Development of the selector electronics

The selector consists of separate RF and IF sections, which will be described in the following subsections:

4.2.2.1 RF section

Figure 4.8 indicates the block diagram of a combined RF section (SANYO part number 4-115V-17400), for operation in band III (175 - 248 MHz) as well as band V (470 - 860 MHz).

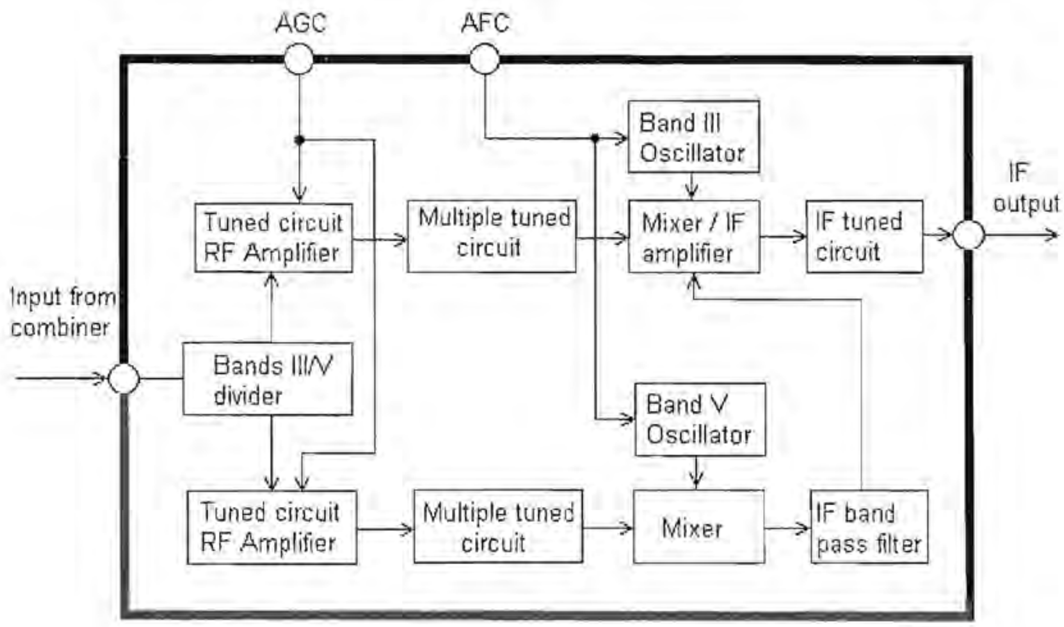


Figure 4.8 Block diagram of RF section

The frequency band divider employs a low pass filter and a high pass filter to divide the frequency bands III and V respectively.

The band III tuner and RF amplifier suppress all signals other than the desired channel signal. The AGC circuit is engaged by the signal strength and the output voltage is uniformly maintained.

The multiple tuned circuit has high selectivity characteristics and suppress all signals other than the desired signal.

The oscillator circuit generates the correct frequency required to convert the desired signal into an intermediate frequency (IF) signal.

The mixer/IF amplifier circuit functions as the band III mixer and band V IF amplifier. When band III signals are received, the output signal of the band III oscillator and the channel being received are mixed and converted into the IF signal. When band V signals are received, the circuit functions as an amplifier of the signal which is converted into the IF signal. The IF tuned circuit ensures that only the IF signal is present at the designated output.

The band V circuitry is a duplication of the tuned circuit/RF amplifier, the multiple tuned circuit, oscillator and the mixer that is also utilised by the band III section.

However, an additional IF bandpass filter with high selectivity characteristics is employed in the band V section to ensure that no unwanted signals are propogated to the mixer/IF amplifier.

4.2.2.2 VIF and SIF circuit

The VIF (vision intermediate frequency) and SIF (sound intermediate frequency) circuit were designed around the TA7607AP and LA1365, both large scale linear integrated circuits indicated as IC1 and IC2 respectively in Figure 4.9.

These circuits serve to obtain the required selectivity response and to amplify and detect the VIF signal and SIF (sound intermediate frequency).

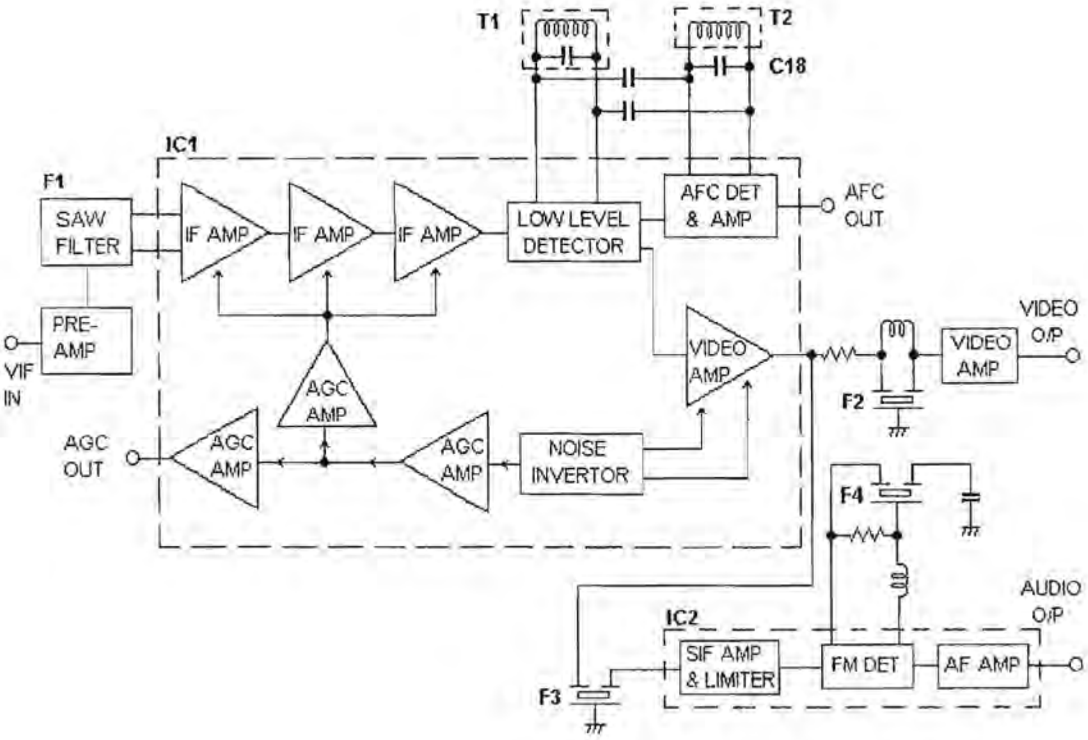


Figure 4.9 Block diagram of VIF and SIF circuits

The selectivity characteristics required are produced by the surface acoustic wave (SAW) filter F1. The SAW filter unfortunately also introduces a loss effect, therefore an amplifier is required to compensate for this loss.

In order to obtain a high signal-to-noise ratio, an emitter-earthed single stage amplifier Q1 is employed before the SAW filter as indicated in Figure 4.10.

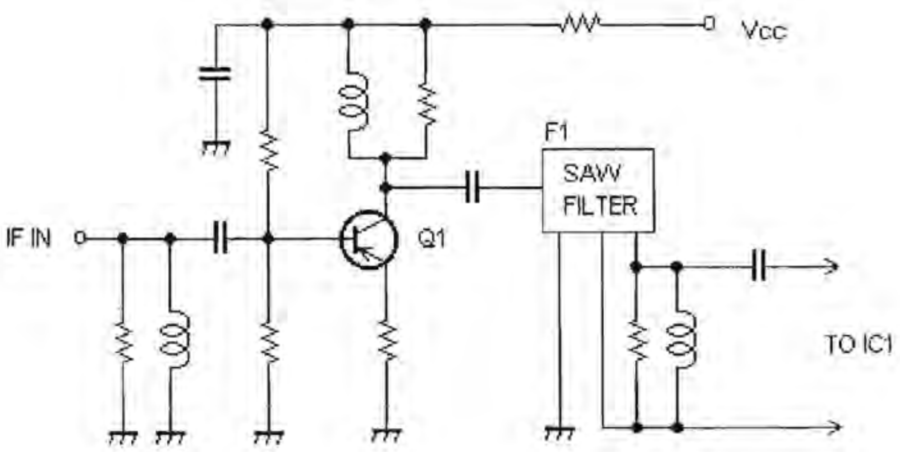


Figure 4.10 Pre-amplifier

The VIF signal enters IC1 from the SAW filter. IC1 contains the VIF amplifier, the video detector, the AFC circuit, as well as the AGC circuit.

The signal is amplified by a 3-stage variable gain IF amplifier. The amplified VIF signal is applied to a tuned circuit contained in T1, and sync detection is performed.

The detected video signal is amplified while its signal-to-noise ratio is simultaneously improved by the noise invertor circuit.

The AFC circuit comprises of T2 and C18, as indicated in Figure 4.11, and includes a phase shifter. The carrier output of the sync detector is applied to the phase shifter and as the phase is detected, a DC voltage proportional in magnitude to the VIF carrier is obtained.

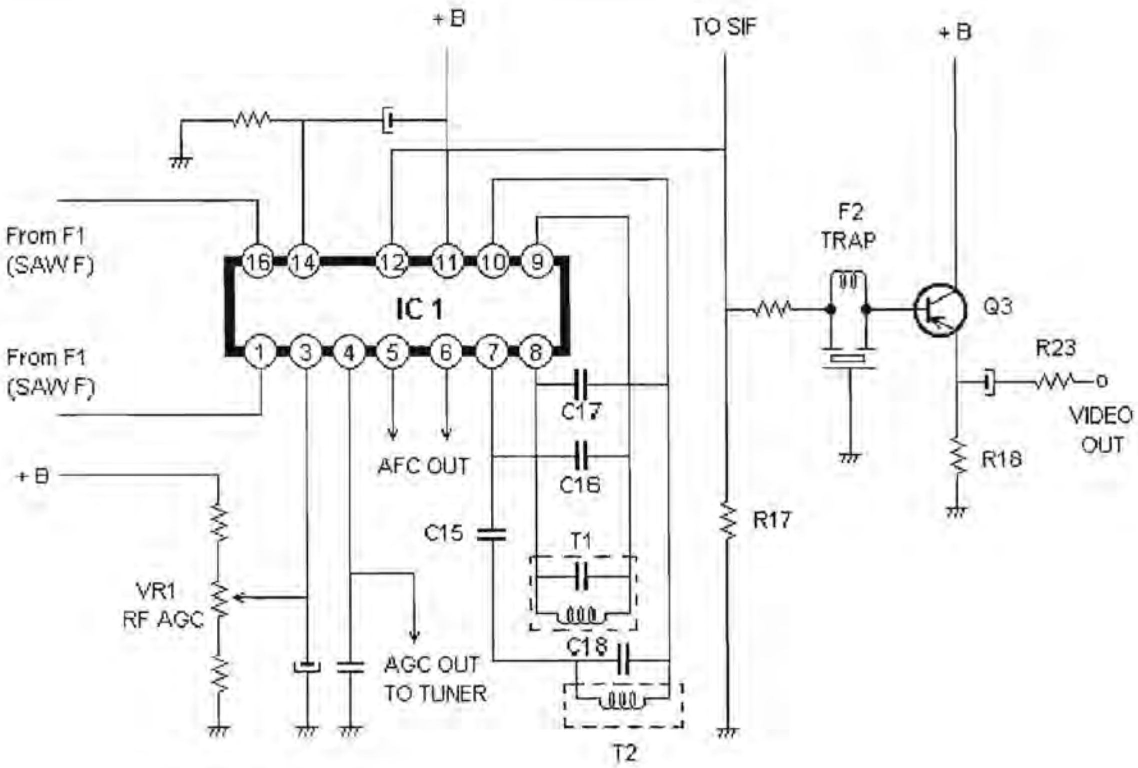


Figure 4.11 AFC and AGC circuits

The AGC circuit is required to maintain the video output at a constant level with respect to fluctuations in the RF input voltage. The control circuit takes the form of a peak AGC system.

The synchronising signal level of the video signal is detected, and the gain of the VIF amplifier circuit is varied accordingly, so that the output level is maintained at a fixed value.

The AGC voltage is supplied to the RF amplifiers contained within the preceding RF stage. When the RF input voltage is low, the RF AGC action ceases and the RF amplifiers function at maximum gain. Once the RF input voltage exceeds approximately 65 dB/uV, VR1 is set so that operation commences.

The level of the video signal detected by IC1 is set by R23 so that a 1V p-p output is obtained at the video output connector (75 ohm termination). The SIF signal contained within the video signal is eliminated by the ceramic trap F2.

Due to the fact that the level of the video signal detected is comparable to the level of the chromanance signal, the latter (4,43 MHz) is attenuated.

Finally, the high ouput impedance is reduced by the emitter follower stage and R18 is applied to set the output impedance to 75 ohms.

The SIF (sound intermediate frequency) signal which is contained within the video signal detected by IC1 is extracted by the ceramic filter F3 and supplied to IC2.

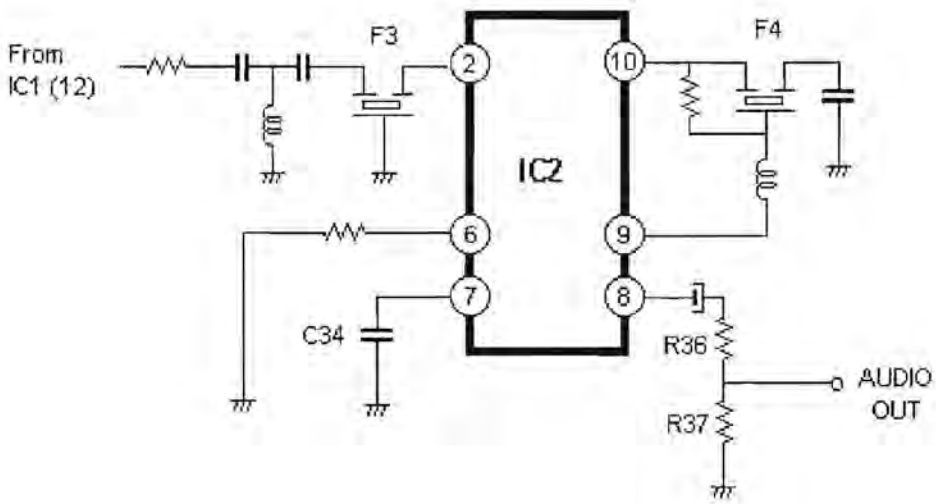


Figure 4.12 SIF circuit

IC2 contains an amplitude limiter circuit and an FM detector circuit. The SIF signal has its AM component removed by the amplitude limiter circuit. The signal is then fed to the differential peak detector circuit, and the sound signal is produced.

The sound signal has undergone pre-emphasis at the transmission end, and is now de-emphasised by C34 combined with the internal impedance of the IC, and the frequency response is smoothed. The audio output level is determined by R36 and R37.

A muting circuit is included for noiseless switching, that may be exploited when considering PC driven manipulation. For this purpose, a voltage forcibly applied to IC1 pin (14) will increase the AGC voltage for the IF amplifier, which will mute the video signal output as well as the SIF signal output.

The RF section utilised for the selection function uses varactor diodes to adjust various tuned circuits required for functional operation. These semiconductor devices are in essence variable tuning capacitors, their values dependend on the tuning voltage applied at any given time.

Tuning is realised by a variable resistor network connected across a 30-VDC power supply rail. (See para 4.2.3). With a desired bandwidth and channel, the wiper of the selector variable resistor is rotated until the channel centre frequency is observed.

Several circuits may be preset to different centre frequencies along the frequency plan. Each circuit is selected by means of a physical switch for manual operation, or reed-relays for manipulation by PC.

An AFC defeat circuit is included to prevent AFC misoperation. When operating AFC mode this circuit momentarily shuts off the AFC during channel-change operation to prevent AFC mis-operation.

The detailed VIF and SIF circuits are indicated in Figures 4.13.a and 4.13.b respectively. Figure 4.13.c depicts the channel pre-set and select circuitry. Labelled terminal blocks represent electrical connections between circuits. The RF section is shown connected to the VIF circuit for operation with AFC and AGC.

The SIF circuit is indicated separately, but forms an integral part of the VIF printed circuit. Plugs are indicated separately but are one and the same for all circuits.

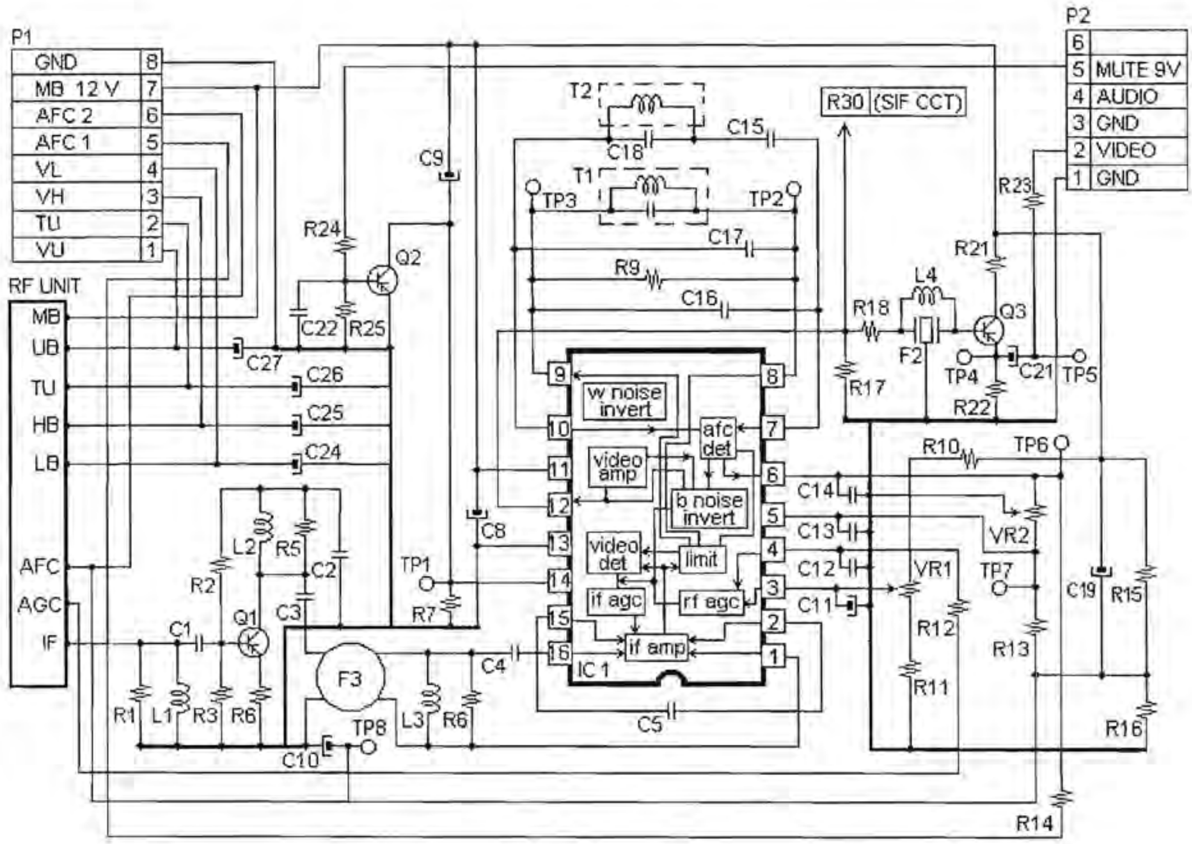


Figure 4.13.a Detailed VIF circuit

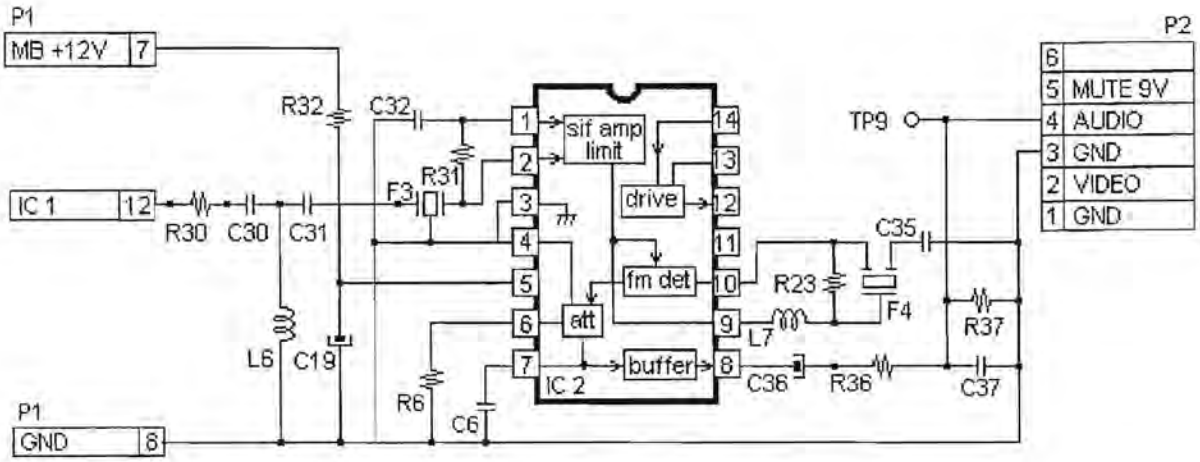


Figure 4.13.b Detailed SIF circuit

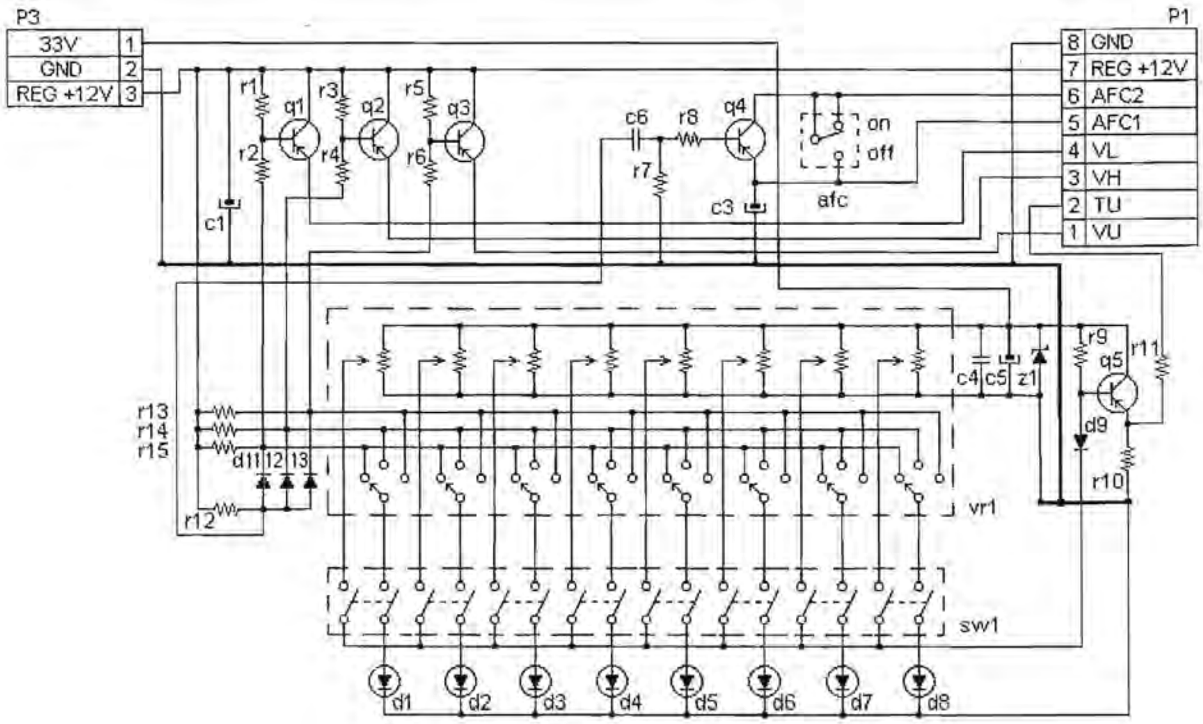


Figure 4.13.c Detailed pre-set/select circuit

4.2.3 **Development of the power supply.**

The RF bandwidth switch was specified to operate from a single external DC power supply voltage of 12-VDC.

Tuning of the RF section of video and television equipment is effected by the varicap method, whereby the voltage is adjusted across a varicap diode to change its capacitance. A mains transformer is normally used to obtain a 33-VAC line for this purpose.

An alternative method had to be devised to generate the regulated 30-VDC tuning rail from the specified operating voltage of 12-VDC. A switch mode power supply module is used to provide the external unregulated 12-VDC to the prototype, and a dual rail internal DC power supply provides unregulated 33-VDC and regulated 12-VDC. The power supply for the prototype bandwidth switch is shown in Figure 4.13.d.

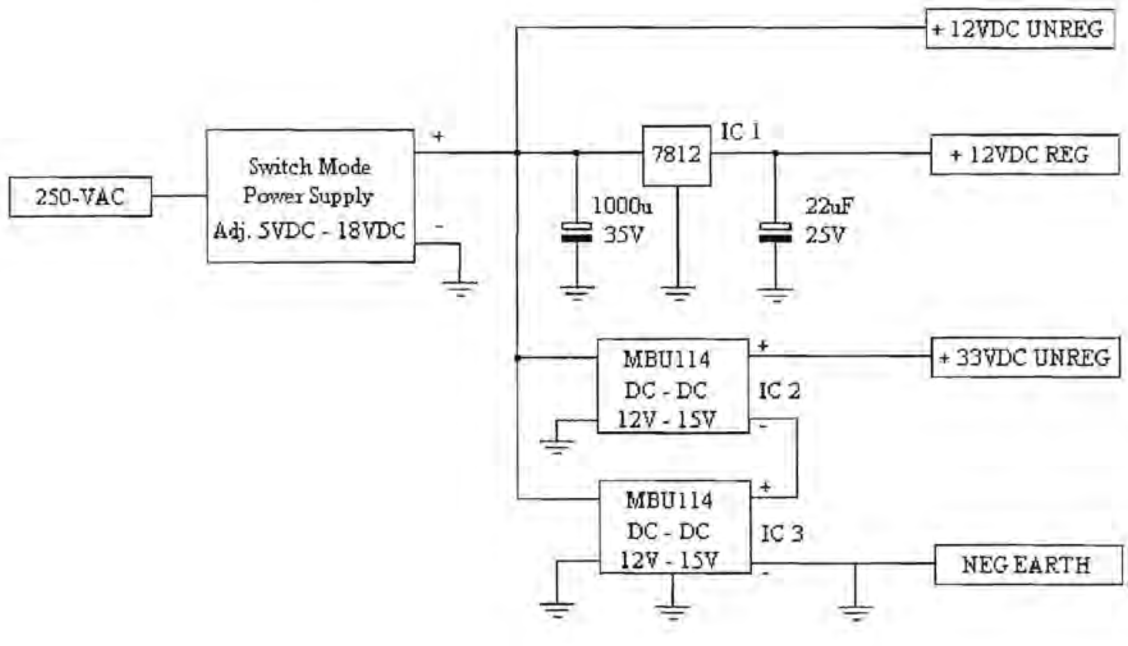


Figure 4.13.d. Dual rail DC power supply for Bandwidth Switch.

The unregulated 33-VDC is applied to a reference diode regulator network as in Figure 4.13.c, consisting of C4 (0,01uF), C5 (22uF/50V) and Z1 (uPC574J). The virtual tuning voltage on the wiper on the selected variable resistor in the voltage dividing network VR1 is applied via D9 to the base of Q5 (2SC536KE). The actual tuning voltage (2-VDC - 28-VDC) TU, is applied via R11 (1K) to pin 2 of P1.

The mounting of the switch mode supply module within the same (metal) enclosure as the bandwidth switch circuitry introduced severe ripple noise in the regulated 12-VDC line. Worst affected was the RF modulator that resulted in a visible noise pattern on the bandwidth switch transmission output.

The regulated 12-VDC line to the RF modulator was screened with copper braiding that was electrically connected to earth within the metal enclosure. This action, together with another wired connection between earth (and earth!) solved all problems experienced with electrical noise introduced from internal, and external sources.

The switch mode power supply module output voltage is adjusted to 13,8-VDC, measured across the input to IC1, the 7812 voltage regulator.

The following table indicates measured values of the power supply output voltages during normal operation of the bandwidth switch:

Parameter	Test point	Adjustment	Value (VDC)
12VDC UNREG	Pin 1 of IC1	External supply	13,8
12VDC REG	Pin 3 of P3	-	11,53
30VDC	Pin 1 of P3	-	29,58

Table 4.8 Dual rail power supply output voltages

4.3 Set-up procedures

The RF and VIF circuitry may be set up for adjustment with either the use of a tracking generator cum spectrum analyser, or by using a sweep- marker generator along with an oscilloscope.

The latter is described, as the financial implications are certainly reduced when considering spectrum analysis equipment.

4.3.1 Pre-test setup

Connect a sweep-marker generator and an oscilloscope as shown in Fig 4.14. The output signal from the sweep-marker generator is applied to the RF unit through an input probe, and the parameter measured is applied to the oscilloscope through an output probe.

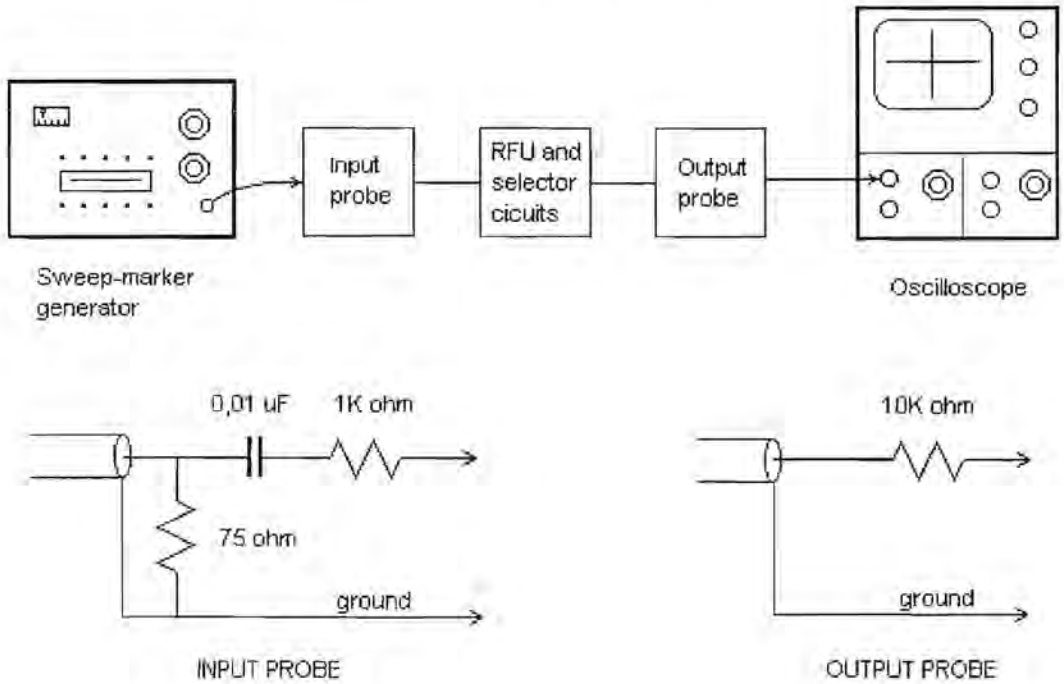


Figure 4.14 Pre-test setup

4.3.2 VIF adjustment

Adjustment location	Measuring point	Measuring equipment	Adjustment condition
Convertor transformer	TP 4 (E of Q3)	Sweep-marker generator. Oscilloscope. DC power supply. Special adj. tool	STOP mode

a. Connect a 22 ohm resistor between test points 3 and 4 (pin 8 and pin 9 of IC 1, as shown in Fig. 4.13.a) and disconnect the P2 connector.

b. Supply DC +0,8V to TP1 (pin 14 of IC 1) as an AGC voltage.

c. Set the output level of the sweep-marker generator to VIF, -25dB (28 mV RMS) and connect the output to the TP terminal of the RF unit through an input probe.

d. Connect the output signal at TP4 (E of Q3) to the oscilloscope through an output probe, and observe the waveform.

e. Adjust the AGC voltage so that the waveform level is 0,8Vp-p.

f. Adjust the RF unit's convertor transformer so that P (the 39,5 MHz marker) becomes 38±2%. See Figure 4.15.a

g. Remove the 22 ohm resistor and reconnect the P2 connector.

4.3.3 Detection transformer adjustment

Adjustment location	Measuring point	Measuring equipment	Adjustment condition
T1	TP4 (E of Q3)	Sweep-marker gen. Oscilloscope DC power supply Special adj. tool	STOP mode

a. Supply DC +2V to TP1 (pin 14 of IC 1, as shown in Fig. 4.13.a) as an AGC voltage, and remove the P2 connector.

b. Set the output level of the sweep-marker generator to VIF, -25dB (28mV RMS) and connect the output to the TP terminal of the RF unit through an input probe.

c. Connect the output signal at TP4 (E of Q3) to the oscilloscope through an output probe, and observe the waveform.

d. Adjust the AGC voltage so that the waveform level is equal to 2,5Vp-p.

e. Adjust the T1 core so that P (the 39,5 MHz marker) becomes 70%. See Figure 4.15.b.

f. Disconnect the P2 connector.

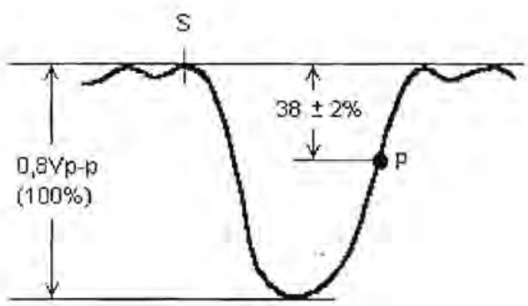


Figure 4.15.a. Wave-form example

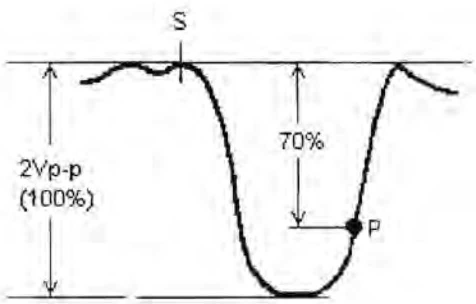


Figure 4.15.b. Wave-form example

4.3.4 AFC circuit (1)

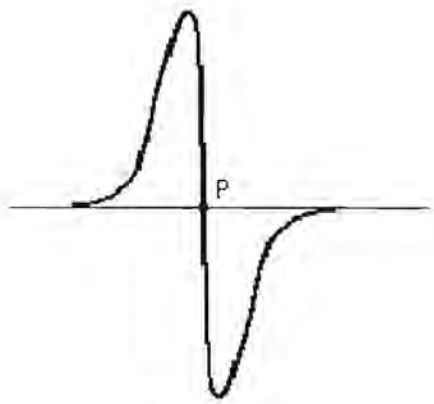
Adjustment location	Measuring point	Measuring equipment	Adjustment condition
T2	Pin 6 of P1	Sweep-marker gen. Oscilloscope DC power supply Special adj. Tool	STOP mode AFC switch "ON"

a. Supply DC +3,5V to TP1 (pin14 of IC 1) as an AGC voltage.

b. Set the output level of the sweep-marker generator to VIF, -13dB (112mV RMS), and connect the output to the TP terminal of the RF unit through an input probe.

c. Connect the output signal at pin 6 of P1 on the circuit board to the oscilloscope through an output probe, and observe thewaveform.

d. Adjust T2 so that P (the 39,5 MHz marker) is positioned on the base line, as in the sketch below.



4.3.5 AFC circuit (2)

Adjustment location	Measuring point	Measuring equipment	Adjustment condition
VR2	TP6 (pin 6 of IC 1) TP7 (pin 5 of IC 1)	DC power supply DC voltmeter	STOP mode AFC switch "ON"
a. Eliminate all input signals. b. Apply DC +5V~ +8V to TP1 (pin 14 of IC 1). c. Adjust VR2 so that the voltage between TP6 and TP7 is 0 ~ 0,5V.			

4.3.6 RF AGC circuit

Adjustment location	Measuring point	Measuring equipment	Adjustment condition
VR1	RF unit AGC terminal	Colour bar generator with RF section. Digital volt-meter. Video monitor (composite video).	STOP mode
a. Connect the colour bar generator to the input of the RF unit. (VHF input, channel 5 or UHF input, channel 31, 69dB/uV, modulation degree 87,5%). b. Connect the video output (pin 2 of P2) to the video monitor composite video input. c. Press channel 1 on the channel selector, and adjust to channel 5 or channel 31, setting for the best possible picture quality observed on video monitor. d. Set the AFC switch to the "ON" position. e. Measure the voltage at the AGC terminal of the RF unit, and adjust VR1 so that the voltage observed is $5{,}7 \pm 0{,}1$VDC for VHF, or $5{,}0 \pm 0{,}1$VDC for UHF.			

4.3.7 Audio output

<table>
<tr><th>Adjustment location</th><th>Measuring point</th><th>Measuring equipment</th><th>Adjustment condition</th></tr>
<tr><td>-</td><td>TP9 (pin 4 of P2)</td><td>Television signal gen.
Oscilloscope.</td><td>STOP mode</td></tr>
<tr><td colspan="4">a. Connect the TV signal generator to the input connector of the RF unit. (VHF input, channel 5, 60dB/uV (1mV RMS) or higher, 60% modulation with sine wave audio signal at 400 Hz).

b. Press a channel, select VHF and adjust for best possible picture quality observed.

c. With AFC "ON", observe thewaveform of thesinewave audio at TP9 on the oscilloscope, and confirm the signal level at 0,69± 0,34V p-p.</td></tr>
</table>

4.4 Test procedures : Internal noise measurement

Before proceeding with the noise measurements on the RF bandwidth switch, we shall have to discuss noise created by any of the passive devices found in receivers. Such noise is generally random, and is thus impossible to treat on an individual voltage basis, but easy to describe statistically since it is truly random.

Since the noise is randomly distributed over the entire radio spectrum there is on average, as much of it at any frequency as at any other, and it may therefore be assumed that random noise power is proportional to the bandwidth over which it is measured.

4.4.1 Thermal agitation noise

This noise is generated in a resistance or the resistive component of any impedance, is random, and is referred to as *thermal, agitation, white,* or *Johnson* noise (after its discoverer) [22]. It is due to the rapid and random motion of the molecules, atoms and electrons of which, according to simplified atomic theory, any such resistor is constructed.

In thermodynamics, kinetic energy shows that the temperature of a particle is a way of expressing its internal kinetic energy, so that the 'temperature' of a body is the statistical RMS value of the velocity of motion of the particles in the body. The theory states that the kinetic energy of these particles becomes approximately zero, at $0\ K$ (Kelvin or absolute), which very nearly equals -273°C. It is thus apparent that the noise power generated by a resistor is proportional to its absolute temperature, in addition to being proportional to the bandwidth over which the noise is to be measured. Thus:

$$P_n \propto T\ \delta f = kT\ \delta f \quad \ldots\ldots\ldots\ldots\ldots\ldots (2)$$

where k = Boltzmann's constant = 1,38 x 10^{-23} J/K (joule/kelvin)

T = absolute temperature, K (kelvin) = 273 + °C

δf = bandwidth of interest

P_n = maximum noise power output of a resistor

If an ordinary resistor at the standard temperature of 17°C (290 K) is not connected to any voltage source, it might at first be thought that there is obviously no voltage to be measured across it. That is correct if the measuring instrument is a DC voltmeter, but it is decidedly incorrect if a very sensitive electronic voltmeter is considered; there may even be quite a large voltage across the resistor, but since it is random and therefore has a definite RMS value but no DC component, only an AC instrument will register a reading.

This noise voltage is caused by the random movement of electrons within the resistor, which constitutes a current; although as many electrons arrive at one end of the resistor as at the other over any long period of time, at any instant of time there are bound to be more electrons arriving at one particular end than at the other because their movement is random.

Equally, over a period of time, this imbalance will be redressed, but as the rate of arrival of electrons at either end of the resistor varies randomly, so does the potential difference between the two ends, thus a random voltage across the resistor definitely exists and may both be measured and calculated.

It must be realised that all formulae referring to random noise are applicable only to the RMS value of such noise, and not to its instantaneous value, which is quite unpredictable. So far as peak noise voltages are concerned, there is reason to believe that they are unlikely to have values in excess of 10 times the RMS value.

From Equation (1), the equivalent circuit of a resistor as a noise generator may be drawn as in Figure 4.16, and from this the resistor's equivalent noise voltage E_n may be calculated.

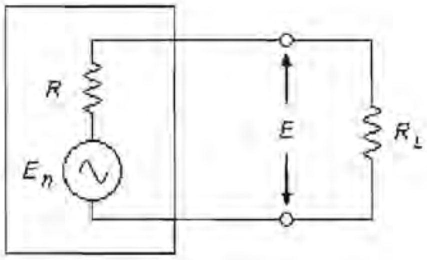


Figure 4.16. Resistance noise generator

Consider that R_L is noiseless and is receiving the maximum noise power generated by R; under these conditions of maximum power transfer, R_L must be equal to R. Then:

$$P_n = \frac{E^2}{R_L} = \frac{E^2}{R} = \frac{(E_n/2)^2}{R} = \frac{E_n^{\,2}}{4R} \quad \ldots\ldots (3)$$

thus $$E_n^{\,2} = 4RP_n = 4RkT\ \delta f \quad \ldots\ldots (4)$$

and $$E_n = \sqrt{4kT\,\delta f\,R} \quad \ldots\ldots (5)$$

It is seen from Eq. (5) that the square of the noise voltage E_n associated with a resistor is proportional to the absolute temperature of the resistor, the value of the resistance, and the bandwidth over which the noise is measured. Note especially that generated noise voltage is quite independent of the frequency at which it is measured; this stems from the fact that it is random and evenly distributed over the frequency range considered.

4.4.2. Noise figure

Measurement of the signal-to-noise ratio (*S/N*) of an amplifier, a receiver or a device is often used for either of two purposes, or sometimes for both; the comparison of two circuits for evaluation of their performance or the comparison of noise and signal at the same point to ensure that the former is not excessive.

Signal-to-noise ratio *S/N* is defined as the ratio of signal *power* to noise *power* , at the same point.

Thus, $$\frac{S}{N} = \frac{P_s}{P_n} = \frac{E_s^{\,2}/R}{E_n^{\,2}/R} = \left(\frac{E_s}{E_n}\right)^2 \quad \ldots\ldots (6)$$

Equation (3) above is a simplification which applies whenever the resistance across which the noise is developed is the same as the resistance across which the signal is developed, and this is almost invariably the case. An effort is naturally made to keep the signal-to-noise ratio as high as possible under a given set of conditions.

For comparison of receivers or amplifiers working at different impedance levels the use of *noise figure*, or sometimes known as *noise factor*, is defined and used. The noise figure F is defined as the ratio of the signal-to-noise power supplied to the input terminals of a receiver or amplifier to the signal-to-noise power supplied to the output or load resistor.

Thus, $$F = \frac{Input\ S/N}{Output\ S/N} \qquad (7)$$

It can be seen immediately that the noise figure is 1 for an ideal receiver or amplifier or device which introduces no noise of its own, so that the signal-to-noise ratio does not deteriorate as a result thereof.

Also known is the alternative definition of noise figure, which states that F is equal to the S/N of an ideal system divided by the S/N of the receiver or amplifier under test, both working at the same temperature over the same bandwidth and fed from the same source. In addition, both systems must be linear.

The noise figure may be expressed as an actual ratio, or in decibels. The noise figure of practical receivers can be kept to below a few decibels up to frequencies in the order of 1-GHz by a suitable choice of the first MMIC or transistor or tube, combined with proper circuit design and the use of low-noise resistors.

4.4.3 Calculation of noise figure

Noise figure may be calculated for an amplifier or receiver or device on the same basis by treating either as a unit; that is, each may be treated as a four-terminal network having an input resistance R_t, an output resistance R_L, and an overall gain A. It is fed from a source (antenna or generator) of internal resistance R_a, which may or may not be equal to R_t as the circumstances vary. A block diagram of such a four terminal network is shown in Figure 4.17.

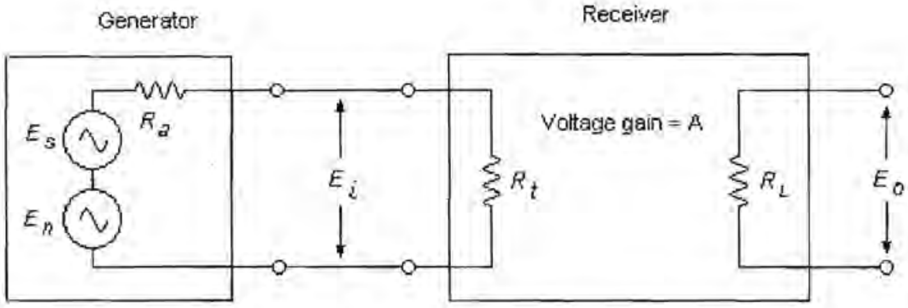


Figure 4.17. Block diagram for noise figure calculation

The calculation procedure may be broken down into the following steps, each followed by the number of relevant equation(s):

1. Determine the signal input power (8, 9).
2. Determine the noise input power (10, 11).
3. Calculate the input signal-to-noise S/N_i, from the ratio of P_{si} and P_{ni} (12).
4. Determine the signal output power P_{so} (13).
5. Write P_{no} for the noise output power to be determined later (14).
6. Calculate the output signal-to-noise ratio S/N_o from the ratio of P_{so} and P_{no} (15).
7. Calculate the generalised form of noise figure from steps 3 and 6 (17)
8. Calculate P_{no} from R_{eq} if possible, and substitute into the general equation for F to obtain the actual formula, or determine P_{no} from measurement and substitute in Eq. (17) to obtain the formula for F.

It is seen from Fig. 4.16 that the signal input voltage will be:

$$E_{si} = \frac{E_s R_t}{R_a + R_t} \quad \text{. . . . . (8)}$$

$$P_{si} = \frac{E_{si}^2}{R_t} = \left(\frac{E_s R_t}{R_a + R_t}\right)^2 \frac{1}{R_t} = \frac{E_s^2 R_t}{\left(R_a + R_t\right)^2} \quad \text{. . . . . (9)}$$

Similarly, the noise input voltage will be:

$$E_{ni}^2 = 4kT\delta f \frac{R_a R_t}{R_a + R_t} \quad \text{. . . . . (10)}$$

$$P_{ni} = \frac{E_{ni}^2}{R_t} = 4kT\delta f \frac{R_a R_t}{R_a + R_t}\frac{1}{R_t} = \frac{4kT\,\delta f\,R_a}{R_a + R_t} \quad \text{. . . . . (11)}$$

The input signal-to-noise ratio will be:

$$\frac{S}{N_i} = \frac{P_{si}}{P_{ni}} = \frac{E_s^2 R_t}{\left(R_a + R_t\right)^2} \div \frac{4kT\ \delta f\ R_a}{R_a + R_t} = \frac{E_s^2 R_t}{4kT\ \delta f\ R_a\left(R_a + R_t\right)} \quad \text{. . . . . (12)}$$

The output signal power will be

$$P_{so} = \frac{E_{so}^2}{R_L} = \frac{\left(AE_{si}\right)^2}{R_L} = \left(\frac{AE_s R_t}{R_a + R_t}\right)^2 \frac{1}{R_L} = \frac{A^2 E_s^2 R_t^2}{\left(R_a + R_t\right)^2 R_L} \quad \text{. . . . . . (13)}$$

The noise output power may be difficult to calculate; for the time being, it may simply be written as:

$$P_{no} = \text{noise output power} \quad \ldots\ldots \quad (14)$$

Thus the output signal-to-noise ratio will be

$$\frac{S}{N_o} = \frac{P_{so}}{P_{no}} = \frac{A^2 E_s{}^2 R_t{}^2}{\left(R_a + R_t\right)^2 R_L P_{no}} \quad \ldots\ldots \quad (15)$$

Finally, the general expression for noise figure is

$$F = \frac{S / N_i}{S / N_o} = \frac{E_s{}^2 R_t}{4kT\ \delta f\ R_a\left(R_a + R_t\right)} \div \frac{A^2 E_s{}^2 R_t{}^2}{\left(R_a + R_t\right)^2 R_L P_{no}} \quad \ldots\ldots \quad (16)$$

$$= \frac{R_L P_{no}\left(R_a + R_t\right)}{4kT\ \delta f\ A^2 R_a R_t} \quad \ldots\ldots \quad (17)$$

Note that Eq. (17) is an intermediate result from which an actual formula for F may be obtained by substitution for the output noise power, from a knowledge of the equivalent noise resistance or from actual measurement.

4.4.4 Measurement of noise figure

The above sections 4.4.1 to 4.4.3 were included to demonstrate that accurate noise measurement remains not an easy task to perform. In an attempt to determine the noise figure of the bandwidth switch, it was decided to obtain sophisticated measuring gear, and to follow the route of measuring the input signal power and noise power, as well as the output signal power and noise power, as opposed to using the equivalent noise resistance. This option may serve to measure not only noise that is generated from within the receiver, but also the noise due to external influences.

A VCR may be regarded as a typical input to the bandwidth switch, and the same was thus used to perform the measurement. The input signal power from the VCR modulator to the RF bandwidth switch measured -26,8-dBm, and the noise power was estimated at -69,2-dBm for a 6-MHz span width, resulting in an input *S/N* of 42,4-dB.

The output signal power of the modulator used in the bandwidth switch measured -24,9-dBm, and the output noise power was estimated at -64,6-dBm, resulting in an output *S/N* of 39,7-dB. It can be seen that although the output modulator has a signal output power 1,9-dB higher than the input, the RF bandwidth switch has added noise to the signal, with a resulting noise figure F of 2,7-dB. The bulk of the noise is most probably generated within the mixer stages of the RF tuner unit, and perhaps a lesser part within the RF modulator.

Signal power was measured at a relatively narrow resolution bandwidth of 30-kHz, otherwise noise generated from external sources may affect the measurement. To estimate the noise power for a 6-MHz span width, the noise power was measured for a resolution bandwidth of 100-kHz, after employing the video averaging feature of the HP8593E spectrum analyser. The result was multiplied by 60 to estimate the noise power for the 6-MHz span of interest.

The expected output *S/N* of 40-dB was not obtained, although the actual figure of 39,7-dB is regarded as acceptable for the purpose of this dissertation. All the other expected electrical inputs and outputs as set in paragraph 4.1.2 were measured to fall within the expected specifications.

4.4.5 Accuracy of the noise measurements

The measurement gear used to perform the noise measurements was the Hewlett Packard HP 8593E Spectrum Analyser with HP 85714A Scalar Measurement Personality option. When resolving signals of equal or different amplitudes in terms of their power levels, one must consider the characteristics of the measuring gear before comments may be made about the accuracy of the measurement. Annexure B considers factors which may affect the accuracy of the noise measurements.

4.4.6 Informal transmission testing

The VCR used for the test was firstly connected directly to a television receiver, and a standard video cassette was viewed by several students at Irene weather station. Transmission was terminated and the VCR output was presented to the input of the prototype RF bandwidth switch, and the output reconnected to the same television receiver and again viewed by the same students. Picture and sound quality was reported to be non-compromised. Following the single input test, four generators were connected to the input of the RF bandwidth switch, and the signals resolved at the output of the combiner. The four generator frequencies were evenly spaced at two channel intervals, situated at channel 30, 32, 34 and 36 respectively. See Figure 4.18 below:

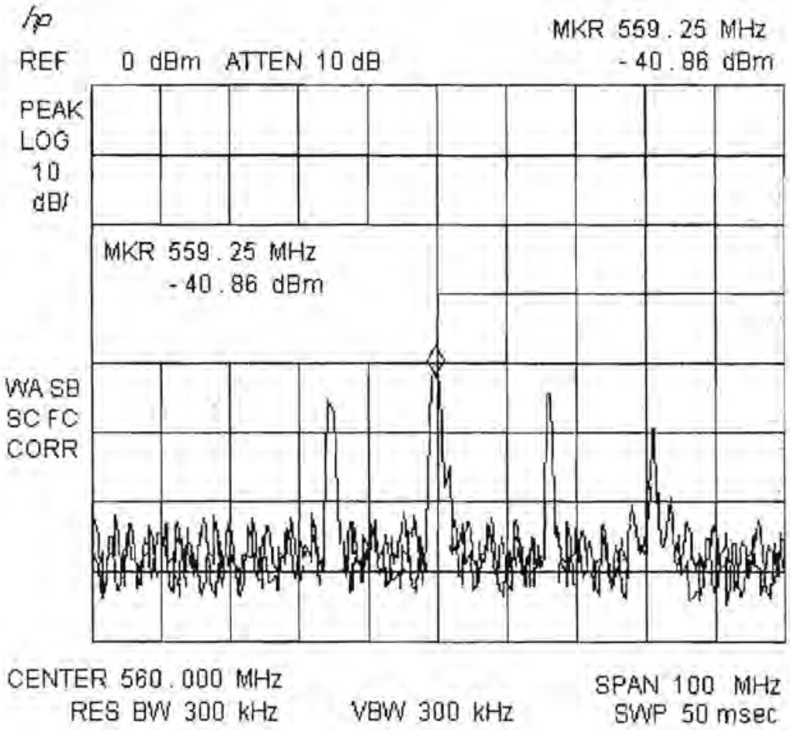


Figure 4.18. Input channel spacing

Next, the preset output of the bandwidth switch modulator was observed, also at channel 36. Attempts to set the output to channel 38 were fruitless, so it was reset to channel 28, to minimise interference with the fourth RF input. The inputs were switched to the output sequentially, and the output observed on a television receiver. Picture and sound quality were reported as acceptable, with no indication of interference between input signals.