FREQUENCY SYNTHESIZER FOR GENERATION OF LOCAL OSCILLATOR SIGNALS.

DISSERTATION

Submitted in partial fulfilment of the requirements for the Degree of M.Sc (Engg.)

by

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DISSERTATION APPROVAL SHEET

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C O N T E N T S

CHAPTER 1

FREQUENCY SYNTHESIZER FOR GENERATION OF LOCAL OSCILLATOR SIGNALS REQUIRED FOR A FILTER BANK RECEIVER

Introduction

Frequency synthesizers are devices used to generate signals whose frequency can be set as desired. The present frequency synthesizer, intended to generate several local oscillator signals required for the 256 channel narrow band filter receiver under construction at the Raman Research Institute, Bangalore. The receiver is intended for the study of molecular lines emitted by several molecules in our Galaxy at millimetre wavelengths. For example, Carbon Monoxide has an emission line centred around 115.271 GHz and it is widely distributed in our Galaxy. The study of such phenomena is called molecular spectroscopy, in which two parameters mainly determine the quality of observations. First is the resolution and second is the total spectrum studied with such a resolution.

An ideal filter receiver should have the widest possible spectrum width and the narrowest possible spectral resolution. The required spectral resolution depends upon the width of the spectral line which we wish to study. At millimetre wavelengths for normal molecular lines, a spectral resolution of 250 KHz is considered sufficient. We need however a number of such narrowband channels to achieve a wide spectrum. In the present filter receiver we cover a 64 MHz wide spectrum with a resolution of 250 KHz per channel.

The mm-wave spectrum is first converted to an intermediate frequency spectrum, say 10-400 MHz, by mixing it with a local oscillator signal at the required frequency. This I.F. Spectrum is then amplified. In this I.F/band molecular spectral information is centred in the range 118 MHz by proper adjustment of the frequency of the first local oscillator. It is difficult to construct passive or active filters with 250 KHz bandwidth in this range of frequencies, because the loaded Q requirement of the filter will be very large. So the entire range is brought down in frequency to where passive filters may be used without any difficulty. The original spectral information can then be studied by either of the following methods: By arranging several filters such that their centre frequencies differ by one bandwidth, or by using several filters of the same centre frequency and using several local oscillators to get the required range of frequencies. The receiver is optimized by using both of these two methods.

The range of frequencies from 118 MHz to 182 MHz is brought down to two identical bands of 32 MHz bandwidth extending from 50 MHz to 82 MHz by beating with local oscillator frequencies of 200 MHz and 232 MHz. Again, each of these bands of frequencies is divided into four identical bands of 8 MHz bandwidth extending from 26 -34 MHz by mixing with frequencies of 108 MHz, 100 MHz, 92 MHz and 84 MHz. Further mixing with local oscillator frequencies of 17.5 MHz, 18.5 MHz, 19.5 MHz, 20.5 MHz, 21.5 MHz, 22.5 MHz, 24.5 MHz results in eight 1 MHz wide signals. These are then separated into four 250 MHz bands by passing through four bandpass filters whose centre frequencies are separated by 250 KHz.

The aim of the present Frequency Synthesizer is to produce the above local oscillator signals with sideband noise levels well below 40 dB. Here

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monic levels should be also below 40 dB. It is important that the second Harmonics of the generated frequencies do not fall into the I.F bands of the receiver. The output levels of the synthesizer should be at a level of +7 dBm.

Chapter 2 introduces the design philosophy of the synthesizer and gives the block diagram of the system.

Chapter 3 is devoted to the design of the components used.

Chapter 4 describes the Mixers and digital techniques used for frequency division. Component values and the performance of each unit are also discussed.

Chapter 5 gives the system Assembly, constructional details and performance of the system.

CHAPTER 2

DESIGN OF THE SYNTHESIZER SYSTEM

Introduction

In the design of the Frequency synthesizer, it is seen that the number of mixers used is a minimum and the mixer outputs are such that the loaded Q requirement of the filters used to separate its output frequencies does not exceed 10. It is important that frequencies are generated nowhere ? whose second harmonic falls into the I.F.band of the receiver. It is preferred to obtain the final frequencies by frequency translation.

The synthesizer has as input a 5 MHz standard source with a stability of 1 in 10^9. Frequencies from 17.5 MHz to 24.5 MHz should be obtained in steps of 1 MHz. The scheme is shown in Fig.2 and the operation is explained below.

Using a transistor frequency multiplier (x 4), 20 MHz is generated from the 5 MHz standard. 1 MHz is obtained by dividing 5 MHz by digital techniques. This 1 MHz is multiplied by a transistor multiplier (x 2) to get a 2 MHz signal. Further multiplication (x 2) of the 2 MHz signal yields a 4 MHz signal. 2.5 MHz is obtained by dividing (÷ 2) the 5 MHz signal.

The final frequencies are now produced by frequency translation. 20 MHz when mixed with 2.5 MHz gives us 22.5 MHz and 17.5 MHz. 17.5 MHz when mixed with 4 MHz gives us 21.5 MHz. Mixing 22.5 MHz and 21.5 MHz. with 2 MHz signals gives us 20.5 MHz, 24.5 MHz, 19.5 MHz and 23.5 MHz respectively. 18.5 MHz is obtained from 20.5 MHz by a separate mixer. We find that to obtain all these frequencies we require 5 Mixers.

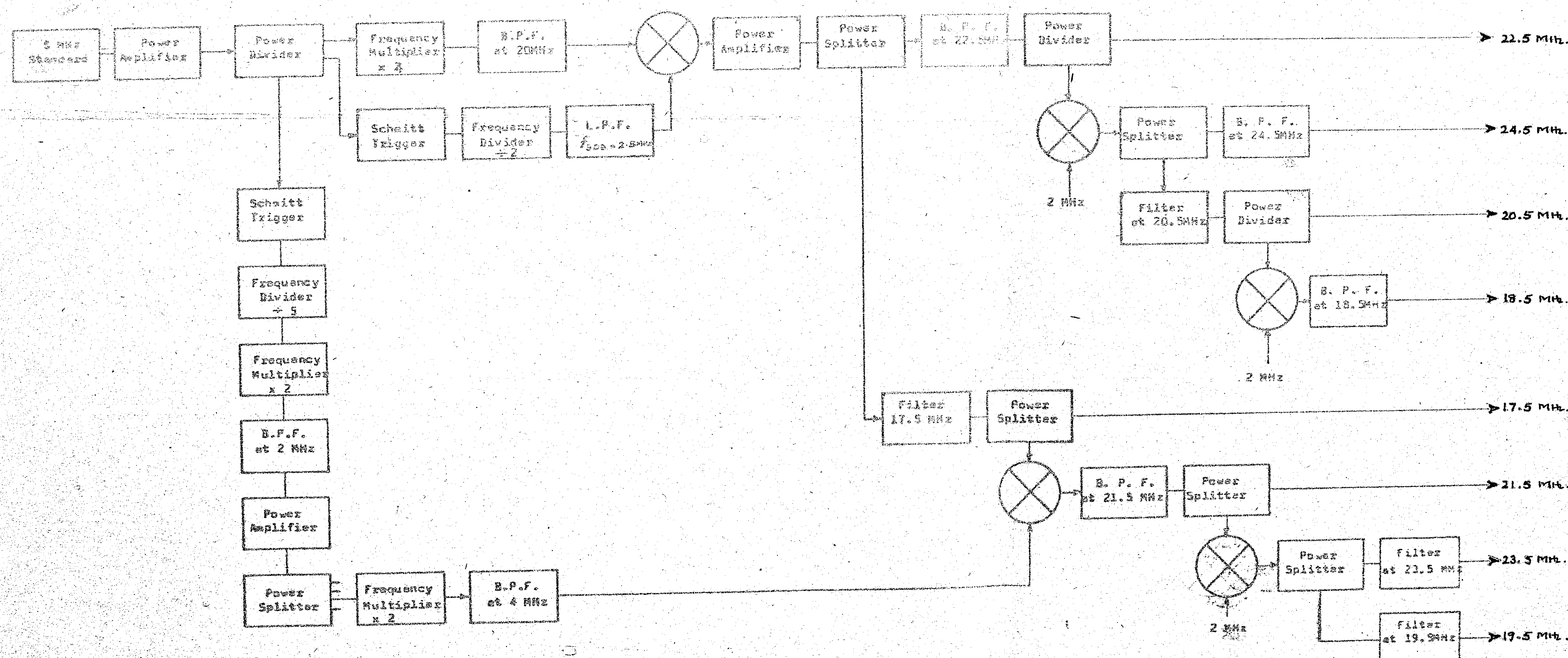

Fig. 2. Synthesizer block diagram.

Power dividers are used in place where it is necessary that power should be fed into separate branches. The power splitters ensures good isolation between the divided ports. Split Tee Power dividers, made by simulating transmission lines using lumped elements, are used in the synthesiser. T

Transistor frequency multipliers are nothing but Power amplifiers where the output is tuned to some harmonic of the input . It is operated with ~~beering~~ biasing under class C conditions because this mode has a good efficiency. As a rule of thumb the efficiency of a second harmonic multiplier is 42 percent, third harmonic 28 percent and fourth harmonic 21 percent. It was found experimentally that a transistor frequency multiplier tuned to its fourth harmonic produced enough output to work the R.F port of the Mixer SN 76514. Hence 20 MHz is directly obtained from 5 MHz through a single frequency multiplier. Two other transistor multipliers are tuned to their second harmonic to get 2 MHz and 4 MHz signals.

Frequency division is done by digital techniques. Divide by 5 and divide by 2 operation can be accomplished by using a single IC Chip SN 7490. Since such divisions are possible only with square waves, we require two schmitt triggers for the divide by 2 and divide by 5 operations. The SN 7413 chip contains dual NANDED Schmitt triggers. By shorting the inputs of the NAND gates conversion to square waves becomes possible. The divided square wave contains the required fundamental and its higher harmonics. A low pass filter is used to extract the sine wave from the square wave.

Bandpass filters must be capable of rejecting unwanted harmonics to a level of at least 40 dB. Modern filter design techniques provide Nomographs and tables to predetermine the nature of the response and the attenuation levels at the required frequency. The polynomial type of filters with coupled resonators are used. The complexity of the filter depends on the characteristics required and the Butterworth type of response has been chosen to characterise the passband.

The Mixer circuit used is the SN 76514 which is available in integrated form. The SN 76514 is a double balanced mixer which utilises two cross-coupled, differential transistor pairs driven by a third balanced pair. The circuit features a flat response over a wide band of frequencies (typically 100 MHz). Most of the circuitry is built into the chip. The only external components are the bypass capacitors to be chosen for optimum performance.

When assembling the components together it is desirable to ensure that the connected terminals are matched. This reduces reflections and minimizes power losses.

CHAPTER 3

DESIGN OF INDIVIDUAL COMPONENTS

3.1

In section 3.1, the types of filters used and their design is described. In section 3.2. the evolution of the split-tee power divider from a transmission line form is given with design equations for simulating a transmission line using inductors and capacitors. Section 3.3. gives the design requirement of the transistor frequency multiplier.

3.1 Ideal and Practical Lowpass Filters

In designing filters, it is convenient to normalise the element values. We begin by designing for a 1 ohm termination and for a cut off at 1 radian per second. This helps us to avoid dealing with numbers containing powers of 10 and 2π. Once the design is completed on this basis, the element values can be scaled in impedance and frequency to get practical component values.

The behaviour of an ideal low pass filter is shown in figure 3.1 This ideal response shows no attenuation at all from zero frequency to w=1 but infinite attenuation at cut off. Practical filters can be constructed to approximate this ideal in any of several different ways.

Figure 3.2 shows the Butterworth type of approximation. Here the passband is seen to be exceedingly flat near zero frequency and a very high attenuation is experienced at high frequency, but the approximation for both pass and stop bands is relatively poor near the cut off region.

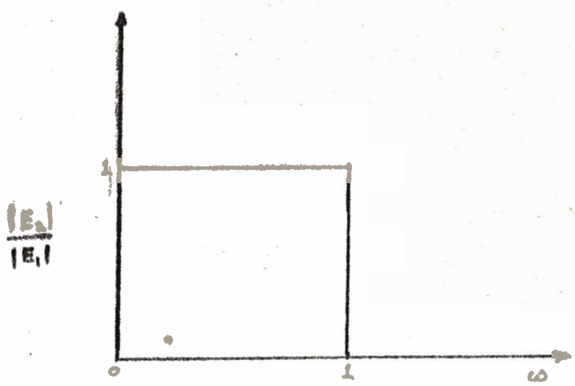

Fig.3.1. The ideal low pass filter.

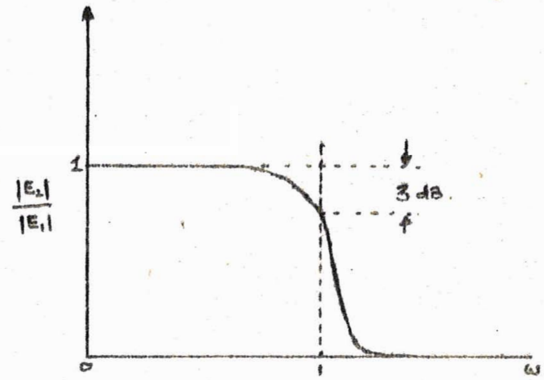

Fig.3.2. Butterworth approximation to ideal response.

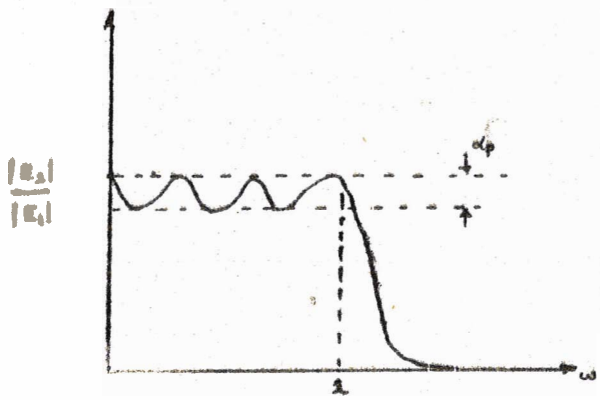

Fig. 3.3. Tchebycheff approximation.

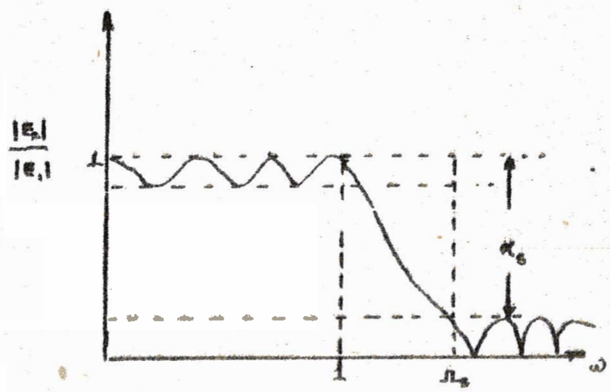

Fig.3.4. Elliptic-function response.

In figure 3.3 the Tchebycheff approximation has certain amount of ripple in the passband. Both the Butterworth types and the Tchebycheff types have zeros of their response only at infinite frequency. Given this condition it can be shown that the Tchebycheff type gives these steepest possible descent into the stopband other things being equal.

The performance can be improved with response zeros positioned near the passband. Figure 3.4 shows the elliptic function type of response, which is characterised by having equal ripples in the passband. This happens to be convenient, because the commonest stopband requirement, from the user's point of view, is that he shall be guaranteed atleast so much attenuation throughout the stopband. The elliptic function type meets this specification and is probably the most commonly used.

The elliptic and Tchebycheff types of filters, have a performance better than the Butterworth type as predicted theoretically. But the Butterworth type is less critical and can meet most of our requirements with fewer components than the other types.

In the simplification of the modern design approach, it is desirable that the element values be precalculated and normalised in a convenient manner and related to the desired filter characteristics. It is also desirable that these relationships be in catalog form, atleast for the types of filters most commonly used. Such tables are available in books on Filter Synthesis (please see reference No.1)

The following effective parameters determine the properties of a filter:-

1. The order of the transmission function which sets the number of reactive elements necessary for realisation.

2. Minimum attenuation in the stopband above a certain limiting frequency, A min.

3. The maximum amplitude of ripples in the passband, A max.

3.2 Estimate of Filter Complexity with Nomograph

Figure 3.5 shows the nomograph published by Kawakami for a Butterworth type of response. This is useful in determining the required degree of transmission function in order to satisfy a given condition. They avoid direct reference to any elliptic function parameter. The maximum value of ripple in the passband (A max.) is given at the right side of the monograph. A straight line is drawn from the value of A max permitted though the desired value of attenuation in the stopband (A min.). The line runs upto the third vertical line and is then rotated to run parallel to the scale. The desired amount of attenuation at a given frequency will be guaranteed if the filtering function is of the order found at the intersection of the vertical line erected from the Ω scale value and the line which runs parallel to the Ω scale. If the crossing appears between two curves, the order that must be chosen is the largest integer written above the curve.

A majority of modern high frequency filters can be designed with the aid of tables of normalised parameters. Normalised values included in the

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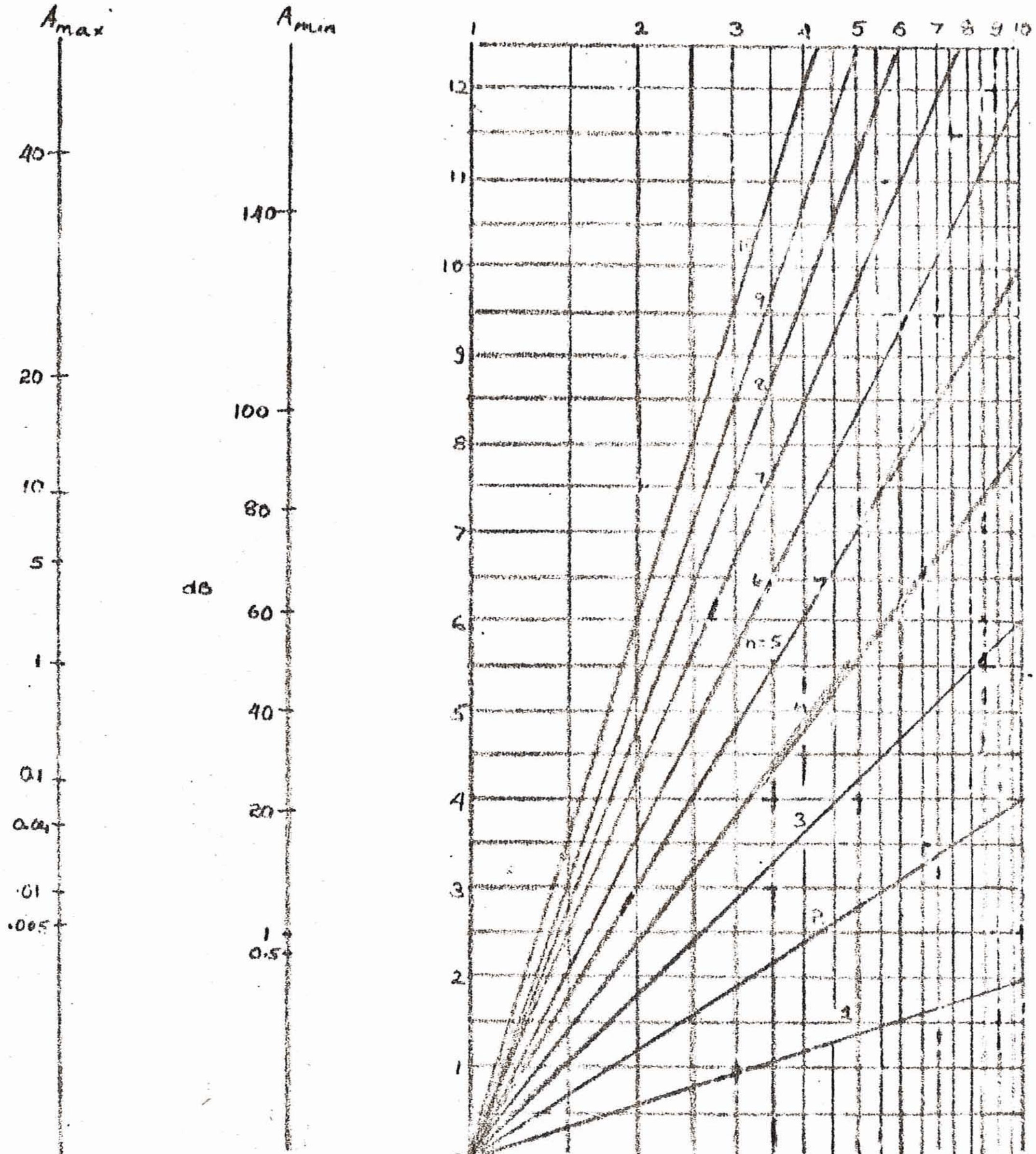


Nomograph for Butterworth filters

tables are dimensionless and designed as follows:-

1. Normalised frequency $\Omega = \frac{f}{f_r}$

2. Normalised resistance $r = \frac{R}{R_r}$

3. Normalised inductance $L' = \frac{\omega_r L}{R_r}$

4. Normalised capacitance $C' = \omega_r C R_r$

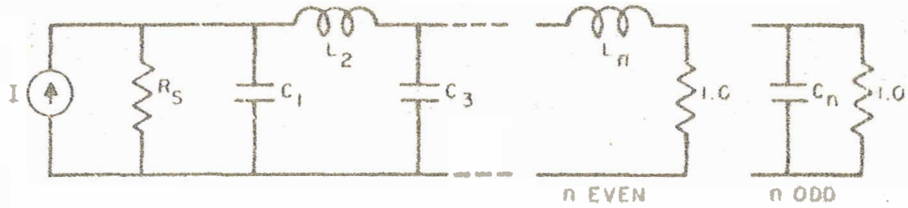
Where Rr and Wr are the reference impedance level and reference frequency respectively.

The normalised cut off frequency is $\Omega_c = 1$ for all low pass filters. Table 3.1 presents the normalised element values for polynomial filters for the Butterworth type of response. The element values listed in the tables correspond to the components of the basic low pass model shown above and below it. The filter shown below the table is the dual of the filter above. All values have been calculated to give a unity 3 db bandwidth.

tables set the load, or output resistance as equal to unity and allow the input or source resistance (Rs) to vary from one to infinity. In the case of n even, R_S increases from unity and for n odd, Rs decreases. When using the tables, it should be noted that the schematic at the top of the page corresponds to the top column headings. This schematic always has a π type input and can be used if desired, with open circuit input resistance corresponding to an ideal current source. On the other hand, if the lower schematic is used, the column headings at the bottom of the tables are to be used. In this case, the input is always of the T type and it should be noted that for this type of schematic it is permissable to set $R_S = 0$, corresponding to the short circuit operating condition of an ideal voltage source.

LOW PASS ELEMENT VALUES

BUTTERWORTH RESPONSE



n	R_s	C_1	L_2	C_3	L_4	C_5	L_6	C_7
5	1.0000	0.6180	1.6180	2.0000	1.6180	0.6180		
	0.9000	0.4416	1.0265	1.9095	1.7562	1.3887		
	0.8000	0.4698	0.8660	2.0605	1.5443	1.7380		
	0.7000	0.5173	0.7313	2.2849	1.3326	2.1083		
	0.6000	0.5860	0.6094	2.5998	1.1255	2.5524		
	0.5000	0.6857	0.4955	3.0510	0.9237	3.1331		
	0.4000	0.8378	0.3877	3.7357	0.7274	3.9648		
	0.3000	1.0937	0.2848	4.8835	0.5367	5.3073		
	0.2000	1.6077	0.1861	7.1849	0.3518	7.9345		
	0.1000	3.1522	0.0912	14.0945	0.1727	15.7103		
	INF.	1.5451	1.6944	1.3820	0.8944	0.3090		
6	1.0000	0.5176	1.4142	1.9319	1.9319	1.4142	0.5176	
	1.1111	0.2890	1.0403	1.3217	2.0539	1.7443	1.3347	
	1.2500	0.2445	1.1163	1.1257	2.2389	1.5498	1.6881	
	1.4286	0.2072	1.2363	0.9567	2.4991	1.3464	2.0618	
	1.6667	0.1732	1.4071	0.8011	2.8580	1.1431	2.5092	
	2.0000	0.1412	1.6531	0.6542	3.3687	0.9423	3.0938	
	2.5000	0.1109	2.0275	0.5139	4.1408	0.7450	3.9305	
	3.3333	0.0816	2.6559	0.3788	5.4325	0.5517	5.2804	
	5.0000	0.0535	3.9170	0.2484	8.0201	0.3628	7.9216	
	10.0000	0.0263	7.7053	0.1222	15.7855	0.1788	15.7375	
	INF.	1.5529	1.7593	1.5529	1.2016	0.7579	0.2588	
7	1.0000	0.4450	1.2470	1.8019	2.0000	1.8019	1.2470	0.4450
	0.9000	0.2985	0.7111	1.4043	1.4891	2.1249	1.7268	1.2961
	0.8000	0.3215	0.6057	1.5174	1.2777	2.3338	1.5461	1.6520
	0.7000	0.3571	0.5154	1.6883	1.0910	2.6177	1.3598	2.0277
	0.6000	0.4075	0.4322	1.9284	0.9170	3.0050	1.1503	2.4771
	0.5000	0.4799	0.3536	2.2726	0.7512	3.5532	0.9513	3.0640
	0.4000	0.5899	0.2782	2.7950	0.5917	4.3799	0.7542	3.9037
	0.3000	0.7745	0.2055	3.6706	0.4373	5.7612	0.5600	5.2583
	0.2000	1.1448	0.1350	5.4267	0.2874	8.5263	0.3692	7.9079
	0.1000	2.2571	0.0665	10.7004	0.1417	16.8222	0.1823	15.7480
	INF.	1.5576	1.7988	1.6588	1.3972	1.0550	0.6560	0.2225
n	$1/R_s$	L_1	C_2	L_3	C_4	L_5	C_6	L_7

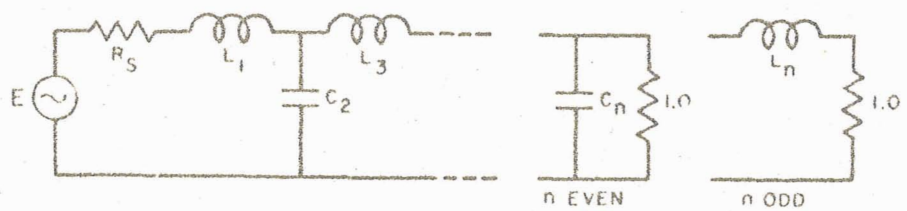


Table 3.1

3.3Low pass design example

The synthesizer designed requires a low pass filter with cut off at 2.5 MHz. Let us choose the Butterworth type of response and let the terminating impedance be 50 ohms. The filter must pass frequencies from 0 to 2.5 MHz, the maximum attenuation at 2.5 MHz being 3 db. Not less than 30db attenuation is to be provided at 5 MHz and above.

From the Nomogram by Kawakami (figure 3.5), it can be determined that the filter must be between fourth and fifth order. The fifth order is chosen. The desired information is given in the tables. We are using the schematic at the top of the page, having a π input since this schematic contains 2 inductors and 3 capacitors.

The table gives only the normalised element values. It should be noted that the element values are symmetrical about the centre of the filter, constituting a peculiarity of equally terminated butterworth and Tchebycheff filters. This factor makes the equally terminated case certainly, the easiest and most popular for most practical applications.

The filter may be denormalized by the usual impedance and frequency scaling. A reference inductor is defined as

$$Lr = \frac{Rr}{Wr}$$

Likewise a reference capacitor is defined as

$$Cr = \frac{1}{Rr\ Wr}$$

Since Rr = 50 ohms and $Wr = 2 \times 2.5 \times 10^6$ rad/sec

$$Lr = 3.1831\ /\mu H$$

$$Cr = 1273.24\ pf.$$

Multiplying the reference values by the normalised element values denormalises the filter and gives the actual element values. The actual element values are given in chapter 4.

3.4 Bandpass Filter Design

A bandpass filter can be obtained from its low pass prototype which will have a geometric symmetry independent of bandwidth. This although theoretically correct, is not always justified practically. The element values may be too small or too large.

Let us assume that we have to adopt a transformation for a third order low pass filter to a bandpass filter by conventional techniques. This requires three coils and three capacitors. Almost the same kind of performance can be expected from three parallel resonators coupled together by mutual inductance by capacitance by inductance or by a combination of these.

Neither image parameter theory nor synthesis theory requires a specific concept of coupling in order to design or explain the physical operation of filters. Half section and full sections can be connected if they have appropriate characteristic impedance. Filter design in terms of coupling has become very popular with electronic engineers where narrow bandpass filters are concerned.

Coupled resonators

There are two types of coupling which are generally used in coupled resonator filters. The first consists of reactive parallel elements which are coupled by single lumped or distributed reactance. The second consists of coupling by mutual ~~reductances~~ inductances. Each stage may have a different type of coupling mechanism. This cascade is then terminated at one or both ends in a pure resistance.

Co-efficient of coupling

From transformer theory, the magnetic coupling co-efficient is given as $K = \frac{M}{\sqrt{LpLs}}$. By analogy with this we can define a co-efficient of coupling to the bandpass filter also. The normalised coupling K, is defined as $K = \frac{Y}{\sqrt{g1g2}}$. Y is the coupling admitance and g1 and g2 are the input and output short circuit admitances of the filter at resonance.

We normalise the quality factor Q as q where q is given as $q = \frac{\Delta f}{f_m} Q$ where f_m is the centre frequency and Δf is the bandwidth.

Choice of coupling

When filter design is performed in terms of coupling co-efficient K and normalised q of the circuit elements, after setting the total nodal capacitance of each node, it is necessary to decide what kind of coupling mechanism can be used between adjacent nodes. The coupling can be either capacitive, inductive, mutual magnetic or a combination of these. No firm rules exist dictating which type of coupling to choose.

Using the design data, we design the network in steps. The tables give the coupling co-efficient K and a simple calculation transforms them into capacitances or inductances necessary to produce the required coupling co-efficients.

The proximity of ground planes can appreciably affect the value of the true capacitance of a capacitor and will also strongly affect the value of self-inductance of an inductor. Therefore, particularly when small element values are called for, the actual value of coupling capacitors and inductors to be used in the filter cannot always be determined, since the actual values of these components could change because of proximity effects. Because of this, adjustable elements are used as coupling elements.

The minimum number of resonators required to satisfy a certain requirement may be obtained from conventional nomographs or curves. After the number of resonators and the coupling mechanism is determined the components can be found. To obtain the first resonator, the Q factor of the first node is selected. This means that the source resistance has determined the first node inductance and capacitance. For a very small relative bandwidth the ratio of coupling to shunt elements is approximately reciprocal to the relative bandwidth and is very small. Physically, this means that capacitor coupling is not always realisable and one must sometimes employ a mutual inductive coupling. The remainder of the circuit element are determined by a step by step design procedure with known values of loaded q and k. To obtain the results predicted by the theory, the unloaded q of each element must be greater than a certain minimum.

The choice of impedance level can be based on the desire to obtain maximum unloaded Q and some mechanical and economical considerations.

<u>Design with table of pre-distorted q and k parameters.</u>

When each of the filter elements has the same finite Q, then the real part of each root can be pre-distorted an amount equal to the normalised unloaded decrement of each element.

3-db DOWN k AND q VALUES

BUTTERWORTH RESPONSE

n	q_0	I. L.	q_1	q_n	k_{12}	k_{23}	k_{34}	k_{45}
2	INF.	0.000	1.4142	1.4142	0.7071			
	14.142	0.915	1.4142	1.4142	0.7071			
	7.071	1.938	1.4142	1.4142	0.7071			
	4.714	3.098	1.4142	1.4142	0.7071			
	3.536	4.437	1.4142	1.4142	0.7071			
	2.828	6.021	1.4142	1.4142	0.7071			
	2.357	7.959	1.4142	1.4142	0.7071			
	2.020	10.458	1.4142	1.4142	0.7071			
3	INF.	0.000	1.0000	1.0000	0.7071	0.7071		
	20.000	0.958	0.8041	1.4156	0.7687	0.6582		
	10.000	2.052	0.8007	1.5359	0.7388	0.6716		
	6.667	3.300	0.8087	1.6301	0.7005	0.6879		
	5.000	4.742	0.8226	1.7115	0.6567	0.7060		
	4.000	6.443	0.8406	1.7844	0.6077	0.7256		
	3.333	8.512	0.8625	1.8497	0.5524	0.7470		
	2.857	11.157	0.8884	1.9068	0.4883	0.7706		
4	INF.	0.000	0.7654	0.7654	0.8409	0.5412	0.8409	
	26.131	1.002	0.5376	1.4782	1.0927	0.5668	0.6670	
	13.066	2.162	0.5355	1.6875	1.0745	0.5546	0.6805	
	8.710	3.489	0.5417	1.8605	1.0411	0.5373	0.6992	
	6.533	5.020	0.5521	2.0170	1.0004	0.5161	0.7207	
	5.226	6.822	0.5656	2.1621	0.9547	0.4906	0.7444	
	4.355	9.003	0.5819	2.2961	0.9051	0.4592	0.7706	
	3.733	11.772	0.6012	2.4159	0.8518	0.4192	0.7998	
5	INF.	0,000	0,6180	0,6180	1,0000	0,5559	0,5559	1,0000
	32.361	1.045	0.4001	1.5527	1.4542	0.6946	0.5285	0.6750
	32.361	1.045	0.5662	0.7261	1.0947	0.5636	0.5800	0.8106
	16.180	2.263	0.3990	1.8372	1.4414	0.6886	0.5200	0.6874
	16.180	2.263	0.5777	0.7577	1.0711	0.5408	0.6160	0.7452
	10.787	3.657	0.4036	2.0825	1.4088	0.6750	0.5080	0.7066
	10.787	3.657	0.5927	0.7869	1.0408	0.5144	0.6520	0.6860
	8.090	5.265	0.4111	2.3118	1.3670	0.6576	0.4927	0.7290
	8.090	5.265	0.6100	0.8157	1.0075	0.4844	0.6887	0.6278
	6.472	7.151	0.4206	2.5307	1.3195	0.6374	0.4732	0.7542
	6.472	7.151	0.6293	0.8449	0.9722	0.4501	0.7267	0.5681
	5.393	9.425	0.4321	2.7375	1.2675	0.6148	0.4479	0.7821
	5.393	9.425	0.6508	0.8748	0.9355	0.4103	0.7663	0.5048

It is a difficult problem to synthesise the pre-distorted network if all the reactive elements, do not have the same Q. Hence the excellent Q of the capacitance must be degraded with the parallel resistance in order to make their Q the same as that of the last inductances. In case where accuracy is not our concern, we can use a value of Q which twice as high as that of inductances and then in the final network the capacitances are left as they are that is their Q is not reduced artificially.

Table 3.2 presents in tabular form k and q values for polynomial filters. The filters, for each type and order of filter given, tabulates the k and q values for various values of q starting from q_o equal to infinity (the lossless case) and then decreasing to a certain finite value. The lossless case, in practice is used where the quality factor of the components of the filter is high enough so that an insertion loss of 1db or less is expected. In many cases of filter design, however, the available elements will not have high enough quality and the tabulated values of pre-distorted k and q values will here be extremely useful. Note that for each case of finite q, the expected insertion loss of the filter is given. Linear interpolation between values of q_o is usually accurate enough for most practical design problems.

In some cases, two or more solutions are given for a particular requirement. The choice of which solution to pick is usually decided by the comparative values of q_1 to q_n and to the range of R values within each solution.

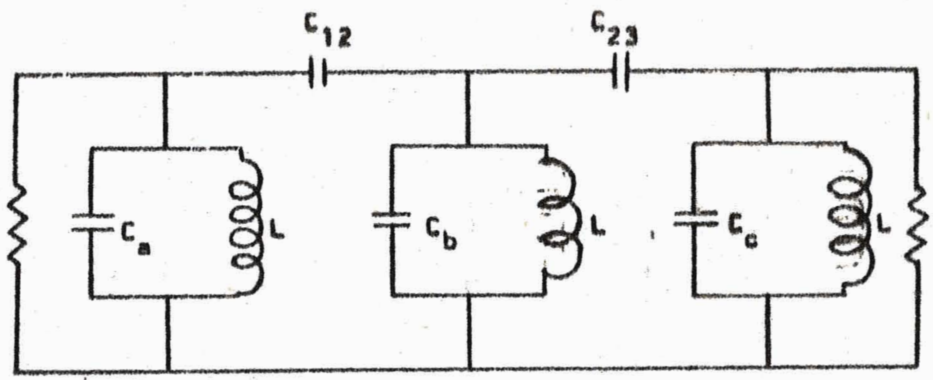


Fig.3.6. Three-pole bandpass filter.

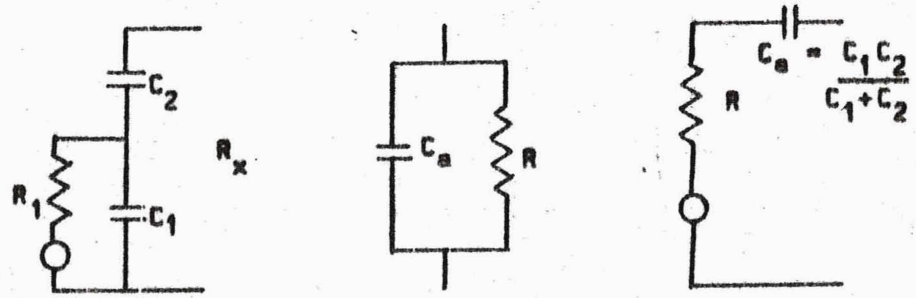


Fig.3.7. Transformation of impedence.

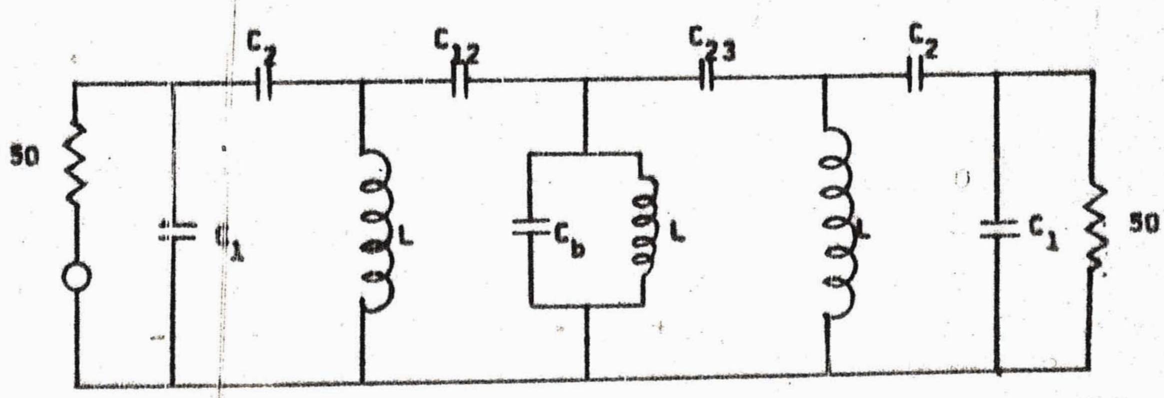


Fig.3.8. Final Circuit of the Completed Design.

3.5 Example of Butterworth Bandpass Filter Design

Let us design a particular filter used in the synthesizer (say filter at 20.5 MHz.

1. Input and output impedance = 50
2. BW 1 db = 1
3. BW 40 db = 8 MHz
4. Centre frequency = 20.5 MHz

Solution

1. Response form factor $\dfrac{bw40db}{bw1db} = 8$

2. Form the nomograph for Butterworth filters, using A max = 1 db, A min = 30 db and to at $\Omega = 8$ the required number of resonators falls between two and three. Therefore we must choose n = 3.

3. With bandwidth 1 db known and by using the curve of the attenuation characteristics for Butterworth filters the value of bw3db can be easily obtained.

$$\frac{bw1db}{bw3db} = 0.84$$

Since bw1db = 1 MHz then Bw3db = 1.25 MHz. From this same curve, the stopband rejection can be checked.

At bw40db/bw3db $= \dfrac{8}{1.26} = 6.4$ for n = 3 the attenuation will be approximately 50 db.

4. The k and q values are found in the table for the Butterworth response table (table 3.2). The infinite q case can be used if it is possible to obtain elements with a high enough Q factor such that the insertion loss of the filter will be 1 db or less. In order to obtain a

4.7 db insertion loss, the value of q must be 5, as read from the table. This value of Qo is chosen because the inductors available had a Q of only 75. The minimum unloaded Q of the internal resonators must be:

$$Q\ min = \frac{qof_m}{bw3db} = \frac{5x20.5}{1.25} = 82$$

The following parameters are read from the table.

$q1 = 0.8226 \qquad K12 = 0.6567$

$q3 = 1.7115 \qquad K23 = 0.7060$

$Q1 = \frac{fm}{\Delta f} q_1 \qquad q1 = 13.49$

$Q3 = \frac{fm}{\Delta f} q_3 \qquad q3 = 28.07$

5. We now make the choice of coupling between resonators. Let us choose a capacitive type of coupling. The required Q of the first resonator may be obtained by choosing the nodal inductance and capacitance (in the case of capacitive coupling) of such a value that the generator resistance produces the required Q1. The alternate method is to use a transforming circuit to couple a non-resonant generator to the first node. For most applications this technique is advisable because it allows to choose values of L and C that are realised in practice. The inductor value is chosen so that the value of coupling capacitor is around 3 to 4pf. This makes it easier to align. An inductor value of 0.6 /μh gives a nodal capacitor value of 101.3 pf. The same value will be used for all the coils which by the same token, will require a nodal capacity of 101.3 pf in all cases. The capacitive coupling elements and the shunt capacitors can be calculated in the following steps.

$$C12 = K12 \frac{\Delta f3db}{fo} \sqrt{C_I C_{II}}$$

$$\sqrt{C_I C_{II}} = \sqrt{C_{II} C_{III}} = 101.3 \text{ pf}$$

$$C12 = .6567 \times \frac{1.25}{20.5} \times 101.3 \text{ pf} = 4.1 \text{ pf}$$

The first nodal capacitor was set as 101.3 pf. So

$$Ca = C_I - C12 = 97.3 \text{ pf}$$

Similarly $C23 = K23 \frac{\Delta f3db}{fo} \sqrt{C_{II} \quad C_{III}} = 4.4$ pf

$\therefore$ Cb = 101.32 - C12 - C23 = 92,9 pf

Cc = 97.3 pf.

The coupling capacitors are however chosen as variable ones in the practical circuits for reasons already discussed.

6. Required load

The required load and source resistances are found from Q1, Q3 and Qo. First the total resistance, including the parasitic loss resistance is found:

R1 = wm LQ1 = 1033. ohms

R3 = wm LQ3 = 2150.9 ohms.

The parasitic resistance due to finite Qo is

Rp = wm LQo = 5747 ohms

Hence the load and source resistances are

$$Rs = \frac{R1 \ Ro}{Rp-R1} \quad 1260.4 \text{ ohms}$$

RL = 3437.4 ohms.

The circuit for the above calculated filter is shown in figure 3.6

7. As the final step of our design, the generator impedance must be transformed to produce the required quality factor in the first resonator. The same transformation must be performed with the last resonator also. The most practical transformation for the above case is shown in figure (3.7), because the shunt capacitor may be used to absorb the distributed capacitance, usually associated with the input and output circuits. The transforming circuit can be calculated in the following fashion

$$\frac{C1}{C2} = \sqrt{\frac{R_x}{R_1}} - 1 \qquad C_1 = C_a\sqrt{\frac{R_x}{R_1}}$$

$$Cs = \frac{C1\,C2}{C1+C2} \qquad C_2 = \frac{C_a\sqrt{R_x/R_1}}{\sqrt{R_x/R_1} - 1}$$

Where Rx is the transformed value of R1 required to produce the specified Q. The account of incidental coil dissipation (neglecting capacitive losses) can be estimated in the following way

$$\frac{1}{R} = \frac{1}{Rx} + \frac{1}{Q1\ X1}$$

Using the above equations we get

C1 = 488.32 pf C2 = 121.45 pf

The set of element values obtained for a band pass filter together with response of typical filters are included in chapter 4.

3.6 Design of Power Splitters

In the frequency synthesizer unit it is necessary at various points to split the power from a single port into two or more ports. A large number of different types of power dividers with and without isolation between output ports, are used for various applications. We have used split-tee power dividers (please see ref No.5) which give both power division as well as isolation between the output ports. A good isolation between the two output ports is required so that reflections, if any, in either of the two ports will cause no problem to the other port.

The design of these power splitters stems from the theory of transmission lines at radio frequencies. Transmission lines in special form as quarter wave transformers can be simulated by lumped inductances and capacitors at UHF frequencies. Such a form of the power splitter is shown in figure 3.9

To ade match at port 1, it follows that

$$\frac{Z12 \cdot Z13}{Z12 + Z13} = Z_{in}$$

Where Z12 and Z13 are the input impedances looking with arms 2 and 3 from port 1.

If the output impedance are chosen, then $Z12 = Z13 = 2Z_{in}$

The expression for the input impedance of a dissipationless line may be written as

$$Z_{in} = R_o \left[\frac{Z_R / \tan(2\pi s/\lambda) + j R_o}{R_o / \tan(2\pi s/\lambda) + j Z_R} \right]$$

If the line is made a quarter wave long or $S = \frac{\lambda}{4}$, $Z_o = \sqrt{Z_{in} Z_R}$

Once the characteristic impedance of the line is calculated, the equivalent form of the line as shown in fig.3.9 can be obtained using the following equations for the values of L and C,

$WC = \tan \theta/2 \, / Z_o$ and $WL = Z_o \sin \theta$

θ being 90°, the required values can be calculated.

With this design, the voltages at port 2 and port 3 are equal. Hence, a resistor may be placed between two ports without causing any power dissi-

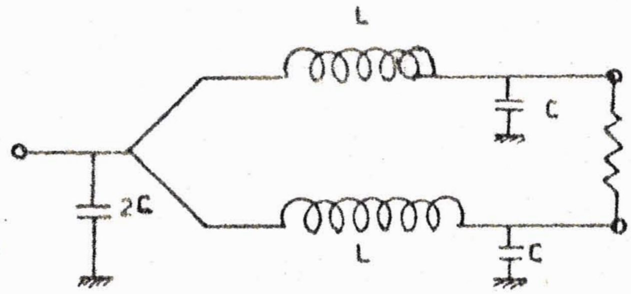


Fig.3.9. Power Splitter.

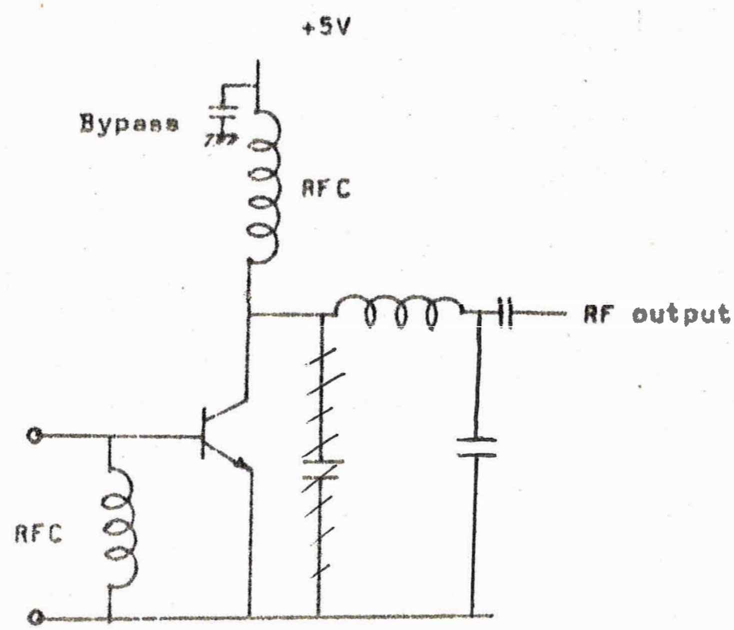


Fig.3.10. Frequency Multiplier

pation. If the power divider is fed from port 2 or port 3 energy will be dissipated in the resistor. Isolation between output ports and a good match seen looking in at any port is obtainable because of this resistor. Although the proper value of the isolation resistor has not been derived, it is observed that a value of $R = 2\ Z_{in}$ yields infinite isolation and a perfect match at the centre frequency.

3.7 Design example

As an example let us assume it is required to split power at 5 into two channels with 50 ohms impedance.

Since the ports of the power splitter are to be matched to 50 ohms, both the arms of the power divider should provide a Z11 and Z12 of 100 ohms. Therefore 50 ohms should be transformed to 100 ohms using the quarter wave line. Therefore, it is required that the line should be having a characteristic impedance of $Z_o = \sqrt{50 \times 100} = 70.7$ ohms

Since $\theta = 90^\circ$

$$C = \frac{1}{wZo} = 450.2 \text{ pf.}$$

$$L = \frac{Zo}{W} \quad 2.25\ /uH$$

In the synthesizer a three way power divider is required at 5 MHz, Therefore Z11 = Z12 = Z13 = 150 ohms provide a Z_{in} = 50 ohms. The characteristic impedance of the transmission line $Zo = \sqrt{50 \times 150} = 86.6$ ohms. In this case the calculated values are C = 367.5 pf and L = 2.7 /uH.

In the practical set up we require a typical isolation of about 40 db between the output ports when the input is terminated by a 50 ohms termination. Tunable coils may be provided for the power splitter to

..16

ensure such a condition. Insertion loss at the power divider is negligible. The only loss is a 3 db one owing to the splitting of power in each arm.

Alternately, it is also possible to build power dividers using resistive elements. This provides operation over a wide frequency range unlike the above power splitter where operation is constrained to a narrow frequency range. But resistive elements dissipate power and contribute to insertion loss. Hence such power dividers were not used.

3.8 Frequency multipliers.

In the initial stages of frequency synthesis, frequencies are stepped up using frequency multipliers. Transistor frequency multipliers are used which when operated in the class B mode produce harmonics along with the fundamental at the output. The desired frequency can be tuned out using a tank circuit. The synthesiser makes use of three such multipliers one of them producing the fourth harmonic of the input while the other two are tuned to the second harmonic.

Transistor frequency multipliers operate on the same principle as power amplifiers. However, in a multiplier circuit, the output must be tuned to a multiple of the output. A multiplier may or may not provide amplification (Please see reference No3 and 4)

A typical circuit of the frequency multiplier is shown in fig.3.10. The transistor remains cut off until a signal is applied. Therefore, the transistor is never conducting for more than 180° (half cycle) of the

360° input signal cycle. In practice, the transistors conduct for about 140° of the input cycle, either on the positive half or negative half, depending on the transistor type. No bias, as such, is required for this class of operation.

It should be noted that emitter is directly grounded. In high frequency applications it is always desirable to have such a direct connection. If the emitter were connected to ground through a resistance, an inductive or capacitive reactance could develop at high frequencies, resulting in desired changes in the network.

An RF choke (RFC) is used to provide RF signal isolation between base and emitter. This also provides a d-c return for the base. The collector is connected to the supply voltage through an RFC. No current limiting resistors are used and it should be noted that the supply voltage is well within the collector breakdown voltage of the transistor. The RFC provides DC return, but RF signal isolation between collector and power supply.

The RF chokes should have as small a DC resistance as possible. The current capability of the coil at low power applications is not of much concern. The minimum current capacity should be greater by atleast 10 percent than the maximum anticipated direct current. The inductance it should present depends on the operating frequency. An inductance which will produce a reactance of about 2000 ohms at the operating frequency is ferred.

The power supply circuit must be bypassed. A bypass capacitor of 0.1 /µf serves the purpose. If RF signals are present on the power supply line, the bypass capacitance or the RFC is not sufficient. Sometimes the presence of RF signals may be due to inadequate shielding. To reduce the RF on the supply side, the capacitor value and inductance value of the RFC should be increased.

Typically, the efficiency of a second harmonic multiplier is about 40 percent. The efficiencies of third and fourth harmonics are around 28 percent and 21 percent respectively. Hence, for example if an amplifier tuned to its second harmonic is to produce a 5 mW RF output, the dc input required is approximately 5/4 = 12.5 mw.

CHAPTER 4: Development and performance of individual components:

In this Chapter the Mixer and the integrated circuits used for frequency division are discussed. Component values and the performance of each block are also included.

4.1 Mixers:

As most frequency multipliers produce a certain amount of excess Phase noise, the design of the final stage of frequency, synthesis incorporates frequency translation using balanced mixers.

The frequencies fed into the Mixer should be chosen in such a manner that the resulting frequencies could be easily seperated by using passive band pass filters with a loaded Q that does not exceed 10. Hence in the design it is seen that the sum and difference of frequencies differ by at least by 4 MHz in the 17.5 MHz to 24.5 MHz range and by 8 MHz in the 100 MHz range. It is also seen that the second harmonic of the generated frequencies do not fall in the intermediate frequency bands of the receiver. The design is such that the number of mixer chips used is also a minimum.

Two mixer units were contemplated for use, one being available in integrated circuit form consisting of differential amplifiers ~~dirving~~ driving a dual differential amplifier and the other being the Mini circuit Laboratories' mixer consisting of transformers and diodes. The SN 76514 I.C. mixer was chosen for the synthesizer, because of its compactness, conversion gain, and flat response over the operating frequency.

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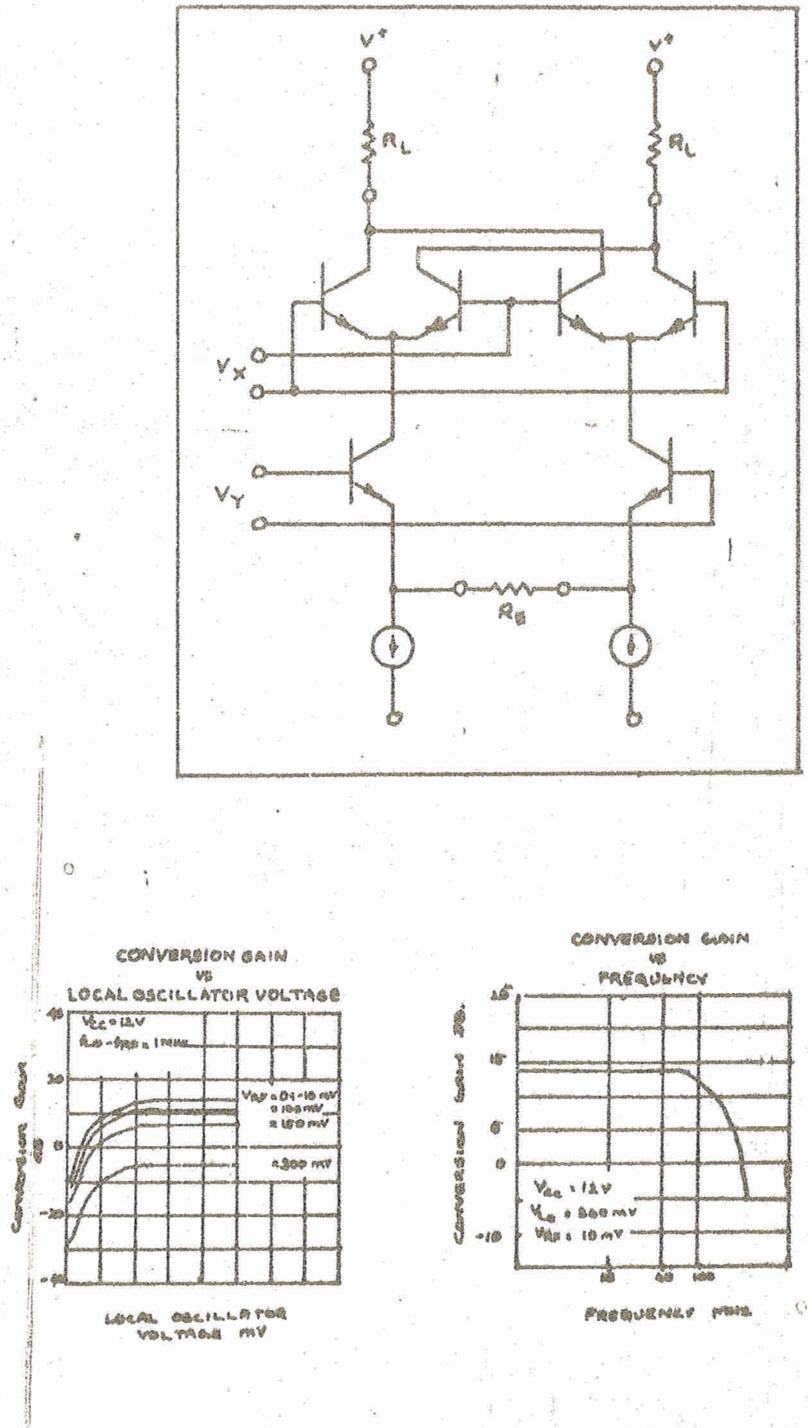

Fig. 4.1. Balanced Mixer.

In addition to its flat response to 100 MHz it had good isolation between the local oscillator, R.F and the I.F. ports.

As stated above the SN 76514 balanced Mixer consists of a differential amplifier driving a dual differential amplifier. In addition, the entire biasing circuit and the circuits for providing a constant current source for the lower differential amplifier are provided in the chip. The only external components required for the SN 76514 are the bypass capacitors shown in Fig.4 which are to be chosen for optimum performance.

The operation of the SN 76514 Mixer is based on the ability of the device to deliver an output which is proportional to the product of the voltages V_x and V_y. This holds true when the magnitudes of V_x and V_y are maintained within the limits of linear operation of the three differential amplifiers in the device. Expressed mathematically, the output (actually output current, which is converted to an output voltage by an external load resistance), V_o is given by:

$$V_o = K V_x V_y$$

where the constant K may be adjusted by a choice of external components. A detailed description of how the mixer circuit configuration performs the basic function of multiplication as expressed by the above equation is available in the literature.

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Device operation.

The most commonly used mode of operation of the I.C. mixer is that of applying a high level input signal to the dual differential amplifiers (carrier input port)and a low level input signal to the lower differential amplifier (modulating signal input port). This results in saturated switching operation of the carrier dual differential amplifier.

The resulting output signal contains only the sum and difference frequency components and the amplitude information of the modulating signal. This is the desired condition for the majority of applications of the SN 76514.

Saturated operation of the carrier-input dual differential amplifiers also generates harmonics (which may be predicted by Fourier Analysis). Reducing the carrier input amplitude to its linear range greatly reduces these harmonics in the output signal. However, it has the disadvantages of reducing the gain and causing the output signal to contain carrier signal amplitude variations.

The carrier input differential amplifiers have no emitter degeneration. Therefore, the carrier input levels for linear and saturated operations are calculated. The cross over point is in the vicinity of 30 mV, r.m.s. with linear operation below this level and saturated operation above it.

Since the SN 76514 generates an output signal consisting of the sum and difference frequencies of only the two input signals, can be used as a double balanced mixer. Figure 4.1 shows the range of operation of the SN 76514 and

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and the dependence of its conversion gain on frequency. With optimum levels of the input signals, the conversion gain can be as high 14 dB. Greater conversion gains can be achieved by using tuned circuits with impedence matching on the signal input port. The output impedence of the mixer being 600 ohms, matching networks have to be provided for use in the 50 ohm system.

Typical performance.

Conversion gain vs local oscillator voltage, and conversion gain vs frequency are given in figure 4. Typically, the SN 76514 has the following characteristics:

Flat response to 100 MHz		
Local Oscillator IF isolation	..	30 dB
Local Oscillator RF isolation	..	60 dB
R.F - I.F isolation	..	30 dB
Conversion gain	..	14 dB

4.2 Frequency division.

To obtain the final frequencies by frequency translation, we have to generate 2 MHz, 2.5 MHz and 4 MHz. Since the standard source available has frequency of 5 MHz, these frequencies can be obtained only by frequency division. Digital techniques are used for this purpose.

When a square wave is applied to a flip flop the output of the flip flop changes state only when the input makes a transition from a positive state to a negative state we get a square wave output divided by two in frequency. Using two or more flip flops in cascade, division by 4 or more is possible.

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By proper feedback from the output to any one of the inputs, division of square waves by any integer is possible.

To obtain 2.5 MHz from 5 MHz we require a flip flop for a divide by 2 operation. To obtain 2 MHz we first divide 5 MHz by 5 to get 1 MHz and by using transistor frequency multipliers we can obtain 2 MHz. Further fre - quency multiplication by 2 gives us 4 MHz.

Division by 2 and by 5 can be carried out using a single I.C. chip, the SN 7490.

The SN 7490 is a high-speed, monolithic decade counter consisting four master slave flip-flops internally interconnected to provide a divide-by-two counter and a divide-by-five counter. Since the SN 7490 responds only to a square wave input, the sine wave obtained from the standard source should be first converted to a square wave. We require two Schmitt triggers for this purpose. The SN 7413 has a dual NANDED Schmitt trigger designed primarily for logic operations with hysteresis, but if the input of the NAND gates are shorted, the SN 7413 provides an excellent means for sine wave to square wave conversion.

Having obtained the divided-by-two square wave, we require a low pass filter to extract the fundamental frequency (2.5 MHz) in it and to reject the harmonics (5 MHz and higher). We do not require an additional low pass filter to extract the 1 MHz sine wave from the 1 MHz square wave obtained by the divide-by-five operation. This is because we require only the 2 MHz signal as a sine wave, which can be obtained after the transistor frequency multiplier.

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Centre Frequency MHz	C'_1 pf	C'_2 pf	L_1 µH	C_{12} pf	L_2 µH	C_b pf	C_{23} pf	L_3 µH	C''_2 pf	C''_1 pf
17.5	428	148	.718	5.4	.718	104	5.8	.718	151	420
18.5	507	139	.65	.5	.65	114	5.2	.65	139	507
19.5	491	127	.63	4.4	.63	96.3	4.7	.63	115	796
20.5	543	118	.6	4.06	.59	93	4.3	.59	121	504
21.5	482	112	.6	3.5	.6	84	3.7	.6	112	482
22.5	520	86	.78	3	.81	67.8	3.2	.81	82	547
23.5	460	110	15	3.2	.5	85	3.4	.5	110	460
24.5	476	102	.48	2.9	.48	81.5	3.15	.48	102	476

Table 4.1 Band pass filter element values.

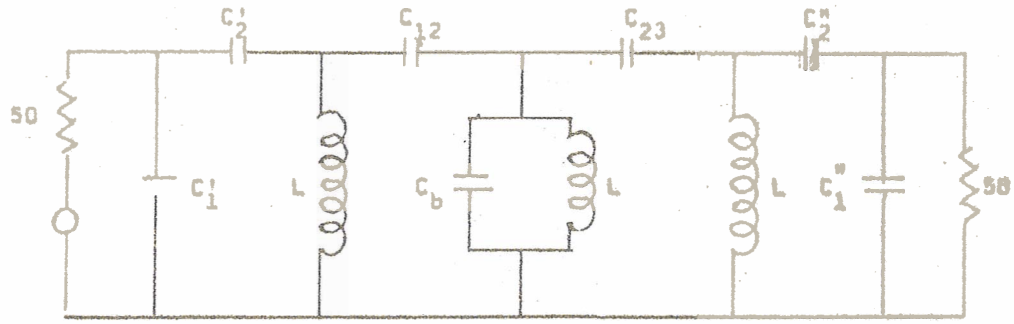

Fig.4.2. Band Pass Filter.

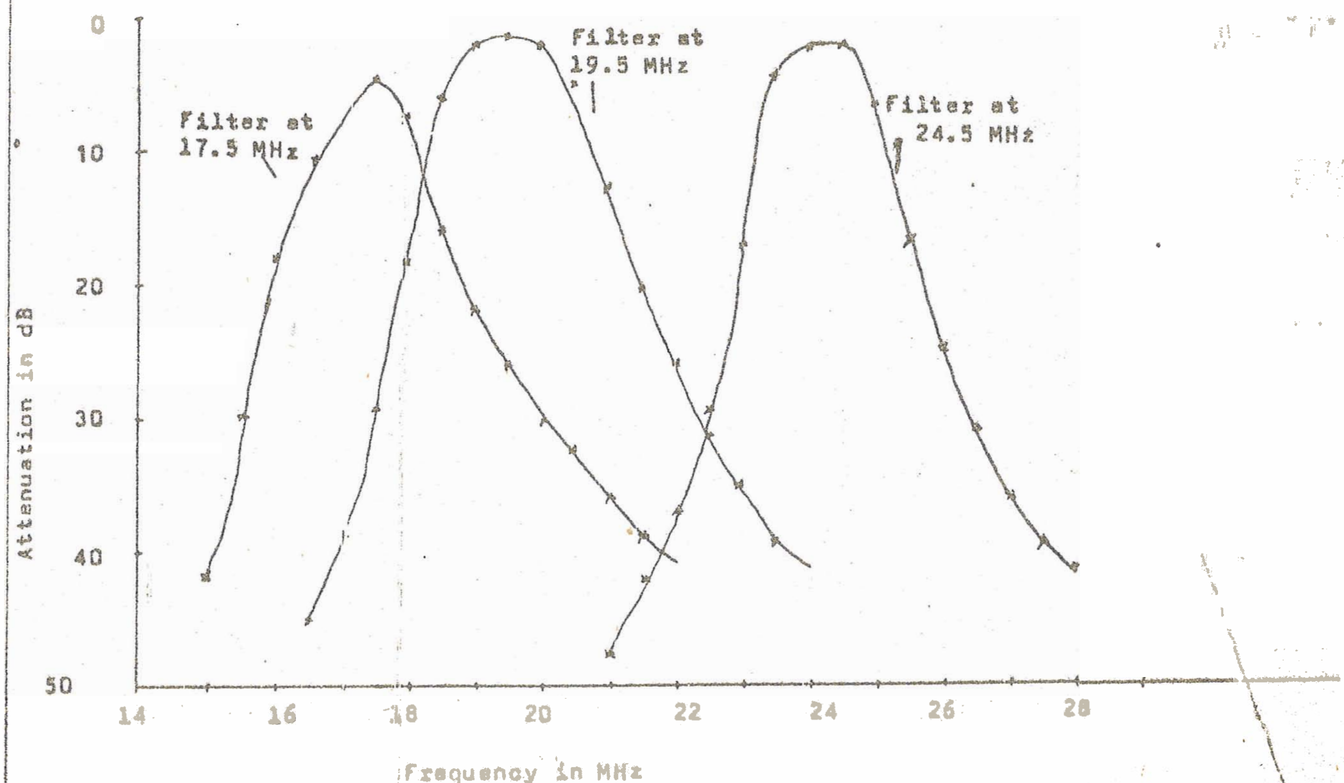

Fig.4.3. Typical response curves.

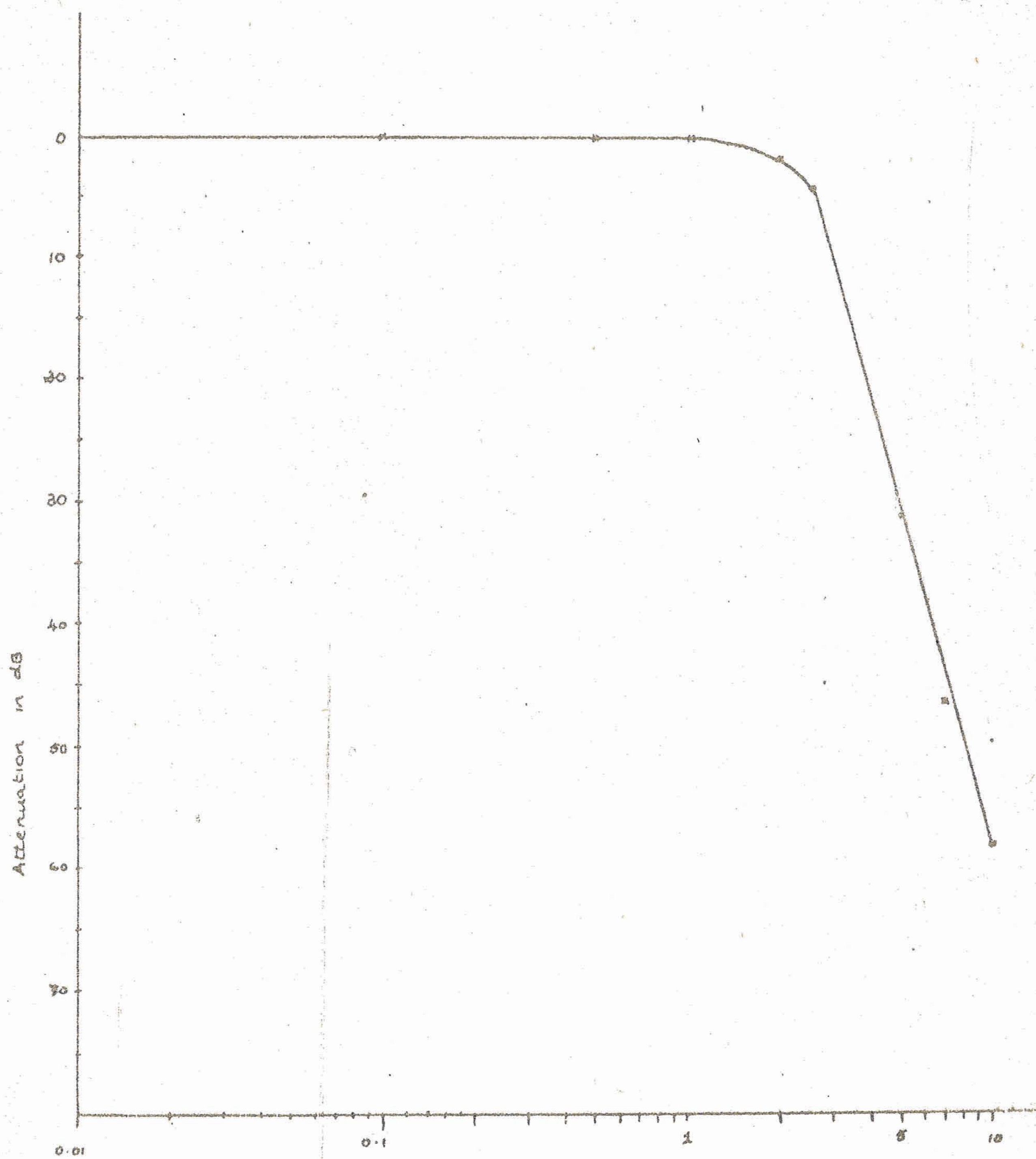

Fig. 4.4. Response of a Low pass filter.

4.3 Performance of the Band Pass filter.

The component values of the Band pass filters are given in Table 4.1. Typical response curves are given in figure 4.3. It is seen that the Band-pass filter provides more than 30 dB attenuation at 4 MHz bandwidth and more than 40 dB attenuation at 8 MHz bandwidth. Insertion loss is not more than 3 dB. The filters are matched to 50 ohms providing a return loss of not less than 22 dB at the input and output port. This corresponds to a standing wave ratio of 1.17.

Performance of the low pass filter.

The element values of the low pass filter are given below:

C_1 = 787 p_f L_2 = 5.2 uH C_3 = 2547 p_f L_4 = 5.2 uH C_4 = 787 p_f.

The performance of the low pass filter is plotted in figure (4.4). It Introduces an insertion loss of 4 dB at 2.5 MHz. The attenuation at 5 MHz is 31 dB. This value is acceptable as the 5 MHz frequency, being the second harmonic content of the square wave, will be already well below the 2.5 MHz level.

CHAPTER 5

ASSEMBLY OF THE SYSTEM & PERFORMANCE TEST

In this chapter, we discuss briefly, the assembly of the system and its performance. The performance of the system is analysed by feeding the synthesized frequency signal to a spectrum analyser and noting the levels of spurious signals and the phase noise near the carrier.

5.1 Assembly of ~~tie~~ the Frequency Dividers and Multipliers

The frequency dividing circuits which are dual in line integrated circuit packages and the frequency multiplier sections were assembled on a printed circuit board.

Additional circuitry had to be incorporated because the power levels at the input of the dividers and multipliers should be made compatible. For example, the power that should be fed to the input of the (x 4) frequency multiplier for optimum output is about +5dBm. Schmitt trigger operation also requires critical adjustment of the voltage and impedance levels as its input, and a matching circuit was provided so that both the schmitt trigger and the transistor multiplier functioned well with the same input power level.

5.2 Assembly of low pass filter and bandpass filters

The low pass filter was constructed on a printed board using Siemen's pot cores. These pot cores, even though they are not tunable, provide a good Q if Litz wire is used.

The bandpass filters on the other hand, were constructed in Module boxes. The Module boxes were made of brass sheets and provided good isolation from external R.F. Tunable coils were used in these filters.

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Matching the input and the output of Band pass filters is rather critical because of stray reactances. The tunable coils help in obtaining the exact centre frequency and also to provide good matching.

5.3 Assembly of frequency multipliers

The CIL 932 transistors which has a high transition frequency(typical value of 500 MHz) was used in the frequency multiplier circuits. Toroidal cores were used to function as RFC. Tunable coils were used to tune to the required harmonic.

The balanced Mixers were assembled on a separate printed circuit board.

5.4. Analysis of the output signal

One of the Frequencies, 22.5 MHz obtained from the Fräquency Synthesizer was analysed on a Tektronix spectrum Analyser. Fig. 5.1 shows the spectrogram with a resolution of 300 Hz and a sweep width of 1 KHz per division. The vertical scale is 10 dB per division. It is seen that the Phase noise is down by 30 dB from the carrier. Figure 5.2 shows the same spectrum with reduced resolution in order to study the levels of the harmonic spurious signals. The resolution is 3 KHz and the sweepwidth is 2 MHz per division. It is seen from the spectrogram that the spurious harmonic at 17.5 MHz is down by atleast 50 db. No trace of the 20 MHz components can be easily identified.

5.5 Conclusions

In this report we have presented the design, construction and performance of a frequency synthesizer system which is required to generate

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[illegible] oscillator signals for use in a filter bank spectral several local oscillator signals for use in a filter bank spectral line receiver. The frequencies generated range from 17.5 MHz to 24.5 MHz in steps of 1 MHz. Tests performed on the system showed that the phase noise near the carrier frequency is atleast down by 30 dB and the spurious harmonics are well below 50 dB.

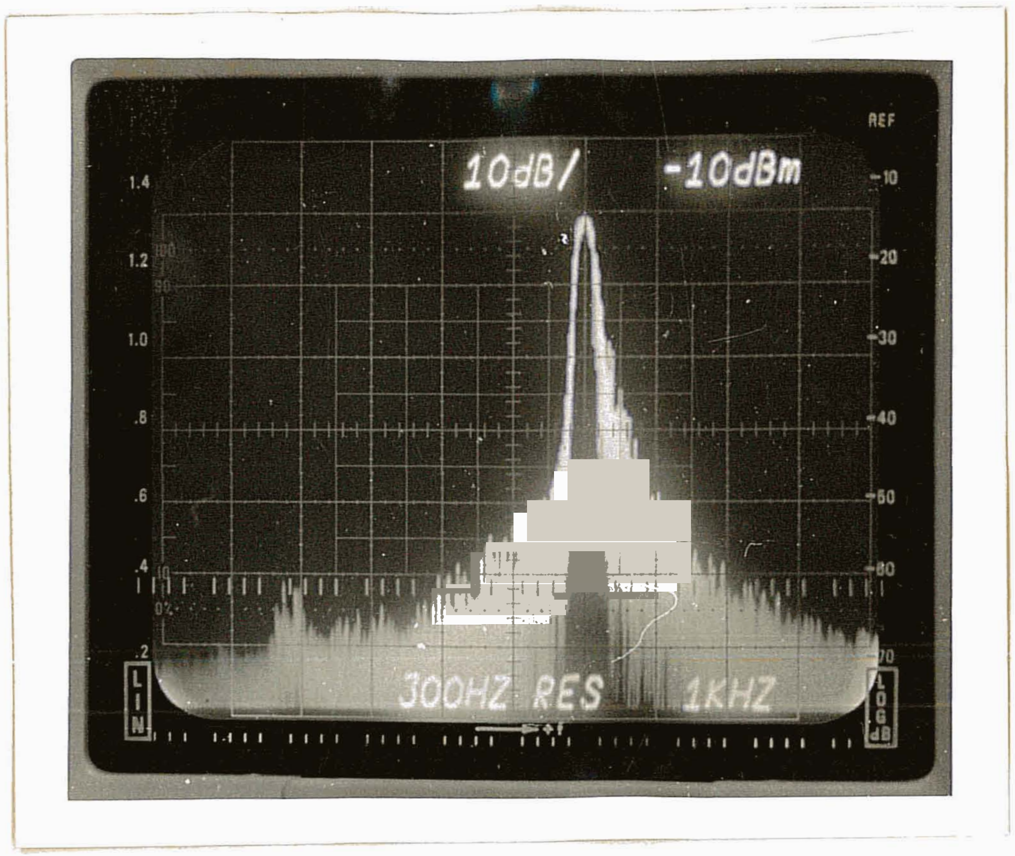
10dB/
-10dBm
REF
300HZ RES
1KHZ
LIN
LOG dB

Fig. 5.1

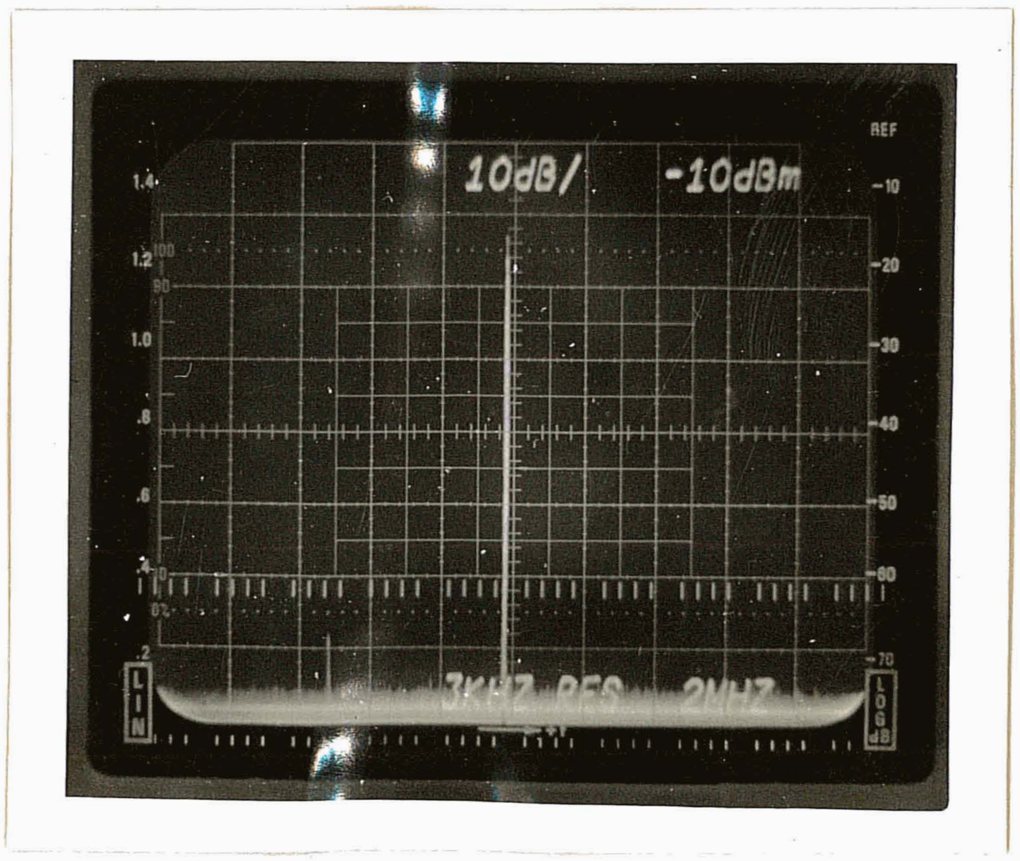
10dB/
-10dBm
REF
3KHZ RES
2MHZ
LIN
LOG dB

Fig. 5.2

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