CHAPTER 3

The Lossless Sequence Asymmetric Polyphase Filter

Ph.D. Thesis "The Synthesis and Application of Polyphase Filters with Sequence Asymmetric Properties"
Michael John Gingell 1975 University of London Faculty of Engineering.

3.1 Introduction

Networks that contain only reactive elements are called 'lossless' because no real power can be dissipated within them. In order that useful filtering characteristics may be obtained the network is usually fed from and terminated by resistors although these are not included in the determination of whether or not the network is lossless. The true criterion of a lossless network is therefore that all the power (and no more) injected into the input of the network is delivered to the load. In the case of a filter in the stopband where very little power reaches the load this can be considered as being because only that amount of power entered the network in the first place and not because it was lost or dissipated on the way. This is the fundamental reason why it is possible to synthesise a lossless network directly from the transfer function.

The basic elements normally permitted in a lossless network are the coil and the capacitor. They are allowed because although they can store energy they cannot dissipate it. Storage is an essential part of the functioning of a filter and only extra stores can raise the order of the transfer function.

Certain devices may be included in a lossless network but they do not add to the total number of useful energy stores and cannot therefore in themselves increase the order of the filter function.

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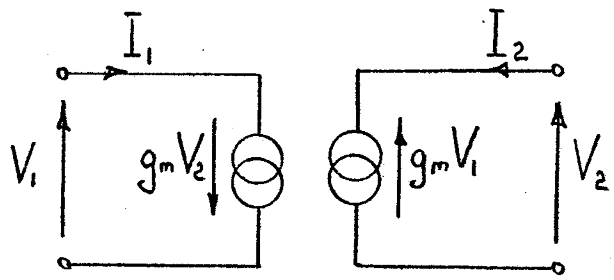
One such device is the gyrator which is notably not only for its circuit properties but for the fact that (except at microwave frequences) no completely passive gyrator has yet been reported. Viewed as a 'black box' the gyrator appears rather like a transformer in that the power entering one port is the same as the power leaving the other port. The fact that, in a practical circuit, it may not be the same power can be regarded as a limitation of current technology and not as a justification of the idea that gyrators are not lossless.

Turning to the constant reactance - what properties are required of it? From a circuit point of view, if a sinusoidal voltage is applied across it then a cosinusoidal current will flow and this means that no real power will be dissipated. It is not required to store energy which follows from arguing that a frequency transformation (such as the one mentioned first in section 2.1) does not increase the order of the transfer function. The constant reactances so introduced seem to act merely as circuit transformers. Lastly the constant reactance is really fictitious as an element in its own right since it implies that one could measure the difference between positive and negative frequencies. 'Positive' and 'negative' frequencies are themselves only mathematical devices which normally have no significance unless for example a polyphase network is considered. It seems reasonable, therefore, that if a constant reactance can be realised at all it will be in a polyphase network. Even then it may not be recognizeable as a discrete device but rather embedded in a polyphase device which exhibits the same properties in any phase.

A number of different solutions have been found to the problem and all use controlled sources in some form or other. A controlled source is defined here as a voltage or current generator whose output is linearly proportional to the voltage at some other node or the current in some other branch of the network in which the source is connected. A gyrator is a good example of a device that can be made using only two controlled sources (Fig. 3.1.1).

Two classes of active realisation have been studied in some depth and are described in succeeding sections.



$$\begin{bmatrix} I_1 \\ I_2 \end{bmatrix} = \begin{bmatrix} 0 & g_m \\ -g_m & 0 \end{bmatrix} \begin{bmatrix} V_1 \\ V_2 \end{bmatrix}$$

Gyrator using controlled sources. Fig. 3.1.1.

Although both methods use controlled sources in the active devices the approach is quite different. The first method is to replace each polyphase set of constant reactances by a polyphase 'gyrator'. This generally requires as many gyrators as there are constant reactances. The second method uses complex impedance transformers and the technique relies on producing an initial filter design with a minimum of capacitors possibly at the expense of extra constant reactances. The practical polyphase network is then

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achieved by placing impedance transformers at the input and output of the filter and between the filter and every capacitor. The constant reactances are then transformed into resistors. Provided the optimum filter structure has been chosen in the first place the number of reactances is not too important since each is replaced by a resistor.

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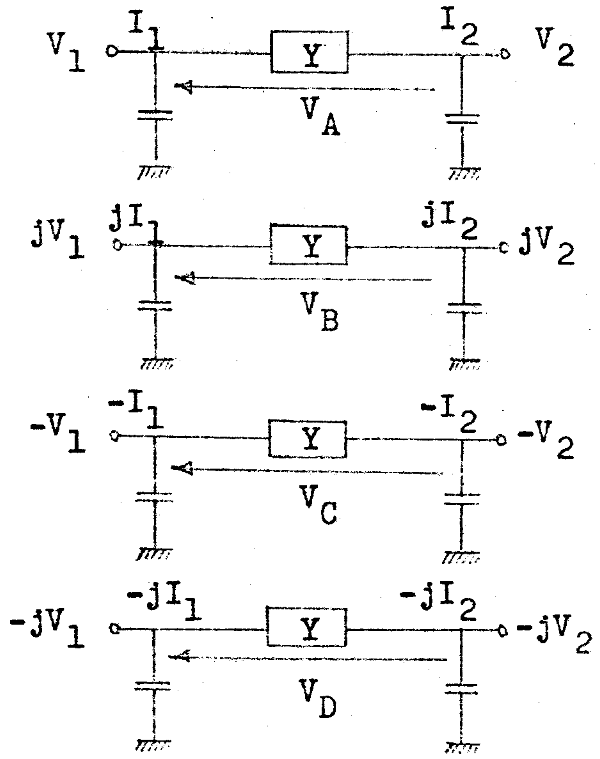
It should be noted that networks containing inductors have not generally been considered because of their disadvantages (against capacitors) of size, cost and the relative difficulty of integration using modern film techniques.

Most of the study that follows is restricted to two and four phase networks because of the fact that in these networks signals in adjacent phases are in quadrature and under these circumstances the simplest circuitry for realising constant reactances is achieved. The two phase case is a special version of the four phase filter where the two phases are in quadrature.

Filters with other numbers of phases can be realised using modifications of the techniques to be described. Some designs for three phase filters are given by way of example.

3.2 Lossless Polyphase Filters Using Gyrators

Consider a four phase filter section containing constant reactances Y driven by a symmetrical four phase input signal and assume it to be symmetrically loaded (Figure 3.2.1 -a). The filter is made up of identical circuits in each phase. If at any point in one of the phases there exists a voltage V and current I



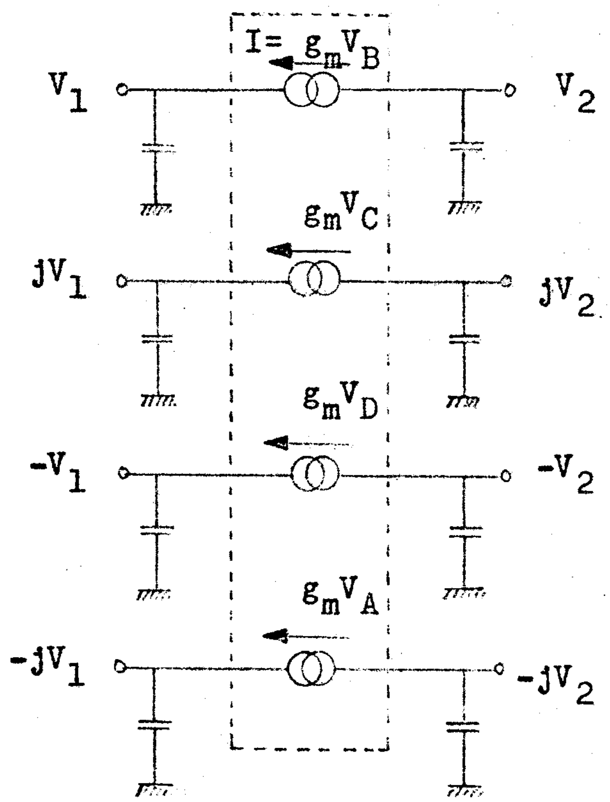
$V_A = V_1 - V_2$

$V_B = jV_1 - jV_2 = jV_A$

$V_C = -V_1 - -V_2 = -V_A$

$V_D = -jV_1 + jV_2 = -jV_A$

A) Four Phase Filter Section.



$Y = jg_m$

B) Realization with a 4 port Gyrator.

Figure 3.2.1 Four Phase Sequence Asymmetric Filter Section using Gyrators.

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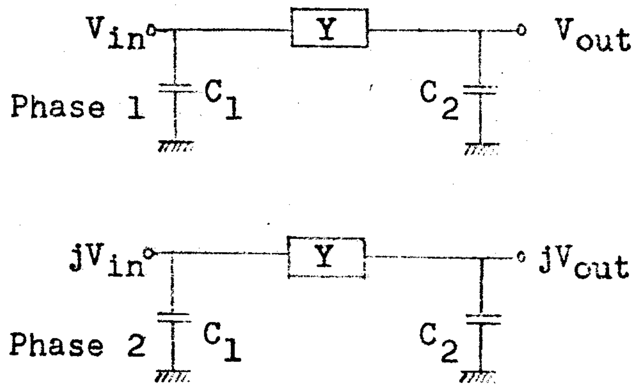
then because the other phases are physically identical at the corresponding points in those phases there will exist jV and jI, -V and -I, -jV and -jI respectively. If controlled sources are connected as in Figure 3.2.1b in place of the constant reactances it can be seen that the impedance looking into each source is imaginary and equal to jg_m where g_m is the mutual conductance of each source.

The set of four controlled sources make up a kind of four port gyrator. When constrained to operate in a four phase mode this device neither increases nor reduces the net power in the circuit. This can be proved by taking the sum of the products of voltage and current at each port of the device which will be found to be zero.

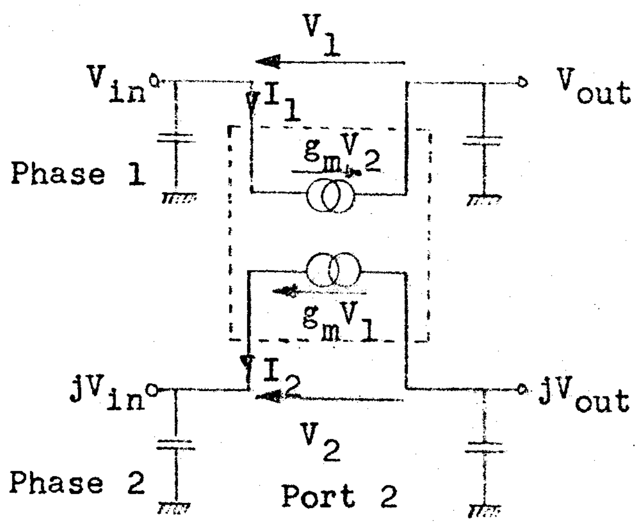
In practice two phase quadrature filters are just as easy to make and they use only half the number of components of the four phase version. Since the -V and -jV phases give no extra phasing information than is provided by their V and jV counterparts they can be eliminated as shown in Figure 3.2.2. Here it can be clearly seen that the two controlled sources which replace the constant reactances form a standard gyrator. A gyrator has the chain matrix

$$\begin{bmatrix} V_1 \\ I_1 \end{bmatrix} = \begin{bmatrix} 0 & \frac{1}{g_m} \\ g_m & 0 \end{bmatrix} \begin{bmatrix} V_2 \\ I_2 \end{bmatrix}$$

ie. $I_1 = g_m V_2$ $\qquad I_2 = g_m V_1$



A) Two Phase (Quadrature) Filter Section.



B) Realization with a Gyrator.

Figure 3.2.2 Two Phase (Quadrature) Sequence Asymmetric Filter Section

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Thus the impedance seen looking into port 1 is

$$Z_1 = \frac{V_1}{I_1} = \frac{V_1}{g_m V_2} = \frac{V_1}{g_m . jV_1} = \frac{1}{jg_m}$$

Similarly at port 2

$$Z_2 = \frac{V_2}{-I_2} = \frac{jV_1}{-g_m V_1} = \frac{1}{jg_m}$$

Therefore looking into either port a reactance of $\frac{1}{jg_m}$ is seen.

3.2.1 An Experimental Filter Section using Gyrators

Before proceeding to build a full scale filter it was thought wise to try a single section to prove the principle worked in practice and to find out what problems might arise. An image designed section was chosen because it was amenable to a number of practical circuit simplifications and the final design did not require floating balanced gyrators.

Figure 3.2.3 shows the basic image design principles of the single phase network. In normalised form the section is a bandpass filter with cut off frequencies at ω=0 and ω=1. The image loss has a peak of attenuation at a finite negative frequency and the image impedance is one ohm at midband rising to infinity at the cut off frequencies. In practice the insertion loss can be expected to differ from the image loss due to reflection and interaction effects between the network and the terminating impedances. Figure 3.2.4 shows the gyrator equivalent design for a two phase quadrature filter. This image section has the particular advantage that the con-

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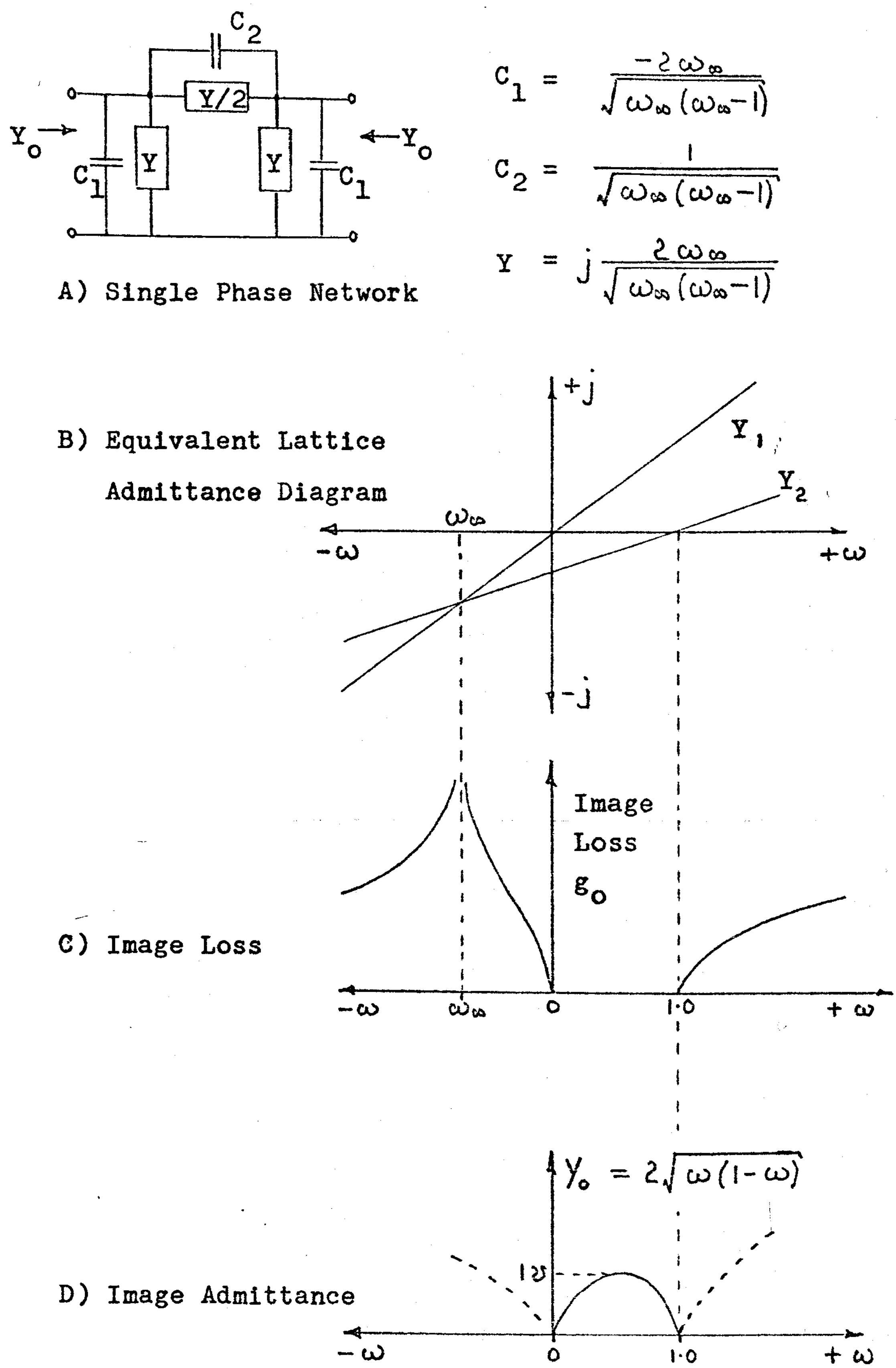

Figure 3.2.3 Image Design of Single Section for first experimental model.

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stant admittance in the top branch is exactly half the value of the constant admittances in each of the shunt branches. By re-arranging the controlled sources as shown in Figure 3.2.4 -B the circuit can be simplified so that only two unbalanced voltage measuring points are required to make the complete filter section.

Figure 3.2.5 shows the final practical circuit. For this the filter was scaled so that the cut off frequencies were zero and 3.8kHz and the image impedance was 7.7kΩ at mid band. By mismatching the terminations and making them 10kΩ a reasonable passband performance was predicted over the desired range from 0.25 to 3.4kHz.

Transistors Q_5, Q_6, Q_7 and Q_8 act as the controlled sources the g_m being controlled almost entirely by the emitter resistors. This is true providing the transistor gains are sufficiently high (greater than 100). These transistors are controlled by voltage sensing amplifiers. The non inverting amplifier comprises Q_1 and Q_3 and drives sources Q_6 and Q_8. The inverting amplifier comprises Q_2 and Q_4 and drives sources Q_5 and Q_7. The amplifier gain is given by

$$\mu = \frac{g_m . R_2}{1+g_m R_2}$$

where g_m is the mutual conductance of the input transistor which is an N-channel MOS type. There is a slight degradation in gain due to the finite h_{fe} of the second transistor in the pair but this is a second order effect. In this design R_2

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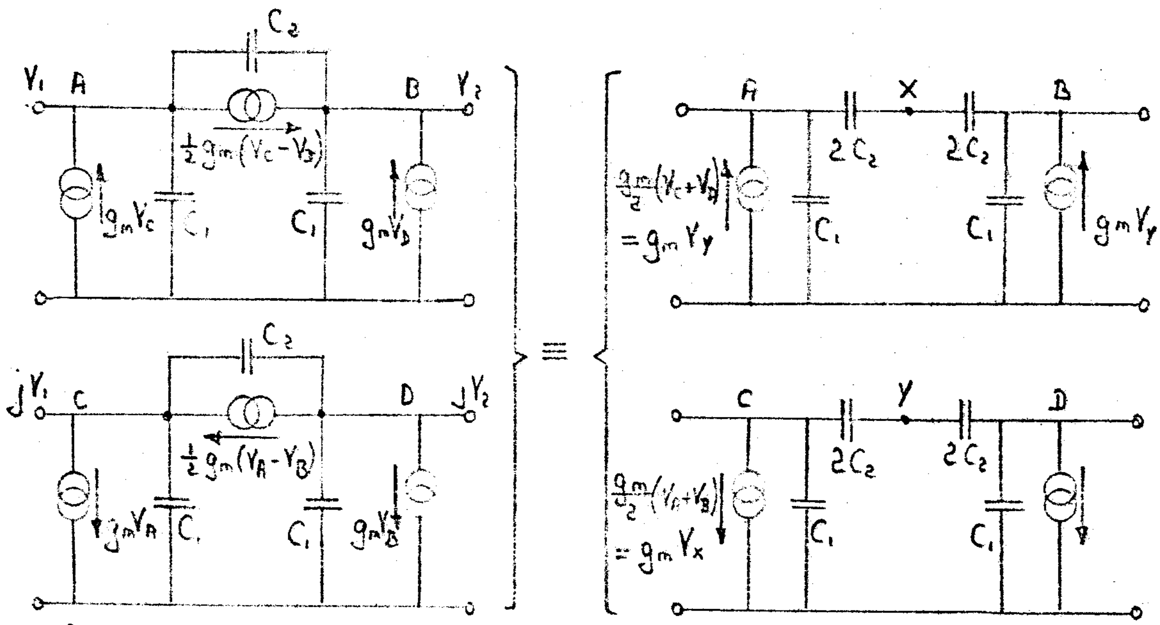
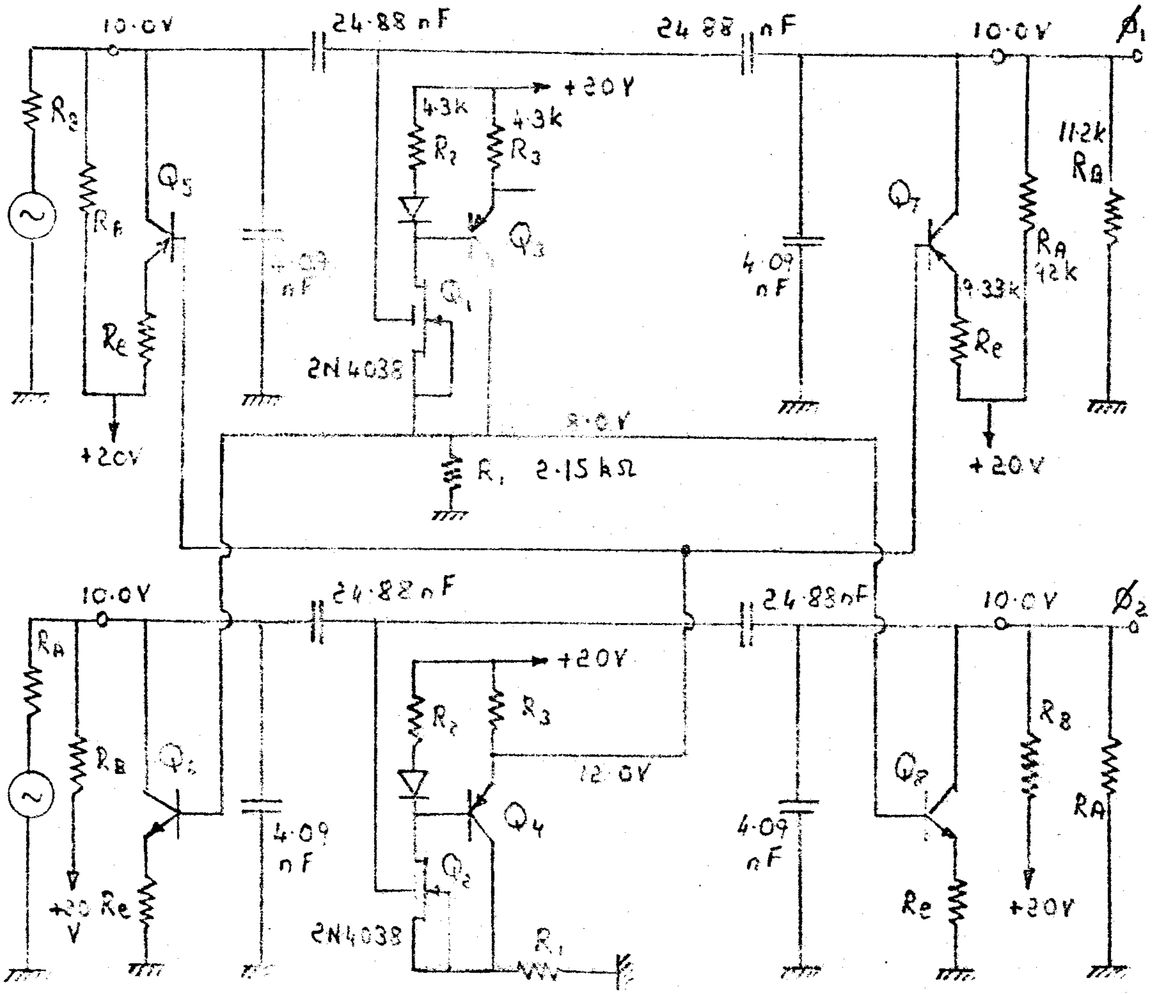


A) Gyrator Equivalent B) Simplified Version

Figure 3.2.4 Two Phase (quadrature) Equivalent of Single Phase Image Section of Fig.3.2.3



Cutoffs 0 &3800 Hz Peak at -625 Hz $Z_o = 7.7k$ R_T= 10kohms

Figure 3.2.5 Practical Filter Section

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is 4300 ohms and the g_m of the 2N4038 input transistors was measured and found to be 2.50 mA/V under the designed operating conditions. From this μ works out to be 0.915 and the controlled source emitter resistors were modified accordingly. An allowance of about 30 ohms was made for the internal emitter resistance of the controlled source transistors. Advantage was taken of the terminating resistors to sink the standing direct current from each controlled source. The terminating resistors were split to provide the correct operating voltage at each collector.

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Computed and Measured Results

Figure 3.2.6 compares the computed response of the theoretical section of Fig. 3.2.3A with the measured response of the practical filter shown in Fig. 3.2.5. The practical response was measured using a two phase oscillator to drive the network and measuring the amplitude at either one of the two output terminals. The two curves show reasonably good agreement allowing for the fact that the phase accuracy and amplitude balance of the oscillator is a severely limiting factor in such measurements. A combination of 2^o phase and 2% amplitude error could cause a 12dB stopband measurement to be out by as much as 0.5dB. As a result, measurements on subsequent networks were made using a single phase drive and quadrature modulators on the output. In this way very much greater measurement precision can be achieved.

3.3 Other types of Polyphase-constrained Networks

The gyrators and N port gyrators used to realise constant reactances may be called polyphase-constrained networks since the n terminals are forced to carry voltages and currents which are

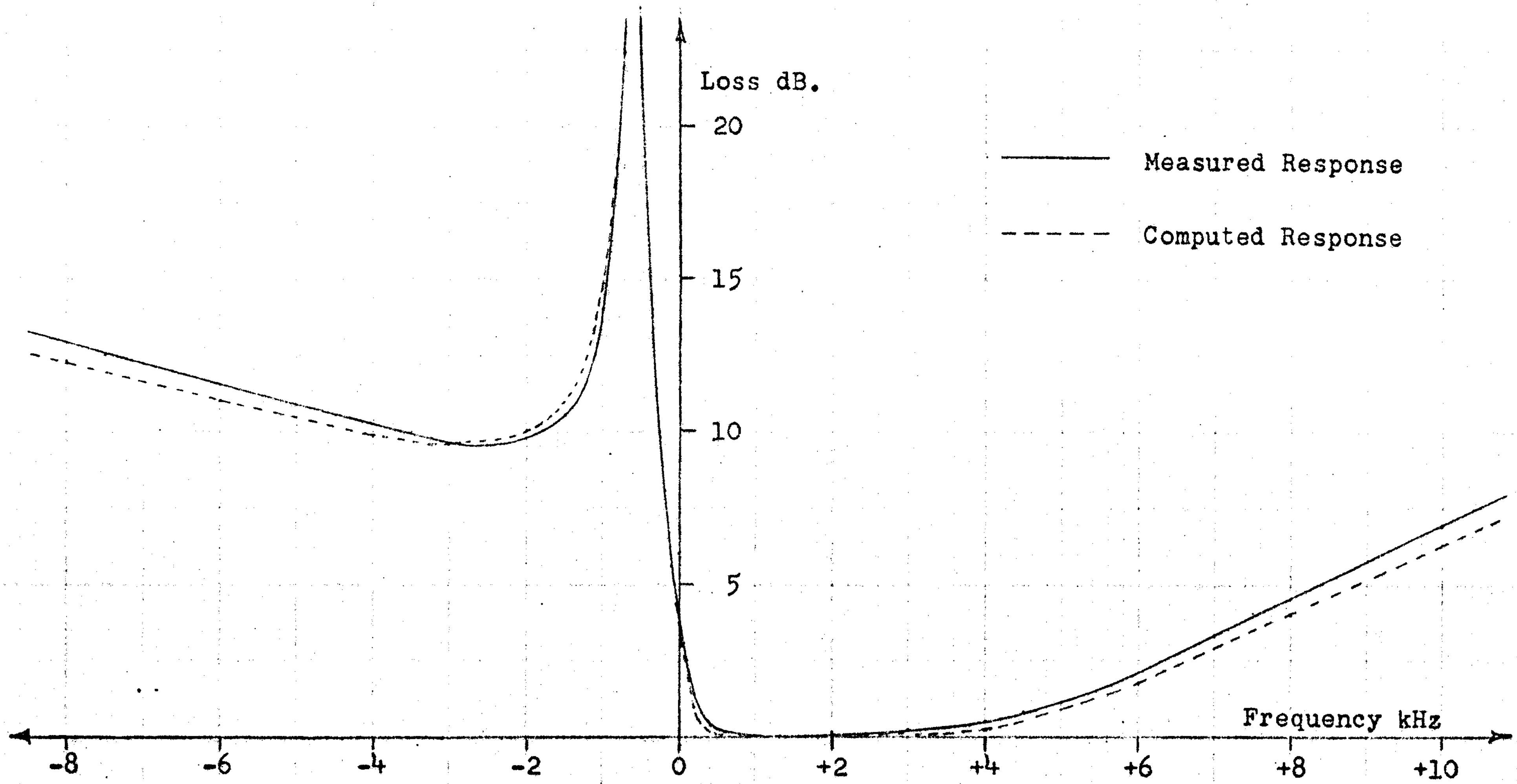

Figure 3.2.6 Frequency response of the filter shown in Fig. 3.2.5

always in n-phase.

An important further class of polyphase constrained networks, particularly useful in polyphase filters are complex impedance transformers of which the most interesting is the 1 : j transformer.

There are a considerable number of possible circuits for their realisation. As an example consider the circuit of Fig. 3.3.1. This is a two phase (quadrature) 1 to j impedance transformer. If impedances Z are placed on the output of each phase, impedances of jZ will be seen looking into either input.

These networks may be used as circuit elements to transform resistors into constant reactances. Fig. 3.3.2 shows a single phase network with a large number of constant reactances. Using 1:j impedance transformers the polyphase version is realised as in Fig. 3.3.3.

If, as in sections with peaks at negative frequency, the reactances X_2, X_4, X_6 etc. are of opposite sign to X_1, X_3, X_5 etc. it would be necessary to add negative impedance convertors to the design of Fig. 3.3.3 and to change the j:1 transformers to 1:j transformers. This detail is shown in Fig. 3.3.4.

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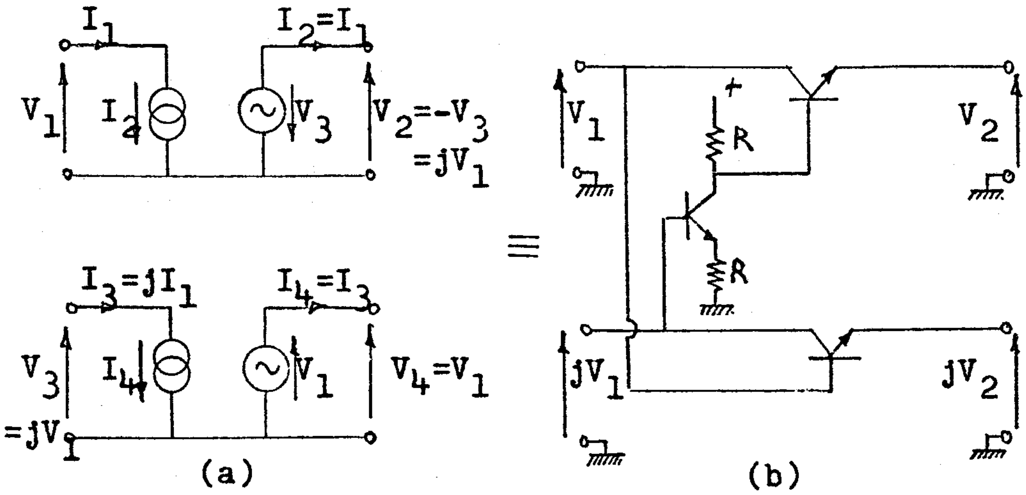

Figure 3.3.1 1:j Impedance Transformer.

(a) Theoretical (b) Transistor realization.

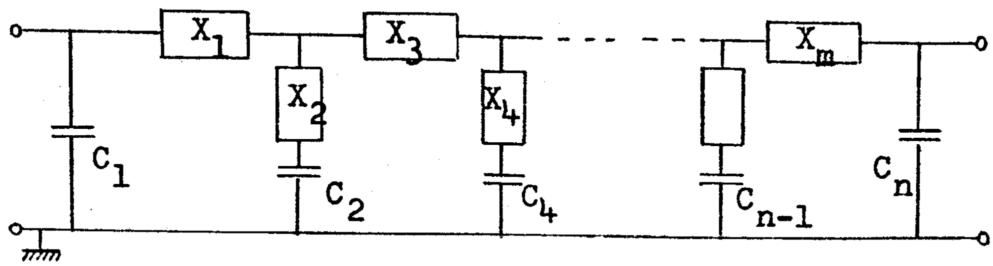

Figure 3.3.2 Filter design requiring a number of constant reactances.

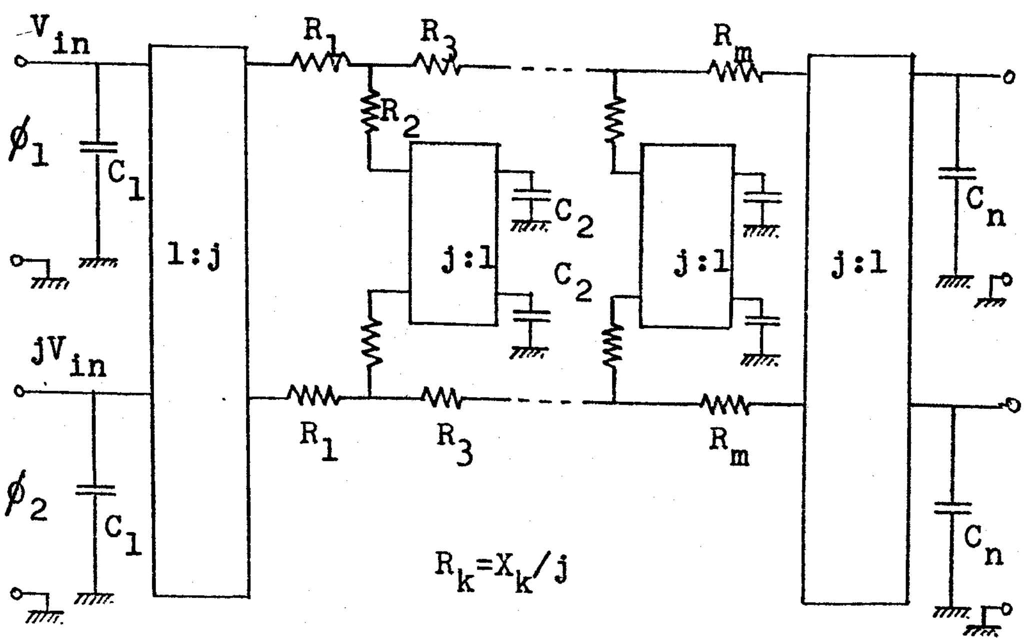

Figure 3.3.3 Two phase (quadrature) realization of Fig 3.3.2 using 1:j transformers.

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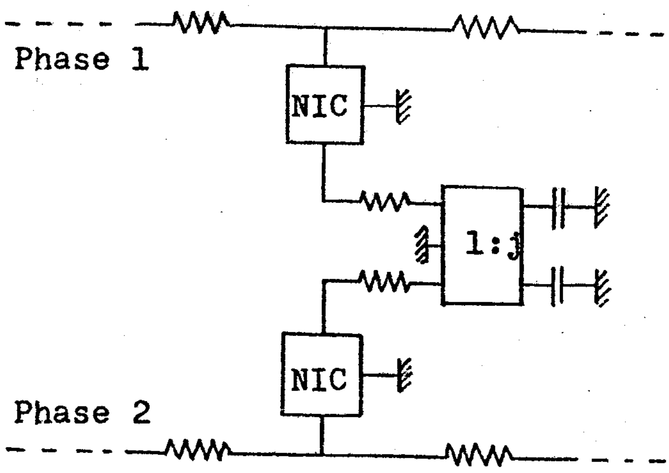

Figure 3.3.4 Section for negative peaks using negative impedance converters.

In practise circuits using negative impedance converters were found to be unstable and the circuit of Figure 3.3.5 is a suitable alternative. The instability was found to be due to the conflicting open circuit/short circuit requirements of the various converters that have to be connected together.

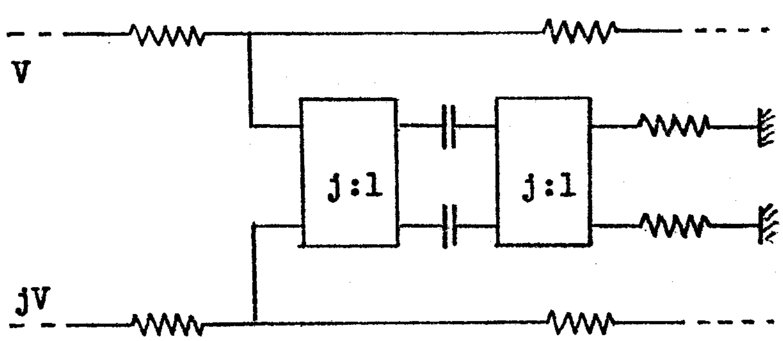

Figure 3.3.5 Stable negative peak section.

By re-arranging the transformer circuit of Fig. 3.3.1 a number of other 1:j impedance transformers can be derived and Figure 3.3.6 shows 4 possibilities. The two phase case is a special version of the 4 phase case which, being symmetrical, contains redundant information in two of its phases.

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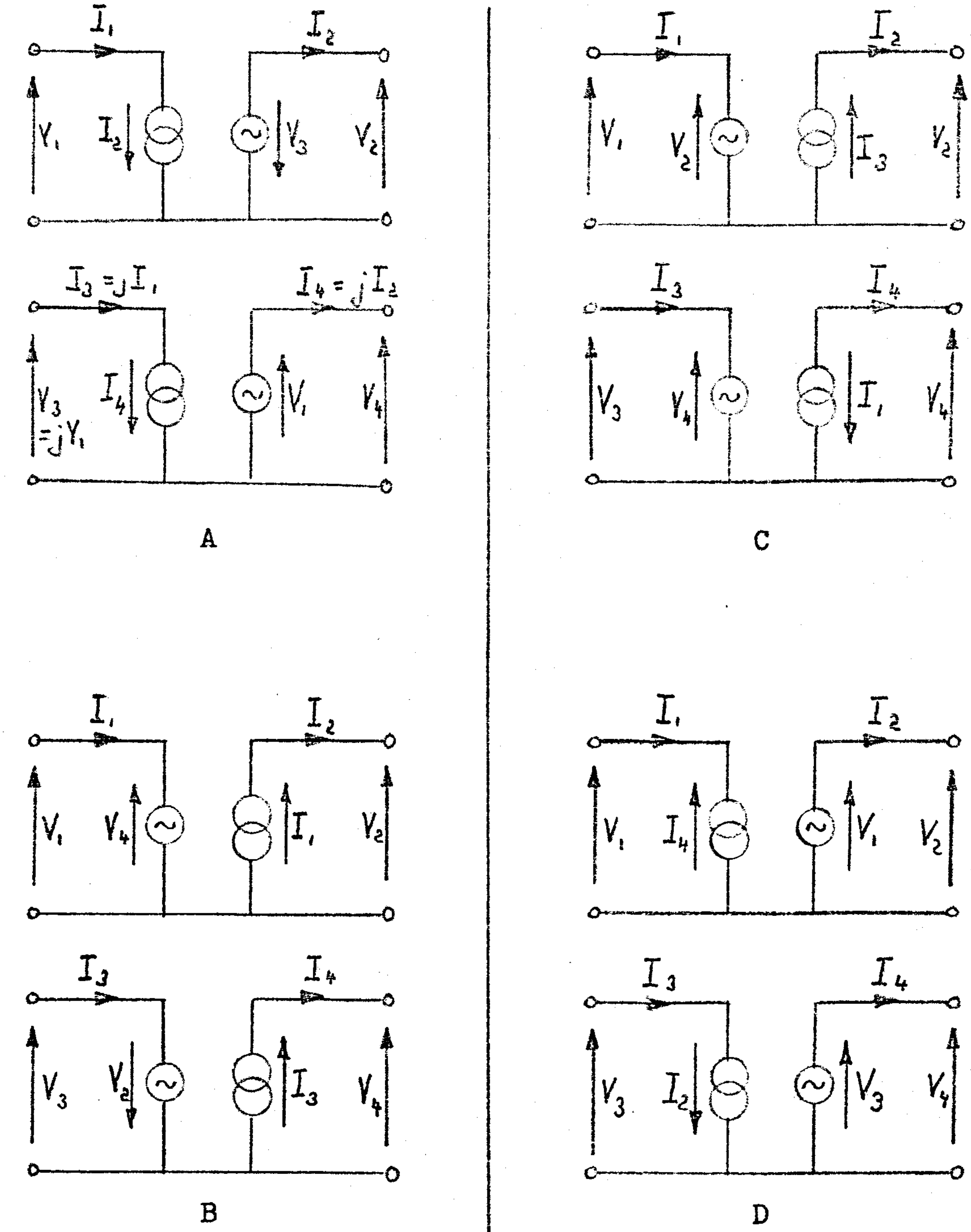

$$\begin{bmatrix} V_1 \\ I_1 \end{bmatrix} = \begin{bmatrix} j & 0 \\ 0 & 1 \end{bmatrix} \begin{bmatrix} V_2 \\ I_2 \end{bmatrix}$$

Voltage shift types.

$$\begin{bmatrix} V_1 \\ I_1 \end{bmatrix} = \begin{bmatrix} 1 & 0 \\ 0 & -j \end{bmatrix} \begin{bmatrix} V_2 \\ I_2 \end{bmatrix}$$

Current shift types.

Figure 3.3.6 Four types of 1:j transformers.

Impedance transformers can be built for any number of phases at the expense of circuit component count. The impedance transformation ratio is not necessarily limited to 1:j as Figure 3.3.7 shows. In this case a 3 phase filter is realised using 1:h transformers where $h = \sqrt[3]{1} = \frac{1}{2} + j\frac{\sqrt{3}}{2}$. Figure 3.3.8 shows one realisation for a 1:h impedance transformer using transistors.

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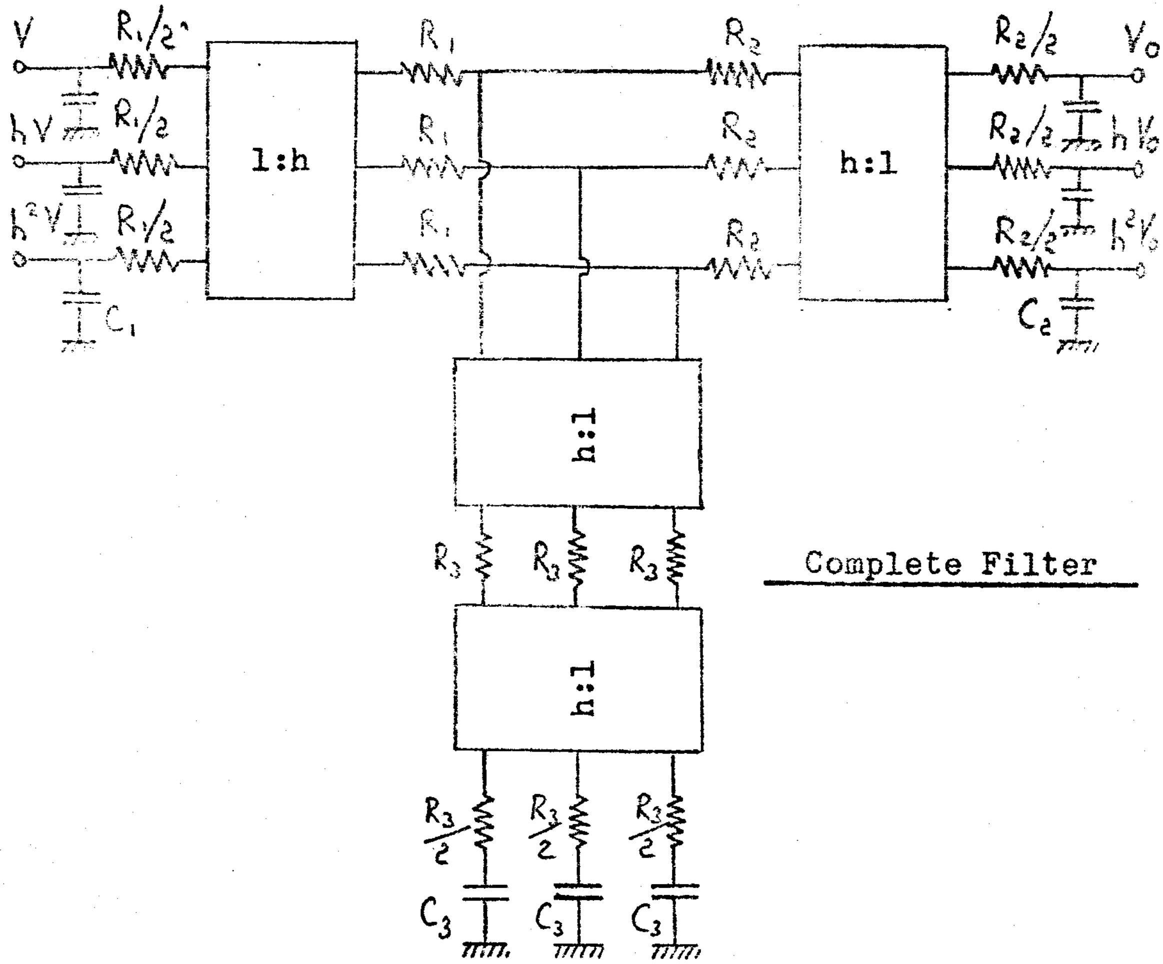
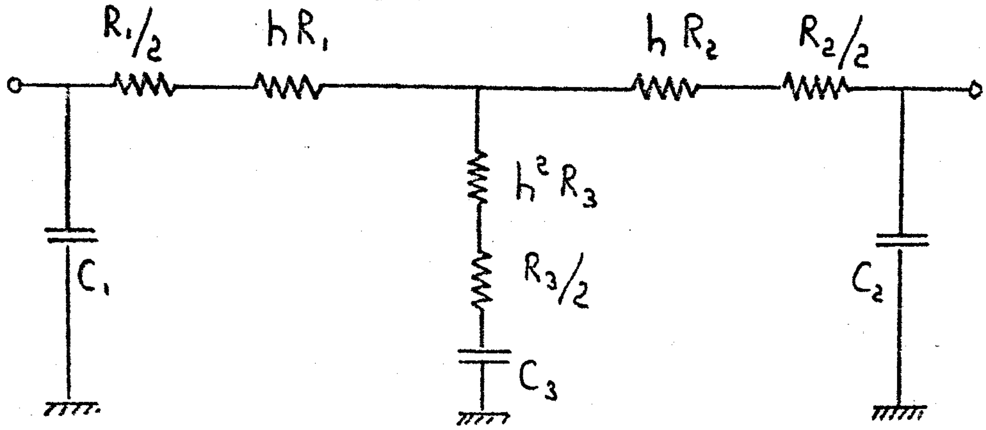


Complete Filter

Single phase equivalent:



Which reduces to:

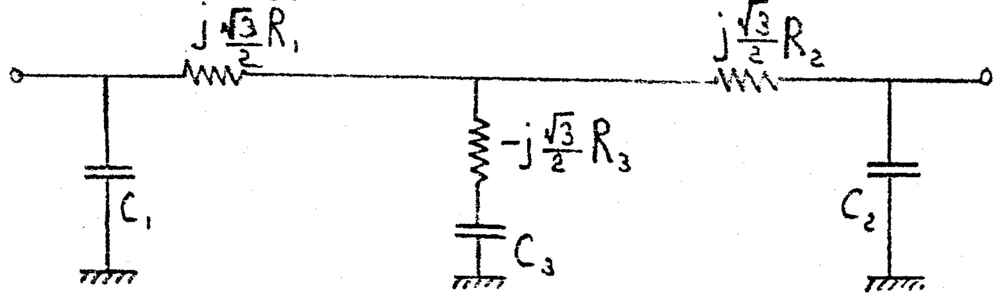


Figure 3.3.7 Example of a 3 phase filter using 1:h transformers. $h=(-1+j\sqrt{3})/2$

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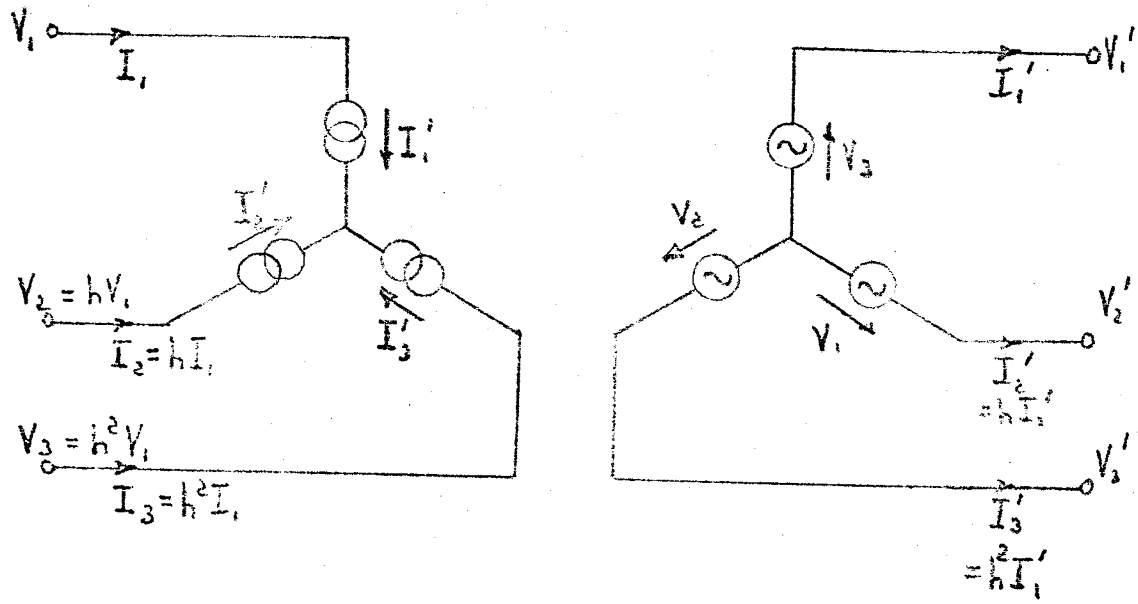

$$\begin{bmatrix} V_1 \\ I_1 \end{bmatrix} = \begin{bmatrix} h & 0 \\ 0 & 1 \end{bmatrix} \cdot \begin{bmatrix} V_1' \\ I_1' \end{bmatrix}$$

$$h = \sqrt[3]{1} = -\tfrac{1}{2} + j\sqrt{3}/2$$

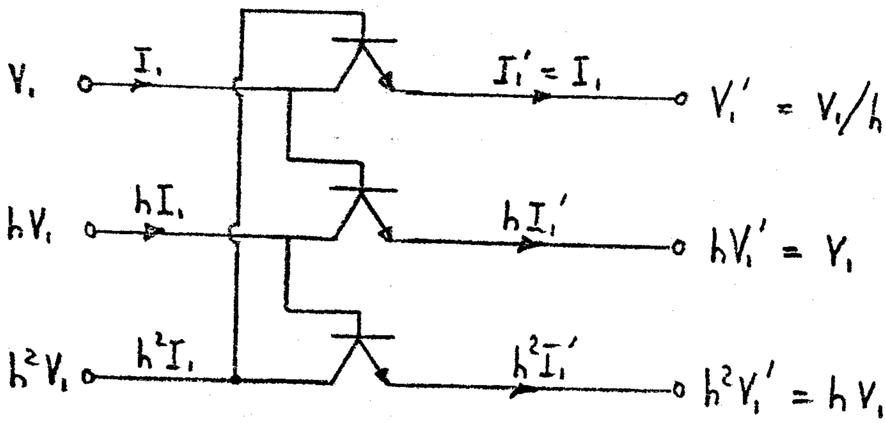

Figure 3.3.8 1:h Impedance transformer.

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3.3.1 Practical Problems in Realizations Using 1:j Impedance Transformers

Stability

The first problem which presents itself when using 1:j transformers is that of stability. Initial attempts resulted in circuits which either latched up in unworkable d.c. bias states on oscillated at high frequencies. This can be explained by considering the ideal equivalent circuit of Figure 3.3.1(a). Each port is connected either to a current source or a voltage source. A current source must not be allowed to go open circuit otherwise an infinite voltage will be developed at its terminals. Similarly an infinite current will result if a voltage source is short circuited. For stability, therefore, it is important to choose the right type of transformers to connect different circuits. In Figure 3.3.3, for example, the first transformer has a capacitor across each input port. Now a capacitor goes to zero impedance at infinite frequency so the port to which it is connected must be short circuit stable. Although the capacitor also has infinite impedance at d.c. there will be a source resistor in the input drive circuit which will prevent the transformer seeing an open circuit so there is no need for open circuit stability. Using a current source port will therefore be satisfactory for this case. By considering the impedance connected to each port of each transformer it is therefore possible to select an appropriate type which will result in overall circuit stability.

Cascading Sections

Since each section requires power for correct

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operation it is necessary to make provision for this when connecting them together. This can be done by providing a d.c. supply current into each end of the network which is then distributed to each section via the resistor ladders which form part of each phase of the filter. Figures 3.3.9 and 3.3.10 show positive and negative peak sections respectively. Current is supplied through the 4 resistors R_1 and passes through the transformer and is terminated by a current sink which does not affect dynamic working of the circuit. In the negative peak section d.c. is blocked by capacitors so an additional current source and sink is needed for the second transformer. The final sink can be provided by the resistor R_2 which also forms part of the filter.

Transformer Imperfections

An ideal 1:j transformer may have the hybrid parameter matrix

$$\begin{bmatrix} V_1 \\ V_2 \\ I_3 \\ I_4 \end{bmatrix} = \begin{bmatrix} 0 & 0 & 0 & -1 \\ 0 & 0 & 1 & 0 \\ -1 & 0 & 0 & 0 \\ 0 & -1 & 0 & 0 \end{bmatrix} \begin{bmatrix} I_1 \\ I_2 \\ V_3 \\ V_4 \end{bmatrix}$$

This state cannot be achieved in practise due to the imperfect nature of the transistor. The principle errors will consist of finite input and output impedances, non unity transfer ratios and unwanted cross coupling. Even with modern high quality silicon planar transistors these effects, although

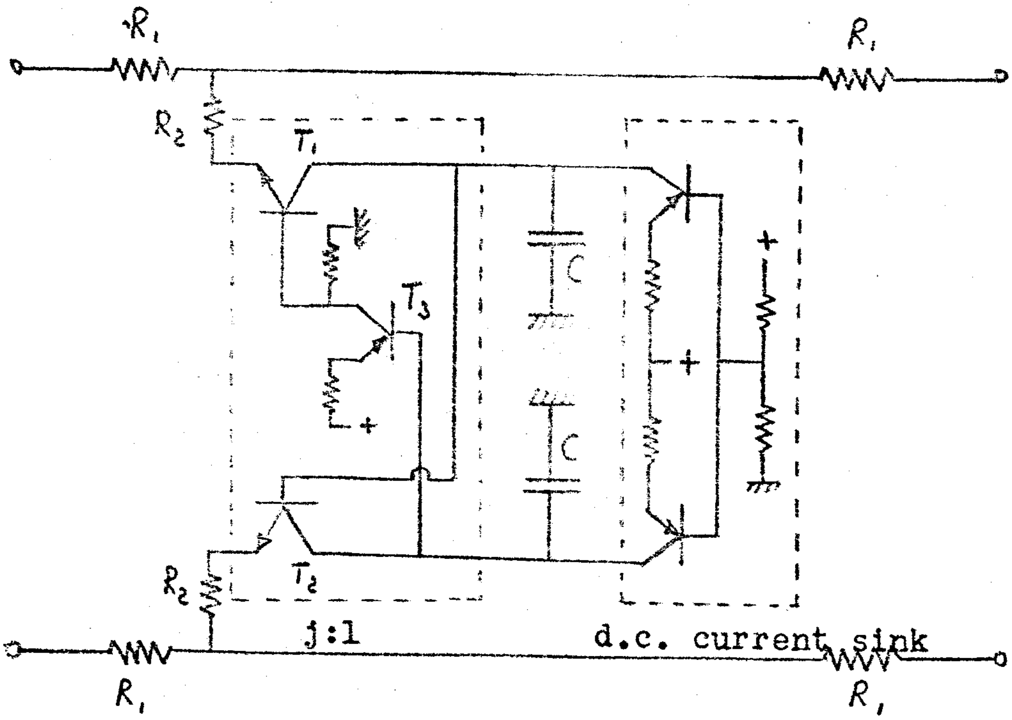
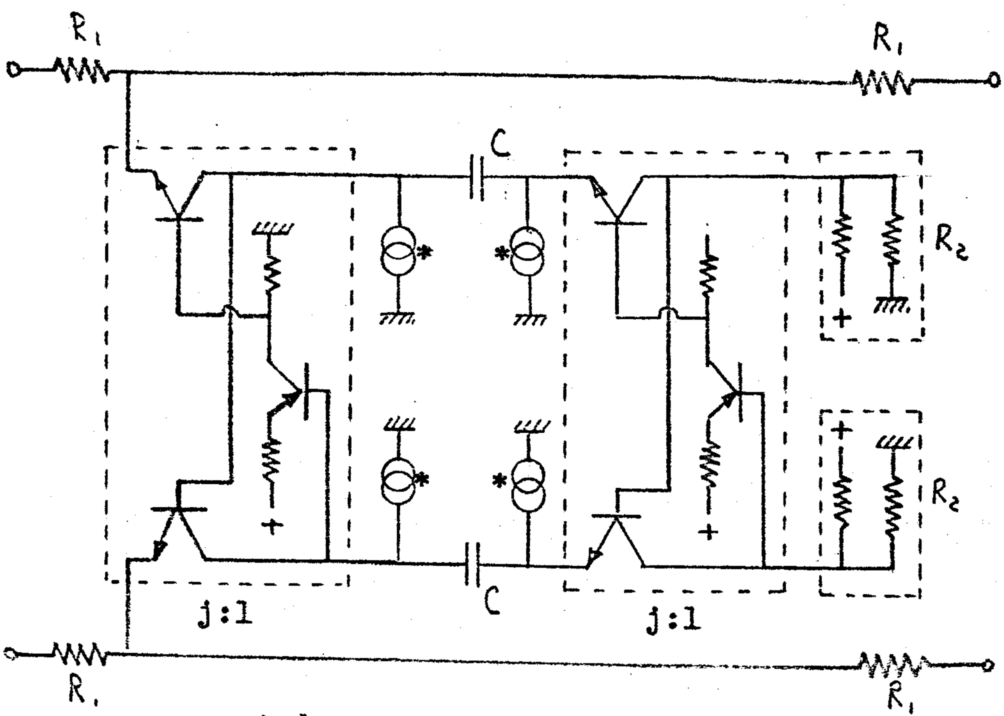


Figure 3.3.9   Positive peak section.



* d.c. current sinks

Figure 3.3.10   Negative peak sections.

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small, are sufficient to seriously degrade the overall filter performance and must therefore be compensated for.

To evaluate the likely performance a nodal analysis computer program was used to determine the equivalent hybrid matrix of transformers using practical transistors. Hybrid parameters are most convenient as the important terms can be seen immediately as equivalent circuit elements. Figure 3.3.11 shows how the major imperfections can be represented in this way.

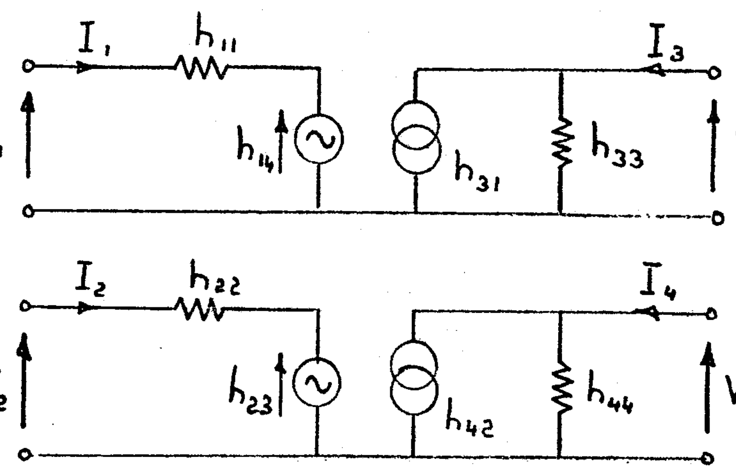

Figure 3.3.11 1:j Impedance transformer imperfections

Transistors (NPN BC107A) were represented in the analysis by an equivalent circuit derived from the d.c. 'h' parameters given in the manufacturers data. It was considered important to operate the transistors at as low a current as possible to minimise power dissipation. Having decided on the current the transistor parameters were obtained from curves (Figure 3.3.12) and substituted in the equivalent circuit for evaluation. Three cases were considered (Figure

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3.3.13) and the results are tabulated in Figure 3.3.14.

Case 1 Basic 3 transistor circuit.

With all transistors operating at 0.1 mA the inverting transistor plays a large part in degrading the performance and significant differences occur between h_{11} and h_{22} the input impedances to phase 1 and 2. This is mainly due to the load resistor (100k) of T_3 being divided by the gain of T (100) appearing effectively in series with the emitter of T_1. By reducing the load resistor and increasing the current of T_3 to 1mA this effect can be reduced considerably.

An additional problem with the basic circuit is that T_2 in phase 2 puts a significant loading on phase 1 due to its finite current gain. This is shown in the results (Fig. 3.3.14) where h_{32} is -.00762.

Case 2 3 transistor circuit using complementary pairs.

By combining an NPN and a PNP transistor together it is possible to make the equivalent of a super NPN transistor. This can improve the performance in some respects although the output impedance (1/hoe) of a complementary pair is low (∿ 30kΩ) and causes unwanted coupling between input and output (h_{13} and h_{24}).

Case 3 Single transistors but with the addition of an emitter follower driving T_2.

The emitter follower acts as a buffer between the output of phase 1 and the base of T_2. As a result h_{32} is

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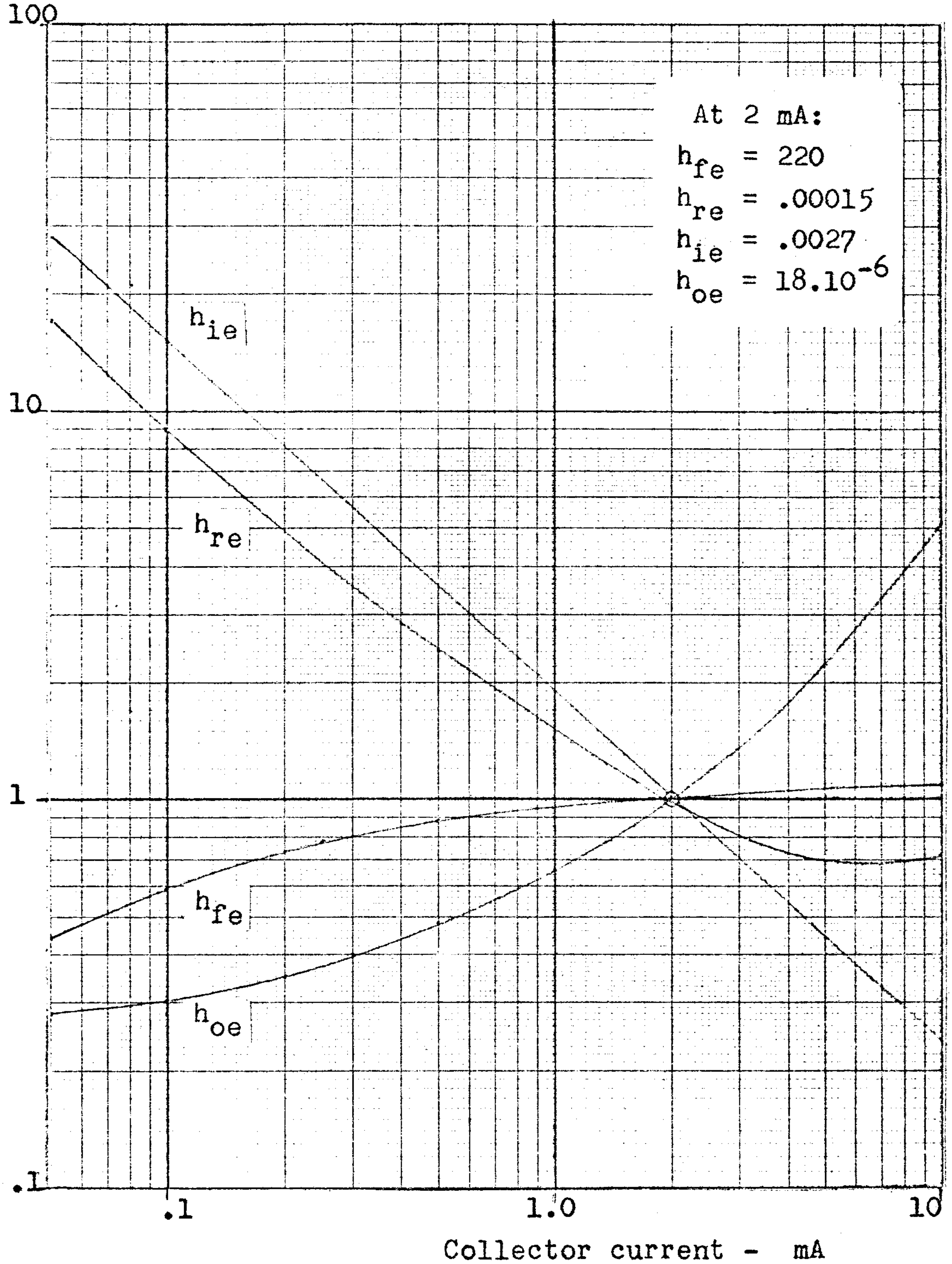

Figure 3.3.12 Variation of Common Emitter Parameters against Collector current for the BC107A transistor.

(Relative to 2 mA I_c with V_{ce} = 5 Volts.)

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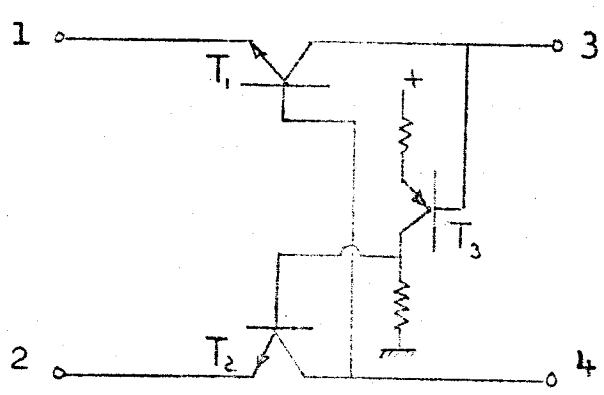
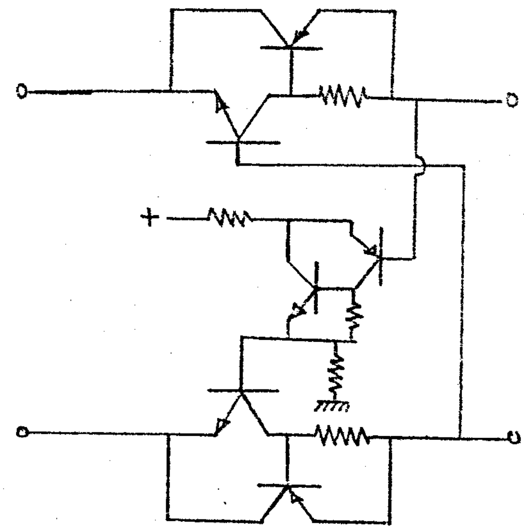
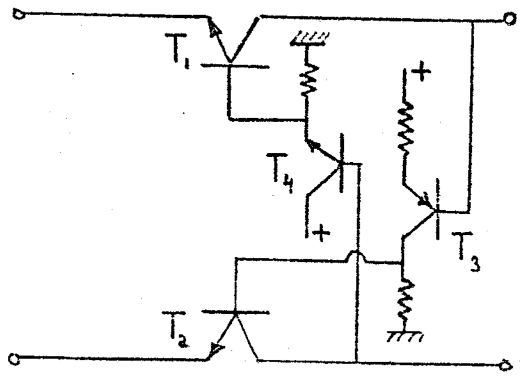


Case 1: Single transistors.



Case 2: Complementary NPN/PNP pairs.



Case 3: With an additional emitter follower.

Figure 3.3.13 Three versions of the basic 1:j transformer studied for comparative performance.

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h Parameter Matrices for 1:j Transformers

Case 1 Three Transistor Transformer

(a) All transistors carrying .1mA Ic

$$\begin{bmatrix} V_1 \\ V_2 \\ I_3 \\ I_4 \end{bmatrix} = \begin{bmatrix} \underline{1058} & .16.10^{-9} & .505.10^{-2} & -.976 \\ -.3.10^{-10} & \underline{305} & .999 & .530.10^{-3} \\ -.9913 & -.762.10^{-2} & .991.10^{-6} & -.10.\ 10^{-5} \\ -.344.10^{-4} & -.9921 & .978.10^{-6} & .11.\ 10^{-5} \end{bmatrix} \begin{bmatrix} I_1 \\ I_2 \\ V_3 \\ V_4 \end{bmatrix}$$

(b) Increasing the inverter current to 1mA

$$\begin{bmatrix} V_1 \\ V_2 \\ I_3 \\ I_4 \end{bmatrix} = \begin{bmatrix} 381 & .8.10^{-10} & .987.10^{-3} & -.998 \\ -.56.10^{-10} & 305 & .999 & .530.10^{-3} \\ -.992 & \underline{-.762.10^{-2}} & .987.10^{-6} & -.102.10^{-5} \\ -.404.10^{-5} & -.992 & .984.10^{-6} & .932.10^{-6} \end{bmatrix} \begin{bmatrix} I_1 \\ I_2 \\ V_3 \\ V_4 \end{bmatrix}$$

Case 2 Complementary Pairs

All transistors biased at .05mA Ic

$$\begin{bmatrix} V_1 \\ V_2 \\ I_3 \\ I_4 \end{bmatrix} = \begin{bmatrix} 202 & .135.10^{-9} & \underline{.873.10^{-2}} & -.972 \\ .5.10^{-9} & 101 & .997 & \underline{.311.10^{-2}} \\ .999 & -.103.10^{-2} & .690.10^{-7} & -.207.10^{-6} \\ -.580.10^{-5} & -.999 & .153.10^{-6} & .134.10^{-6} \end{bmatrix} \begin{bmatrix} I_1 \\ I_2 \\ V_3 \\ V_4 \end{bmatrix}$$

Case 3 Using an Emitter follower, inverting transistor at 1mA

$$\begin{bmatrix} V_1 \\ V_2 \\ I_3 \\ I_4 \end{bmatrix} = \begin{bmatrix} 381 & .294.10^{-9} & .363.10^{-3} & -.994 \\ .585.10^{-10} & 307 & .996 & .544.10^{-3} \\ -.992 & -.579.10^{-4} & .103.10^{-6} & .213.10^{-7} \\ -.418.10^{-5} & -.992 & -.479.10^{-7} & .653.10^{-6} \end{bmatrix} \begin{bmatrix} I_1 \\ I_2 \\ V_3 \\ V_4 \end{bmatrix}$$

Fig. 3.3.14

reduced to a more acceptable level. Transfer ratios are in error by up to 0.8% although the match between phases is much better at 0.2%. Parameters h_{13} and h_{24} remain significant and can only be improved with this simple circuit by choosing transistors with lower output admittance (hoe).

Input impedances h_{11} and h_{22} appear as resistors in series with the input which can be compensated for by absorbing them into the resistors already in the network. In negative peak sections, however, there are two j:1 transformers (Figure 3.3.10) only one of which can be compensated for by increasing the value of the network resistor R by an equal amount. The input resistance of the second transformer appears as a real resistor in series with the network capacitor. In practise this can reduce the circuit Q to as low as 20. The input resistance can be reduced at the expense of power dissipation by increasing the operating current. Complementary pairs reduce the input impedance but they also introduce unwanted coupling terms h_{13} and h_{24} which also worsens overall filter performance particularly in the negative frequency side of the stopband.

Output impedance in the 1:j transformer is sufficiently large to be negligible. The lowest impedance found was of the order of 1MΩ. Output impedance affects either the Q factor, producing Q's of the order of several hundred, or modifies the resonant frequency in negative peak sections by about 0.1%. This should normally be unimportant providing each phase is modified by the same amount.

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3.3.2 A Complete Active Polyphase Filter

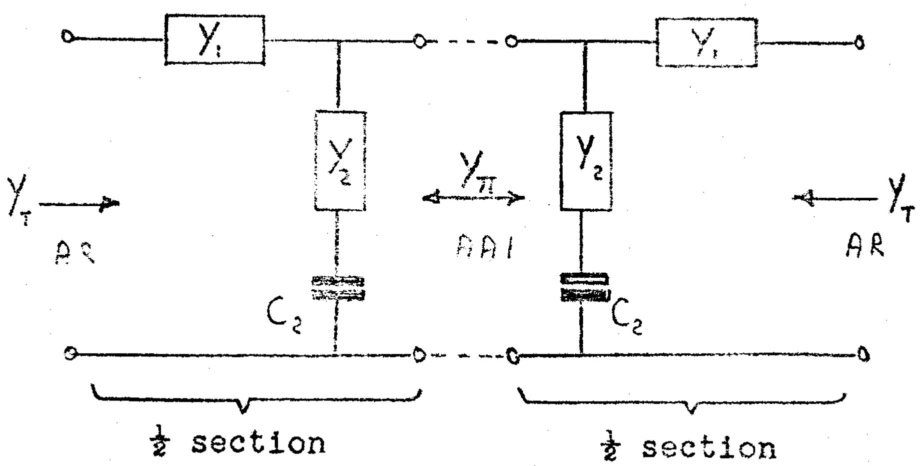
A complete 5 section 2 phase filter suitable for single sideband modulation was designed built and tested, the results being compared with computer analysis of the equivalent circuit. The design was based on the 1:j transformer technique and the theoretical network was designed using the image techniques of section 2.2.

3.3.3 Theoretical Design

Design was based on the following requirements

Passband:	300 to 3400 Hz	$<\frac{1}{2}$dB ripple
Stopband:	$-\infty$ to -600Hz	>50dB
	$+4300$ to $+\infty$Hz	>50dB

An image design was chosen because it allows the simple canonic minimum capacitor structure of Figure 3.3.2 to be achieved directly without recourse to an insertion loss design which would not guarantee the same simplicity. An additional advantage during experimentation is that section order can be changed at any time even to the point of leaving sections out. Further, any purely shunt admittance might cause a severe power supply drain and cause additional complications in the power supply arrangements. A single type of image section is used as shown in Figure 3.3.15 for both positive and negative stopband peaks and also for terminating half sections with peaks at infinity. The normalised section has image cut off frequencies at $\omega=0$ and $\omega=1$ which will be modified by the terminating resistors to give an overall bandwidth of somewhat less than 1.0. Full T sections are connected together at the T port which has an



$Y_1 = -2j/m$

$Y_2 = -2jm/(1-m^2)$

$C_2 = 2m$

$m^2 = \dfrac{\omega_\infty - 1}{\omega_\infty}$

$Y_T = 2\sqrt{\dfrac{\omega}{1-\omega}}$

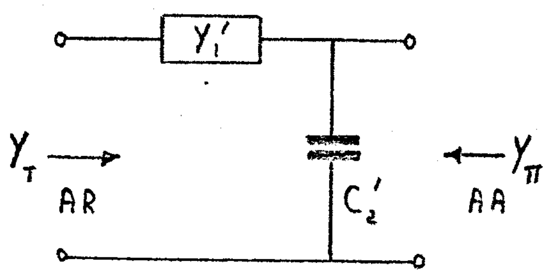
$Y_\pi = \dfrac{2\sqrt{\omega(1-\omega)}}{1+\omega(1-m^2)}$

$q = m\sqrt{\dfrac{\omega}{\omega-1}}$

Image loss (full sect.)
$= 20\log_{10}\left[\dfrac{1+q}{1-q}\right]$ dB.

Special Case: $m=1$ $\qquad \omega_\infty = \infty$

(for terminating ½ sections)



$Y_1' = -2j$

$C_2' = 2$

$Y_\pi = 2\sqrt{\omega(1-\omega)}$

Figure 3.3.15 Minimum Capacitor Image Section.

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anti resonant-resonant type image admittance independant of the peak frequency. The network is terminated by half sections where m=1 so that ω_ω is at infinity and the Π port image admittance is presented to the outside and is of the anti resonant-anti resonant form ie. zero at each cut-off rising to unity at midband.

The pole frequencies were determined using an interactive computer program which sums the image losses due to N sections, listing the frequency response and allowing manual optimisation to meet any arbitary stopband specification. An alternative completely manual method is possible using templates. The frequency specification is transformed to the λ scale where $\frac{\omega}{\omega-1} = e^{4\lambda}$ and then the image loss is the sum of a number of identical templates of the form

$$20 \log_{10} \mathrm{Coth}(\lambda) \quad \mathrm{dB}$$

each displaced by an amount selected by the designer. For each template displaced by an amount λ, the corresponding section m value is given by

$$m = e^{-2\lambda}$$

Five sections were needed to meet the specification, two poles in the positive stopband, two in the negative stopband and one at infinity split into half sections at each end of the filter to provide suitable image impedances for termination. A standard technique with image filters is to

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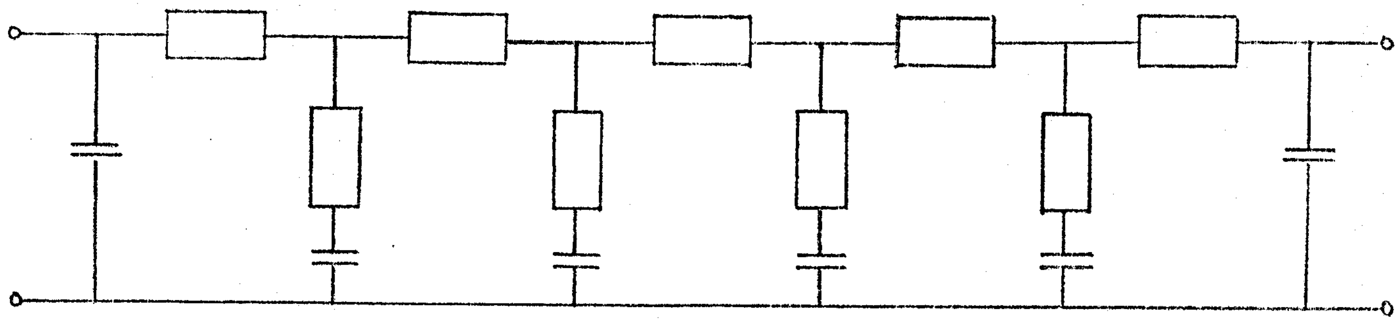
mismatch the terminating resistors to extend the effective bandwidth. In this case the filter impedance level was scaled up to 10kΩ while actually terminating in 14kΩ since the image impedance will rise from 10kΩ at midband to infinity at the cut-offs. The theoretical cut off frequencies are at zero and 3.8kHz and this degree of mismatch provides 0.5dB passband ripple from 0.245 to 3.56kHz.

The complete circuit for the theoretical filter is given in Figure 3.3.16. The section order was chosen to give the two poles closest to the passband the closest tolerance components, placing them at the output end of the filter where accuracy also helps phase balance.

3.3.4 Practical Implementation

The complete filter was constructed using the 1:j transformer technique with transformers including the additional emitter follower (Figure 3.3.13 - Case 3).

An overall block diagram of the filter is given in Fig. 3.3.17. The diagram shows a single phase input drive for single sideband modulation. Power supply arrangements are also shown. A current of 0.1 mA flows out of each transformer and shunt branch. All currents flow towards the middle of the network summing to a total of 0.6 mA which is absorbed in a high a.c. impedance sink. A constant potential difference of 5 volts between the two phases is maintained throughout for convenience in biassing the various transformers. Figures 3.3.18 to 3.3.20 show the input and output transformers and details of the shunt branches for positive and negative poles.



R		j7.2488		j12.391		j11.749		j15.290		j18.444		kilohms
"			j3.7811		-j3.6895		j5.8468		-j5.7921			"
C	8.377		8.335		33.183		6.1865		45.0455		8.377	nF.
F			+5050		-1300		+4400		-610			Hz.

Image impedance at midband : 10k Ω
Source and load resistors : 14k Ω

<u>Figure 3.3.16</u> <u>Theoretical single phase filter design.</u>

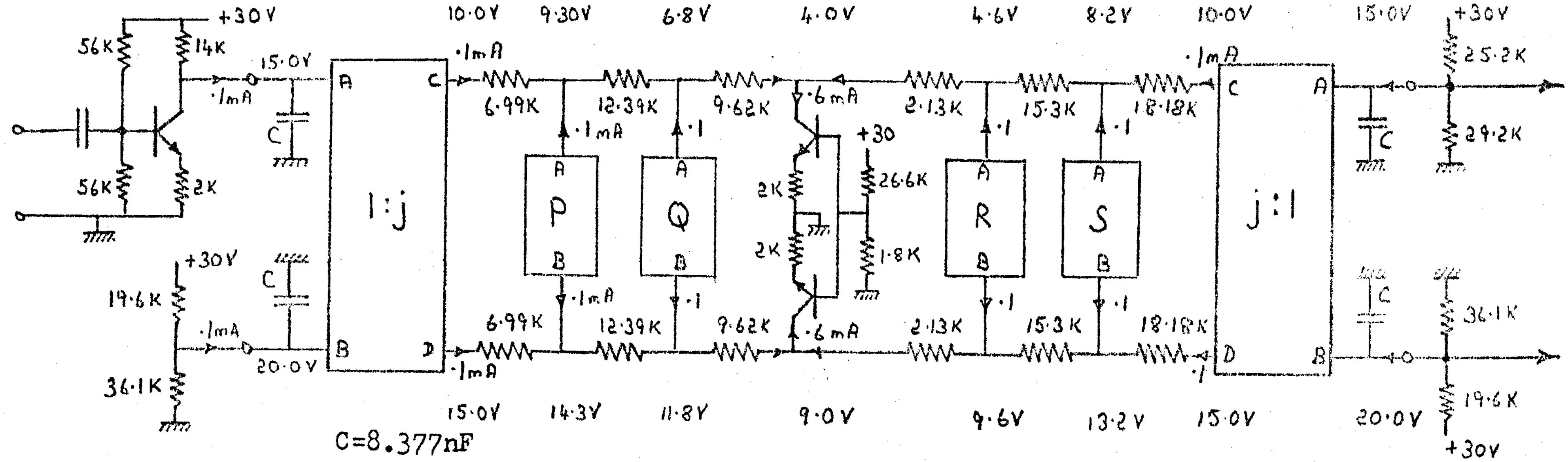


Figure 3.3.17 Overall Filter Structure
-including power supply arrangements.

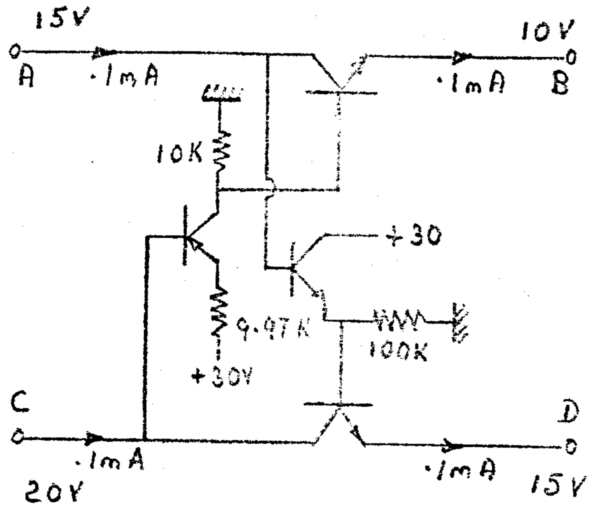
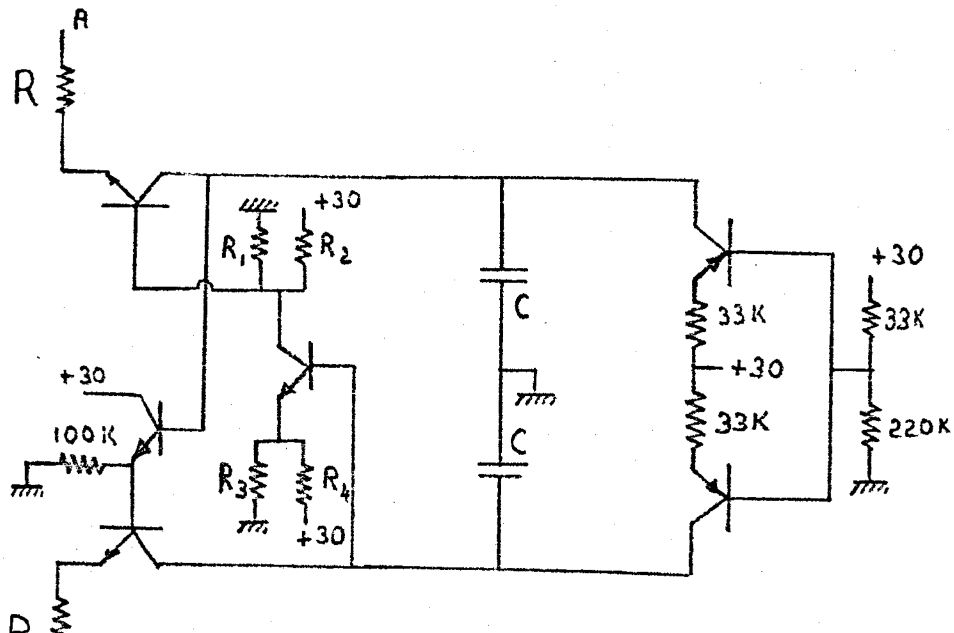


Figure 3.3.18 Input and Output Transformers.



	Section P	Section R
R	3.521 k	5.587 k
C	8.335 nF	6.186 nF
R_1	15.6 k	10.0 k
R_2	29.0 k	--
R_3	--	37.4 k
R_4	10.0 k	13.5 k

Figure 3.3.19 Positive peak sections P & R

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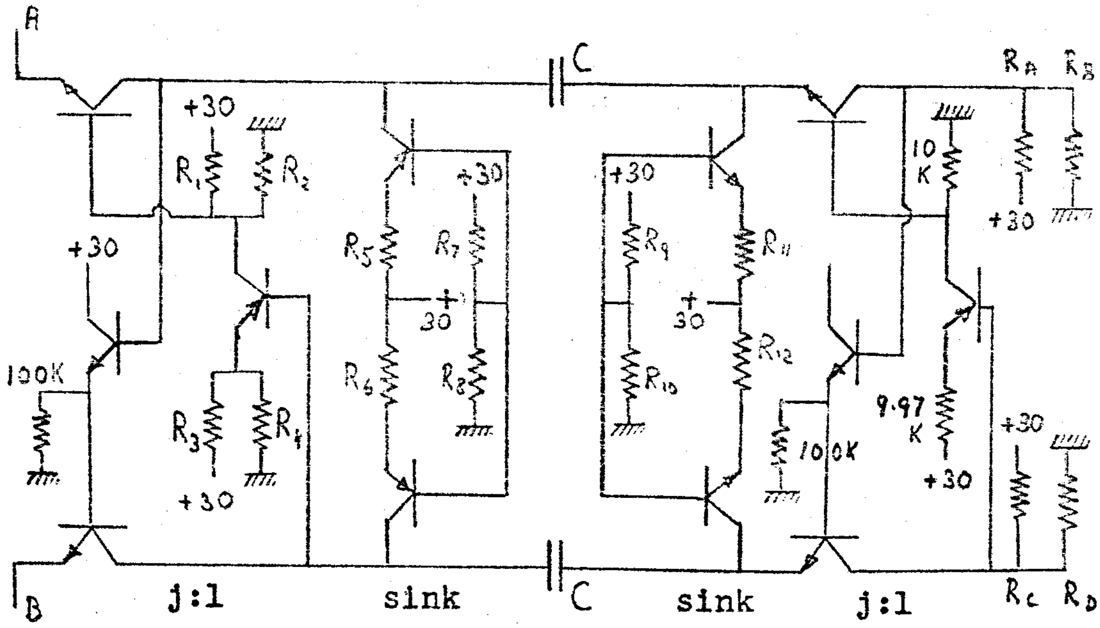

	Section Q	Section S
C	33.18 nF	45.045 nF
R_A	8.097 k	12.407 k
R_B	8.097 k	12.407 k
R_C	6.072 k	9.305 k
R_D	12.14 k	18.610 k
R_1	29 k	-- k
R_2	15.6 k	10 k
R_3	10 k	10.6 k
R_4	--	160 k
R_5	47 k	33 k
R_6	47 k	33 k
R_7	68 k	33 k
R_8	330 k	220 k
R_9	330 k	220 k
R_{10}	68 k	33 k
R_{11}	47 k	33 k
R_{12}	47 k	33 k

Figure 3.3.20 Negative peak sections Q & S

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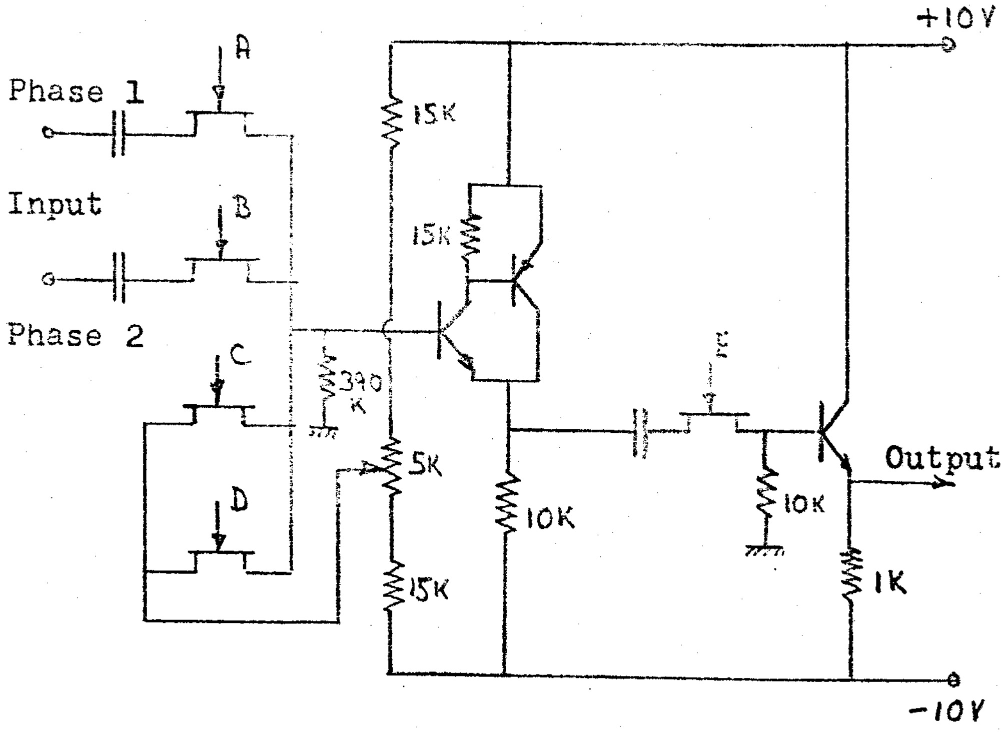

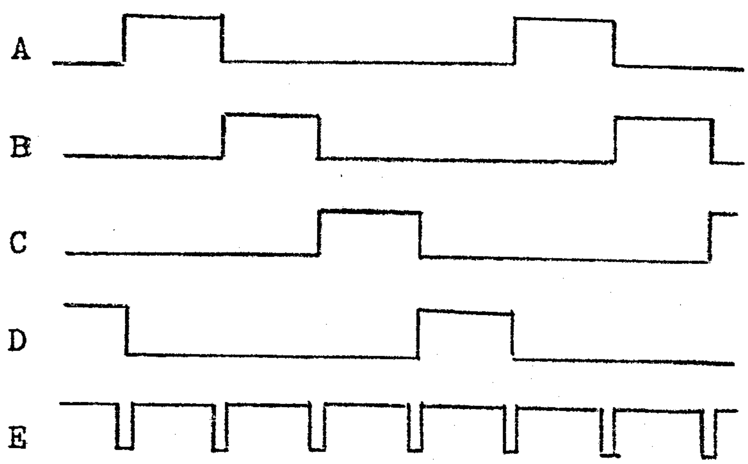

Modulator drive waveforms

Figure 3.3.21 Two phase modulator for testing two phase filters.

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For positive poles the non unity transfer ratio of the transformers gives an error of 1.25% with an out of balance of 0.1% and this has been corrected for by decreasing the capacitor on phase 1 by 1.3% and that on phase 2 by 1.2%. Negative pole section capacitors must be modified in the same way but in addition the resistors are affected by two transformers and need increasing by 2.5% in both phases.

Errors in the input and output transformer transfer ratios have little effect as the filter will be insensitive to overall impedance changes of such small amounts. Unbalance between the two phases will be more critical; particularly at the output where a 1% unbalance can reduce the maximum stopband loss at negative frequencies (up to -4kHz) to 45dB.

Correction was also made for the h_{11} and h_{22} input impedances of the transformers. At the input and output and for the positive peak sections the corresponding resistors were reduced by 260Ω equivalent to the transistor internal emitter resistances for 0.1 mA d.c. operating current. For negative peak sections resistors were compensated, increasing by 260Ω in addition to the correction for non-unity transfer ratio. No correction was possible for the input resistance of the second transformer and so experiments were tried using compound transistor pairs for T_1 and T_2 to observe the effect of a reduction in h_{11} and h_{22}.

Figure 3.3.21 shows the modulator used for testing the filter. Series field effect transistors were used as these can behave as near perfect on-off switches at modest carrier frequencies (up to 200kHz). Two extra modulators are provided for symmetry and to reduce carrier leak. The modulators drive into a high input impedance buffer amplifier followed by a retiming gate which improves the precision of the quadrature modulation. Control voltages for the modulators are derived from a divide by four logic circuit driven from an input at four times carrier frequency.

3.3.5 Results and Conclusions

Figure 3.3.22 shows the ideal filter frequency response (a) when correctly terminated together with various curves corresponding to measured and computer results.

Curve (b) shows the measured response with a substantial loss of discrimination at negative frequencies. A significant slope across the passband and a reduced upper cut off frequency are also in evidence. These effects were confirmed as being largely due to the 1:j transformer imperfections by a computer nodal analysis - curve (d). The analysis was limited by the number of nodes to simulating only the last 2½ sections with a full transistor equivalent circuit. The results do show very similar degradations and their nature indicates that the fault lies in asymmetry between the two phases. Since the practical model was measured using modulators on the output some small part of the degradation will be due to modulator imperfections such as phase and amplitude unbalance. Subsequent tests on the modulators showed phase and amplitude errors of less than 0.1° and 0.01dB respectively so that the effect would be negligible at 40dB rising to significance only at 60dB filter discrimination.

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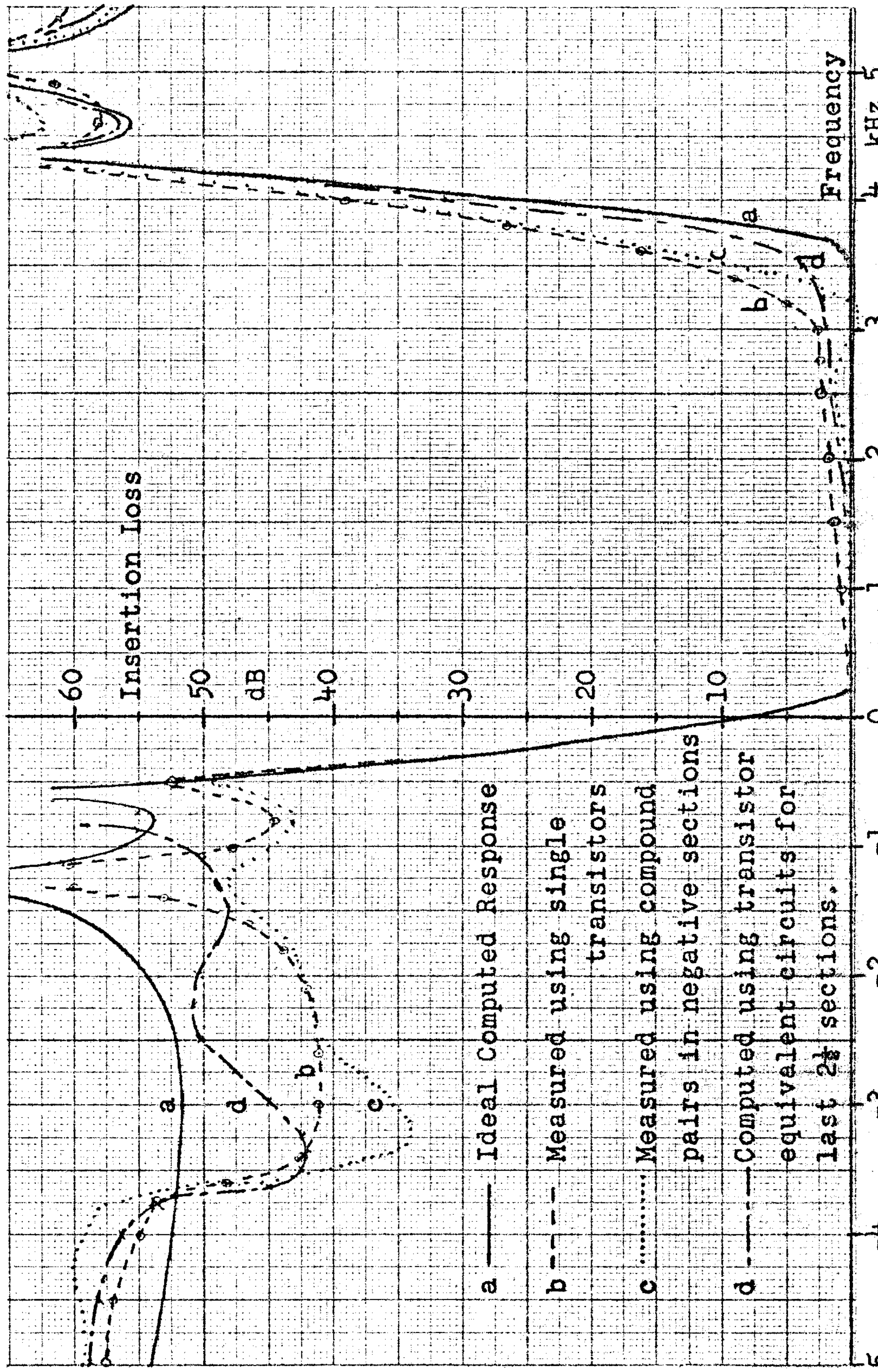

Figure 3.3.22 5 Section Active 2-phase Filter.

Measured and Computed Responses.

A limited study of component tolerances required with ideal transformers showed that a component match of better than 1% is required in the last section if 45dB minimum discrimination is to be maintained while up to 5% in the input section can be allowed.

Curve (d) in Figure 3.3.22 shows the measured response when the transistors in the second transformer in each negative peak section are replaced by compound pairs. The main effect is to improve the Q of the resonant parts of section 'Q' from a value of 14.3 and section 'S' from 22.3 to values which are approximately double. However h_{13} and h_{24} the coupling terms between input and output are made worse so that the Qs are degraded due to the fact that the transformation ratio has an asymmetric real component in it. The passband is improved by the increased Q but the negative stopband is made worse by the effect of h_{13} and h_{24}.

Linearity and harmonic distortion was measured; 2nd harmonic distortion being 50dB down at the maximum signal level the filter can handle before overload, this level at the output being - 40dBm (600 ohms). Third harmonic distortion is more than 70dB down at the same level. Distortion varies with frequency being worst near the cut-offs due to voltage magnification occurring near the poles and zeros of the transfer function. The linearity curve was found to depart from a straight line at the point where the first section inverter became non linear suggesting that an improvement could be achieved by increasing the dynamic range available at each transformer inverter.

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Power dissipation of the filter was 540mW using a single 30 volt supply rail. If common supplies were used for the current sinks, the resistors of the negative peak sections were taken to suitable bias voltages and the inverting amplifiers run at 0.1mA with reduced voltages to keep the load resistors small a total consumption of about 75mW would be possible together with a reduction in component count.

Building this model has proved that such filters can be made to work and that providing care is taken in the design stopbands of up to 50dB can be produced. Beyond that level the precision required from the active elements is difficult to provide. A solution using operational amplifiers might provide the accuracy and stability required but no such ready made economic answer has been found to date. Some of the problems found might be alleviated with a four phase design which being completely symmetrical would not need inverters.

3.4 Summary of Lossless Filter Methods

New classes of network elements, enabling the polyphase realization of constant reactances, have been presented. The constant reactances go together with coils and capacitors to make practical sequence asymmetric polyphase filters.

The network elements required have to be built using active devices and the practical design problems involved have been considered. Two principal techniques have been evaluated. The gyrator technique which replaces an N-phase set of constant reactance is best made using a network with a minimum constant reactances. The

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complex transformer technique tends to favour a minimum capacitor design for minimum complexity.

A complete five section 2 phase filter equivalent to a 10th order bandpass filter has been constructed and tested. The results indicate that such filters are a practical proposition but that considerable difficulties exist when trying to achieve stopbands in excess of 50dB.

The next chapter describes lossy passive polyphase filters which, with their natural physical symmetry make stopbands of up to 70dB possible with readily available standard components.

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