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Annexure A

For the purpose of explanation, with specific reference to Chapter 4, it was decided to extract relevant information from a literature study conducted by the student, to be presented in this Annexure to the thesis.

Broadband transformers

Broadband transformers are employed in circuits which must have a uniform response over a substantial spread of frequency. To be able to consider all factors involved, three specific papers were consulted [14, 15, 16] and some of the information included in the following sections, before the detailed description in chapter 4.2 of the hybrid combiners that were developed for the purpose of this thesis. Other references [17 - 20] are referred to for the sake of completeness.

A.1 Broadband transformer requirements

The broadband transformer is a device that will step up or down a voltage or current, or step up or down the impedance, or provide DC isolation, or any combination of these. In the process of providing the above, the transformer reduces the transmitted signal slightly. This is called attenuation or insertion loss. A typical plot of the insertion loss versus frequency is shown in Fig. A.1.

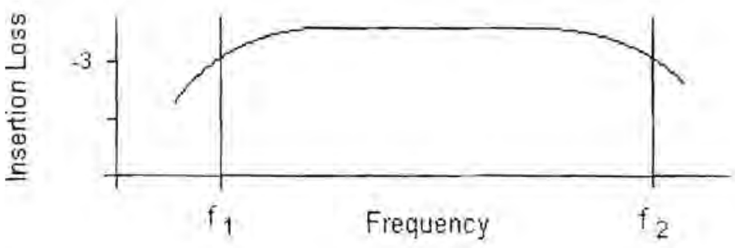


Figure A.1. Typical curve of Insertion loss vs. Frequency for broadband transformers

From this graph it can be seen that the insertion loss increases at frequencies below f_1 and above f_2, which are called the cut-off frequencies of the device. The useful range of the transformer is the frequency range between f_1 and f_2, and the problem of designing a broadband transformer is to extend the bandwidth of the unit while keeping the insertion loss at a minimum.

A.2 The equivalent circuit

The equivalent circuit of the device must be reviewed to understand the design of the broadband transformer. In Figure A.2 is shown the ideal transformer (one that does not have any losses), with resistors, capacitors and inductors that represent those factors that generate losses within the unit. These elements determine the cut-off frequencies of the device, where R_c = coil resistance, L_l = leakage reactance, L_p = parallel inductance of core, and C_d = winding capacitance.

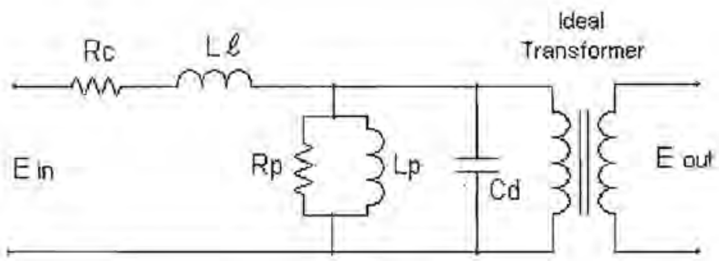


Figure A.2. Lumped element equivalent circuit of a transformer

In this diagram elements marked L_p and R_p represent the inductance and losses in the magnetic core of the transformer. These items are shown in parallel with the ideal transformer and it indicates therefore, that if the values for inductive reactance (X_p) and resistance (R_p) of the core are reduced, the output of the ideal transformer is also reduced. Since the values of L_p and R_p are usually lowest at low frequencies, they are the main elements responsible for the low frequency cut-off, f_1.

The elements marked L_l and C_d in the diagram represent the leakage inductance and winding capacitance respectively of the transformer. As frequency is increased the inductive reactance of L_l increases and the capacitive reactance of C_d decreases which also reduces the output of the transformer. Therefore these elements are responsible for the high frequency cut-off, f_2.

Since these elements largely determine both the high and low cut-off frequencies of the transformer, their values for particular cores must be known for design purposes, if they are to be changed or modified to meet bandwidth requirements.

A.3 The ferrite material data

The magnetic properties of ferrites are usually presented as the variation of permeability and loss factor of the ferrite materials at various frequencies.

The initial permeability and loss factor of a ferrite material is a result of measurements on a core, expressed as though it were an inductor and a resistor in series. Such measurements are generally useless to the designer because he needs his information in a different form.

As described in para A.2 and Figure A.2, the designer needs material and core information in terms of the parallel components of the inductance and resistance values. This data is the result of measurements of cores expressed as an inductor and a resistor in parallel.

The manufacturer often refers to the material cut-off frequency in describing ferrite materials, which is the point where the series permeability has dropped a significant amount from its low frequency value. For broadband transformers the important parameter is the parallel inductive reactance (X_p), which is very nearly equal to the series permeability multiplied by the frequency. Thus, although the permeability has become less with increasing frequency, the X_p is either increasing or remaining constant and the material remains quite useful as a broadband transformer.

Therefore, when selecting a material for a broadband transformer application, it is necessary to compare the parallel components of the magnetic parameters such as the parallel inductive reactance (X_p) and the parallel resistance (R_p), as a function of frequency, rather than as the initial permeability and loss factors of various materials. It is for this reason that the Fair-Rite Products Corporation supply curves of the parallel parameters versus frequency for all their materials commonly used for broadband transformers [14].

Three materials that best cover the high frequency range, were selected by measuring the parallel parameters of all materials versus frequency. In Fig. A.3, a graph is shown of the total impedance for a particular balun core, per turn, versus frequency, for each of the three materials, namely #73 material, #65 material and #43 material.

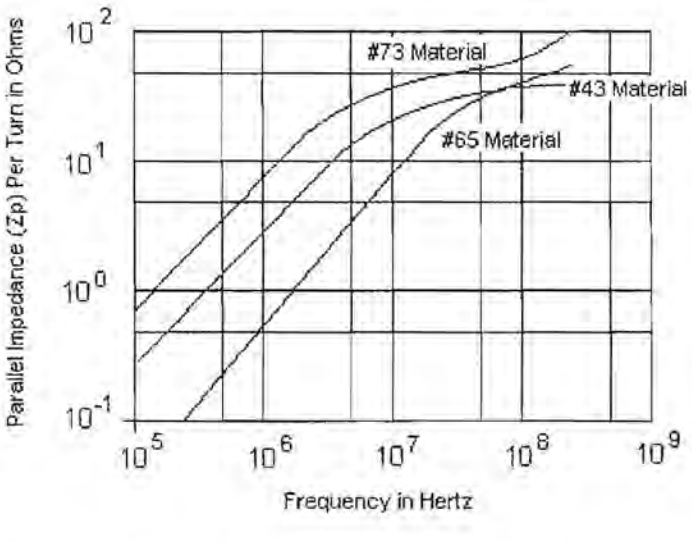


Figure A.3. Parallel impedance per turn for balun type core in 3 materials [14].

Fig. A.3 indicates the typical differences in impedance that can be expected between these three materials in cores of the same size and shape. The graph shows that #73 material has a higher impedance at all frequencies than either of the other two materials and should therefore be considered as the best material for broadband transformers.

For a broadband transformer, it is necessary that the lowest impedance within the frequency span of the transformer be used in determining its attenuation, which in most cases occurs at the lowest end of the transformer bandwidth. Since the curves in Fig. A.3 all have a positive slope, the critical impedance for them is regarded as the one at the low end of the desired frequency range.

However, the resistivity of this material is low (typically 100-ohm per cm), which means that if the windings of the core become shorted to the ferrite, the input impedance of the wound transformer is decreased. To prevent this, the core is usually wound with an insulating tape or a protective coating is added, which increases the cost of the complete device.

As an alternative, the #65 material may be considered for frequencies above 80 MHz, since its impedance is higher than #43 material at this frequency. Below 80 MHz, #43 material is preferred, as it has the higher impedance at this instance.

A.4 The core shapes

After following the process of selecting a material for a particular application, it is necessary to select the core and the number of turns needed for the finished transformer. In order to assist in this selection, the manufacturer included in their application notes a series of graphs depicting the impedance, reactance and resistance per turn versus frequency for each of the standard broadband transformer cores in each of the three transformer materials [14]. This information for #65 material is shown in Fig. A.4, Fig. A.5 and Fig. A.6, with each of the parameters for three core shapes shown on one graph. All the other cores' characteristics fall within the limits of item 5 and item 8.

These graphs were all made using a single turn through both holes of the core. Therefore the magnetic parameter will equal the number taken from the graph, multiplied by the number of turns squared.

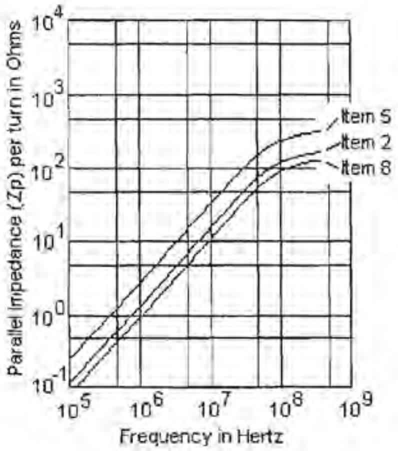


Fig. A.4.
Parallel Impedance per turn vs. Frequency (Fair-Rite #65 material)

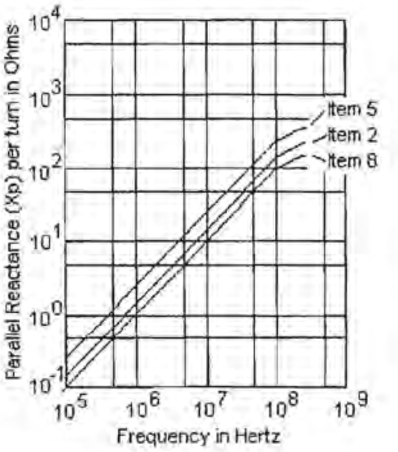


Fig. A.5.
Parallel Reactance per turn vs. Frequency (Fair-Rite #65 material)

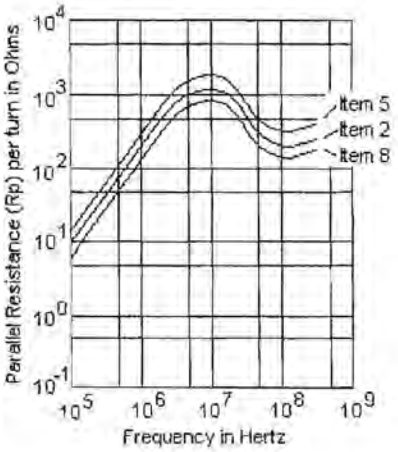


Fig. A.6.
Parallel Resistance per turn vs. Frequency (Fair-Rite #65 material)

The cores most often used for winding broadband transformers at high frequencies are the two hole balun types as shown in Fig. A.7. These cores may be wound either with the winding through one hole and around the outside or through both holes.

The latter method produces the higher inductance per turn and is the method used in generating the impedance curves shown in Fig's. A.4, A.5 and A.6 [14].

Many designers use the single hole ferrite bead for winding broadband transformers with good results [12]. The main advantage of using a shield bead with a single hole for a transformer core is that it is less costly than the balun type. However, it will not produce a transformer with as wide a bandwidth as the two hole type [14].

A.5 The form factor

In previous sections it has been mentioned that the core inductance, leakage reactance and winding capacitance all have an effect on the cut-off frequencies of the transformer. Since these parameters vary between core sizes and shapes, it would be useful to have a numerical factor for each core that denotes its relative value as a broadband transformer. It can be shown that such a number may be generated so that the lower the number determined for a particular core, the wider the frequency range of the finished transformer. This number, known as the form factor, is defined as follows: Form factor = $\ell_w C_\ell$, where $C_\ell = \dfrac{\ell_e}{A_e}$, ℓ_w = length of one complete turn of wire, ℓ_e = effective magnetic path length and A_e = effective magnetic area. This form factor has been calculated by Fair-Rite Products for each core listed in Fig. A.7 [14].

The lowest form factors (and highest frequency response) are for items 4 through 7, which are types 3 and 4 balun cores. The next lowest form factor is for the type 1 balun core while the toroid has the highest.

The toroid is sometimes used with several cores in parallel [12]. The form factor for such an assembly would be markedly lower than the factor for a single toroid, due to several increased parameters, but mainly the effective magnetic area A_e. Figure A.7 also shows that the form factor for shield beads (type 7) is somewhat higher than for balun cores, but they are in the same general vicinity and therefore, make acceptable transformers.

Interesting to note is that items 1, 2 and 3 in Fig. A.7 are the two hole balun cores with similar form factors. The main difference between them is their overall size and hole size. In general the smallest is the least expensive in first cost and mounting space, but it also has a small hole for the winding. If the hole size cannot accommodate the number of turns and wire size required, the next size core has to be considered.

It is obvious, therefore, that many factors other than the form factor alone should be considered in choosing a core, but it is an important tool which can make the designer's job easier.

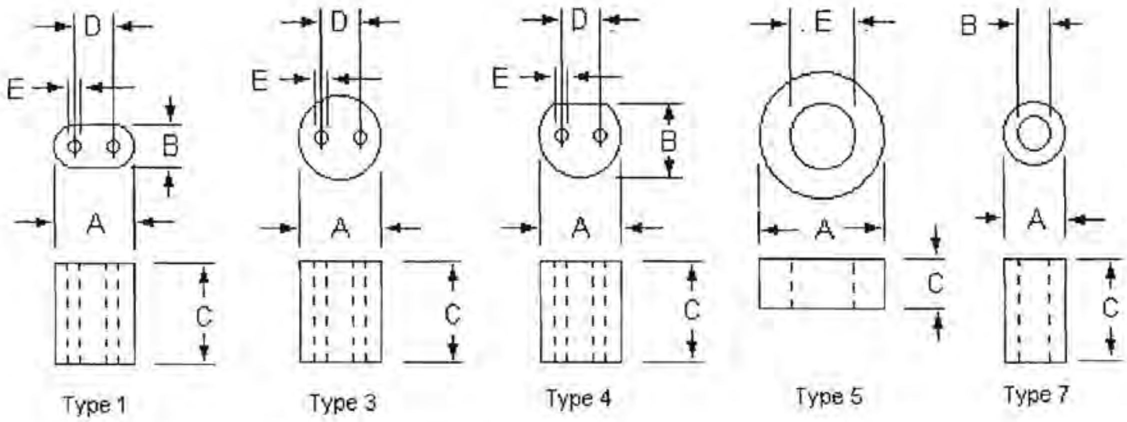


Figure A.7. List of ferrite core shapes for broadband transformer design [14].

Item	Part Number	Core Shape	Nominal Dimensions (inches)					Form Factor
			A	B	C	D	E	
1	28_ _000302	Type 1	.525	.295	.407	.225	.150	13.0
2	28_ _002402	Type 1	.277	.160	.244	.114	.071	14.3
3	28_ _002302	Type 1	.136	.079	.093	.057	.034	14.0
4	28_ _001802	Type 3	.250	-	.242	.100	.050	9.5
5	28_ _001702	Type 3	.250	-	.471	.100	.052	8.6
6	28_ _000902	Type 4	.284	-	.218	.104	.035	8.8
7	28_ _002802	Type 3	.220	-	.250	.090	.197	7.8
8	26_ _002402	Type 5	.350	-	.190	-	-	29.0
9	26_ _000101	Type 7	.138	.051	.118	-	-	17.3
10	26_ _000201	Type 7	.076	.043	.150	-	-	24.3
11	26_ _000301	Type 7	.138	.051	.236	-	-	14.9
12	26_ _000401	Type 7	.138	.051	.500	-	-	13.7
13	26_ _000801	Type 7	.295	.094	.297	-	-	11.8
14	26_ _002401	Type 7	.200	.082	.250	-	-	15.2
15	26_ _021801	Type 7	.200	.082	.437	-	-	12.3

A.6 The circuit configurations

Broadband transformers have been described in detail by C. L. Ruthroff [15] as well as G. Guanella [16], resulting in two models, aptly named the 'Ruthroff balun' and the 'Guanella balun'. Certain variations of these unique configurations have also been discussed elsewhere, and are included here for the sake of completeness [17, 18, 19, 20]. The transformer models are shown in Figs. A.8 to Fig. A.16. When drawn in the transmission line form, the transforming properties are sometimes difficult to see. For this reason, a more conventional form is shown with the transmission line form, with some winding arrangements [15].

In conventional transformers the interwinding capacitance resonates with the leakage inductance producing a loss peak. This mechanism limits the high frequency response. In transmission line transformers, the coils are so arranged that the interwinding capacitance is a component of the characteristic impedance of the line, and as such forms no resonance to seriously limit the bandwidth. Also, for this reason, the windings can be spaced closely together maintaining good coupling.

The net result is that transformers can be built this way which have good high frequency response. In all of the transformers for which experimental data are presented, the transmission lines take the form of twisted pairs. In some configurations the high frequency response is determined by the length of the windings and while any type of transmission line can be used in principle, it is quite convenient to make very small windings with twisted pairs. The sketches showing the conventional form of transformer indicates clearly that the low frequency response is determined in the usual way, by the primary inductance. The larger the core permeability, the fewer the turns required for a given low frequency response and the larger the overall bandwidth. Thus a good core material is desirable.

The transformer configurations are shown using ferrite toroids, for explanatory purposes. Ferrite toroids have been found very satisfactory [14]. The permeability of some ferrites is very high at low frequencies and falls off at higher frequencies. Thus, at low frequencies, large reactance can be obtained with few turns.

When the permeability falls off, the reactance is maintained by the increase in frequency and good response is obtained over a large frequency range. It is important that the coupling is high at all frequencies or the transformer action fails. Fortunately, the bifilar winding tends to give good coupling.

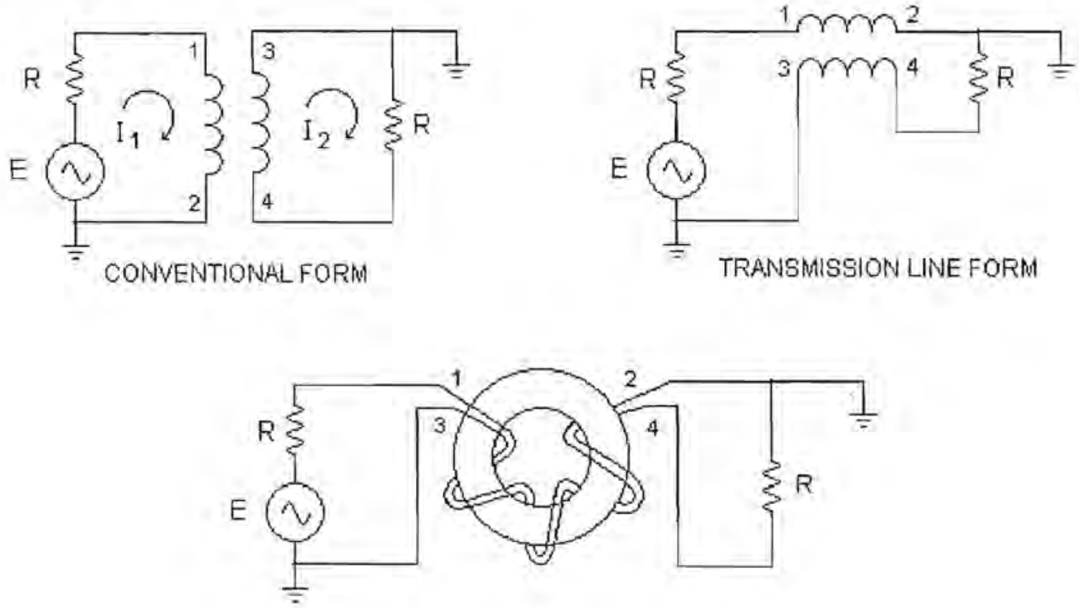


Fig. A.8. Reversing transformer

A.6.1 Ruthroff's polarity reversing transformer - Fig A.8 [15]

This transformer consists of a single bilfilar winding and is the basic building block for most of the transformer configurations presented here.

The fact that a reversal is obtained can be seen from the conventional form, which indicates the current polarities. Both ends of the load resistor are isolated from ground by coil reactance. Either end of the load resistor can then be grounded, depending upon the output polarity desired. If the centre of the resistor is grounded, the output is balanced. A suitable winding consists of a twisted pair of insulated wire. In such a winding, the primary and secondary are very close together, insuring good coupling. The interwinding capacitance is absorbed by the characteristic impedance of the line.

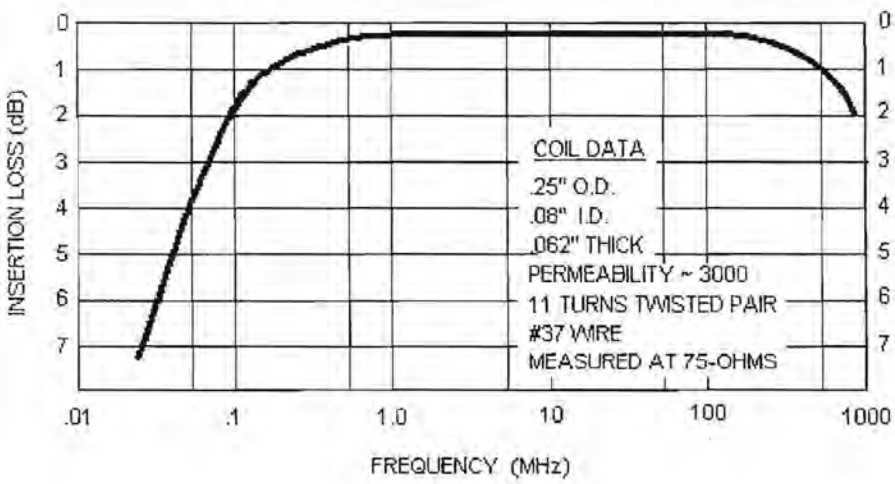


Fig. A.8 (a) Ruthroff's reversing transformer. Insertion loss vs. frequency [15].

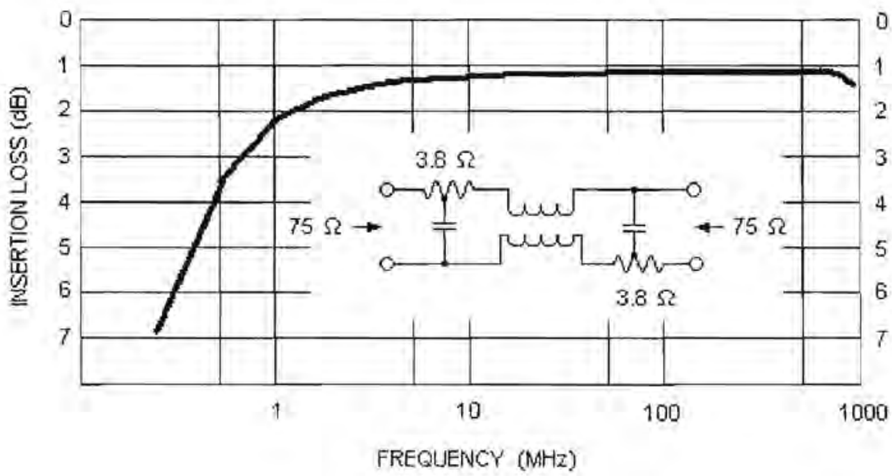


Fig. A.8 (b) Ruthroff's matched reversing transformer [15].

At high frequencies this transformer may be regarded as an ideal reversing transformer plus a length of transmission line. If the characteristic impedance of the line is equal to the terminating impedances, the transmission line is inherently broadband. If not, there will be a dip in the response at the frequency at which the transmission line is a quarter wavelength long. The depth of the dip is a function of the ratio of terminating impedance to line impedance and is easily calculated.

Experimental data on a reversing transformer are shown in Fig. A.8 (a) and Fig. A.8 (b) [15]. Fig. A.8 (a) is the response of a transformer with no extra impedance matching. The return loss of this transformer to a 3 nanosecond pulse is 20-dB.

The transformer of Fig A.8 (b) has been adjusted to provide more than 40-dB return loss to a 3 nanosecond pulse by C. L. Ruthroff [15]. The transformer loss (about 0,5-dB before matching) is matched to 75-ohms with the two 3,8-ohm resistors. The inductance is tuned out with the capacitance of the resistors to the ground plane. The match was adjusted while observing the reflection of a 3 nanosecond pulse.

A.6.2 **Unbalanced-to-Balanced 1:1 Impedance transformer - Fig. A.9 [15]**

This is similar to Fig. A.8 except that an extra length of winding is added. This is necessary to complete the path for the magnetising current.

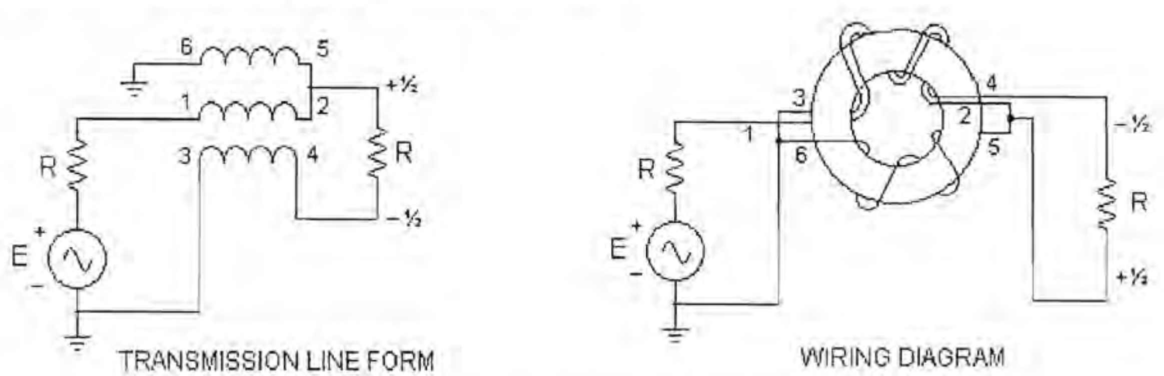


Fig. A.9. Unbalanced to balanced transformer [15].

A.6.3 Unbalanced-Unsymmetrical 4:1 Impedance transformer - Fig. A.10.

This transformer is interesting because with it a 4:1 impedance transformation is obtained with a single bifilar winding such as used in the reversing transformer. The transforming properties are evident from Fig. A.10. Not so easily seen is the high frequency cut-off characteristic.

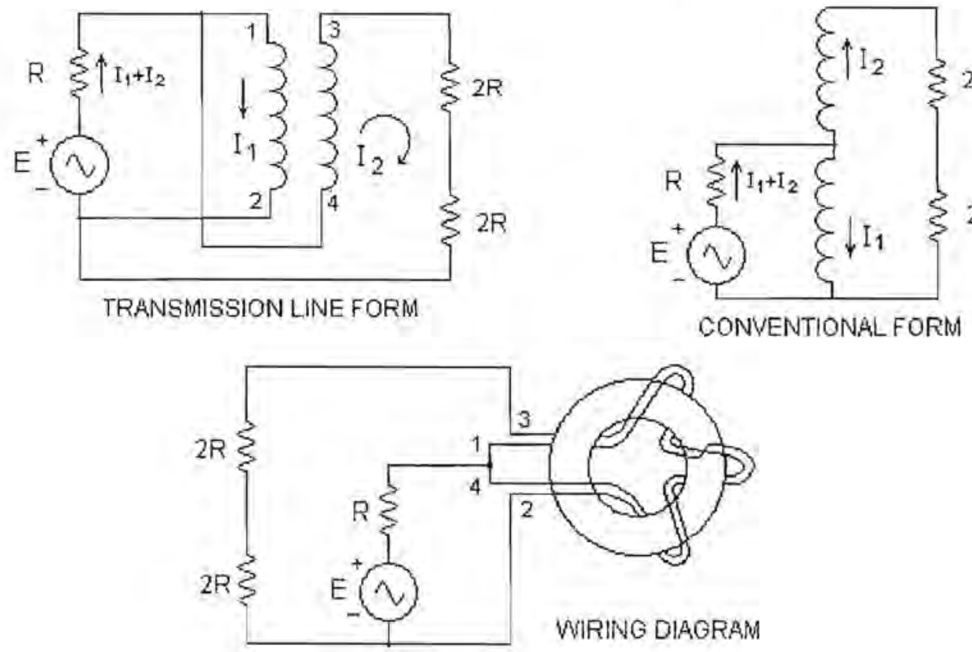


Fig. A.10. Unbalanced-Unsymmetrical 4:1 impedance transformer [15].

The response of this device at high frequencies is derived from the transmission line equivalent. With the characteristic impedance of the winding denoted as Z_o, the reactance of winding W may be denoted as $X >> R_L$, R_g. The loop equations are as follows:

$$e = (I_1 + I_2)R_g + W_1$$

$$e = (I_1 + I_2)R_g - W_2 + I_2R_L$$

$$V_1 = V_2 \cos\beta.\ell + jI_2 Zo \sin\beta.\ell \quad \ldots (1)$$

$$I_1 = I_2 \cos\beta.\ell + j\frac{V_2}{Zo}\sin\beta.\ell \quad \ldots (2)$$

This set of equations is solved for the output power P_O, where $P_O = |I_2|^2 R_L$, from which follows:

$$P_O = |I_2|^2 R_L = \frac{e^2(1+\cos\beta.\ell)^2 R_L}{\left[2R_g(1+\cos\beta.\ell) + R_L\cos\beta.\ell\right]^2 + \left[\frac{R_g R_L + Z_o^{\ 2}}{Z_o}\right]^2 \sin^2\beta.\ell} \quad \ldots\ldots (3)$$

From this expression, the conditions for maximum power transmission are obtained by setting $\ell = 0$ and setting $\left|\frac{dP_o}{dR_L}\right|_{\ell=0} = 0$. The transformer is matched when $R_L = 4R_g$. The optimum value for Zo is obtained by minimising the coefficient of $\sin^2\beta.\ell$ in (3). In this manner the proper value for Zo is found to be $Z_o = 2R_g$.

Now, setting $R_L = 4R_g$ and $Z_o = 2R_g$, (3) reduces to:

$$P_o = \frac{e^2(1+\cos\beta.\ell)^2}{R_g\left[(1+3\cos\beta.\ell)^2 + 4\sin^2\beta.\ell\right]} \quad \ldots\ldots (4)$$

Also,

$$P_{available} = \frac{e^2}{4R_g} \quad \ldots\ldots (5)$$

and dividing (4) by (3):

$$\frac{Power\ available}{Power\ Output} = \frac{(1+3\cos\beta.\ell)^2}{R_g\left[(1+3\cos\beta.\ell)^2 + 4\sin^2\beta.\ell\right]} \quad \ldots\ldots (6)$$

where β is the phase constant of the line, and *l* is the length of the line. Thus, the response is down 1 dB when the line length is λ/4 wavelengths and the response is zero at λ/2.

This function is plotted in Figure A.11.

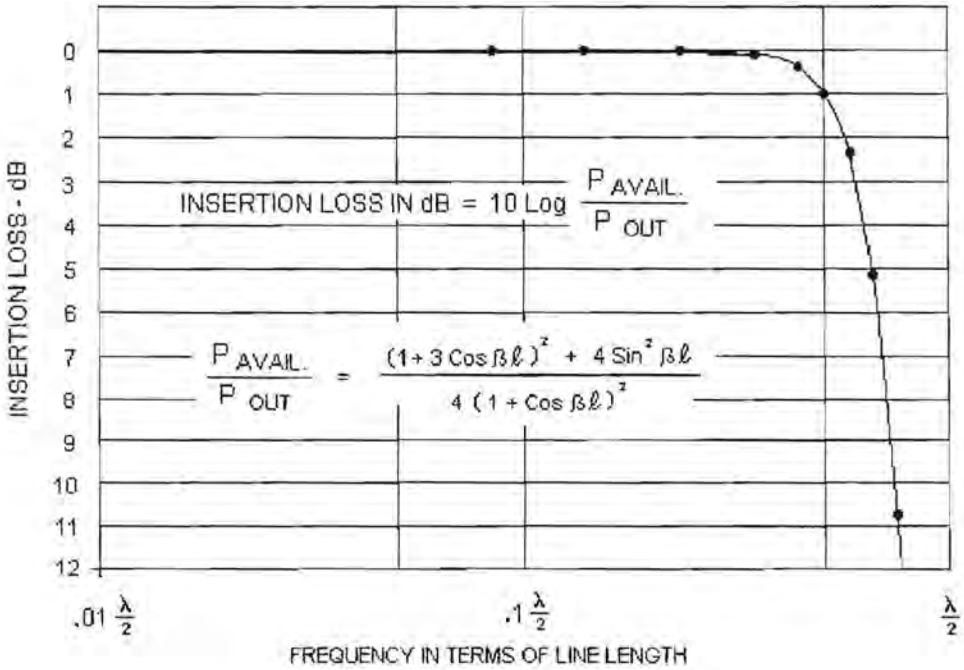


Figure A.11. Theoretical insertion loss vs frequency [15]

The impedances seen at either end of the transformer, with the other end terminated in Z_L have been derived [15]. They are:

$$Z_{in} \text{ (low impedance end)} = Z_o\left(\frac{Z_L\cos\beta\ \ell + jZ_o\sin\beta\ \ell}{2Z_o(1+\cos\beta\ \ell) + jZ_L\sin\beta\ \ell}\right) \quad \ldots\ldots\ldots (7)$$

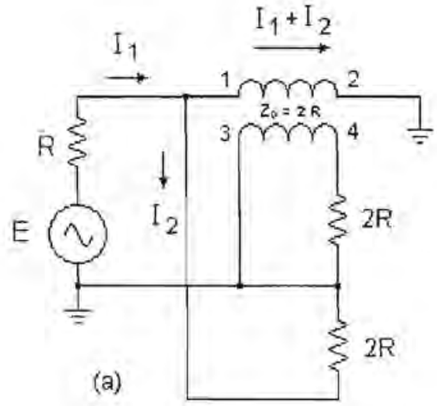
and

$$Z_{in} \text{ (high impedance end)} = Z_o\left(\frac{2Z_L(1+\cos\beta\ \ell) + jZ_o\sin\beta\ \ell}{Z_o\cos\beta\ \ell + jZ_L\sin\beta\ \ell}\right) \quad \ldots\ldots\ldots (8)$$

A.6.4 Balanced-to-Unbalanced 4:1 impedance transformers - Fig. A.12.

The circuitry in Fig. A.12 is rather simple. The single bifilar winding is used as a reversing transformer as in Fig. A.8. The high frequency cut-off is the same as that for the transformer of Fig. A.10.

In some applications it is desirable to omit the physical ground on the balanced end. In such cases, Fig. A.12.(b) can be used. The high frequency cut-off is the same as for the transformer of Fig. A.10.



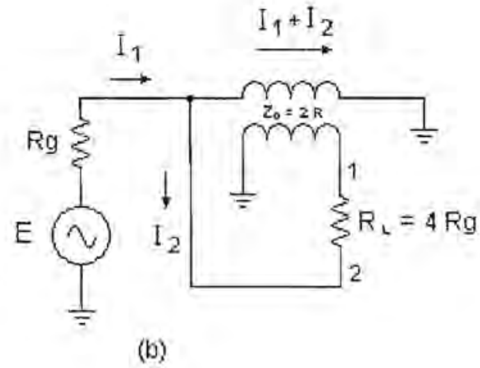


Fig. A.12. Balanced-Unbalanced 4:1 impedance transformer [15]

In the low frequency analysis of the transformer in Fig. A.12.(b), the series impedance of each half of the bifilar winding is denoted by Z.

The loop equations are:

$$E = \left(R_g + Z\right)I_1 - (Z + kZ)I_2$$

$$E = \left(R_g - kZ\right)I_1 + \left(R_L + Z + kZ\right)I_2 \quad \text{.......................} (9)$$

from which
$$\frac{I_1}{I_2} = \frac{R_L + 2Z(1+k)}{Z(1+k)} \approx 2 \text{ } if \text{ } Z >> R_L \quad \text{..................} (10)$$

The voltages from points 1 and 2 to ground may now be calculated as $V_{2G} = I_1 R_g$,

and when the transformer is matched, $E = 2I_1 R_g$, and $V_{2G} = I_1 R_g$ .... (11)

Similarly, $V_{1G} = I_2 Z - kZ(I_1 - I_2)$. With the aid of (10), this may be rearranged to:

$$V_{1G} = ZI_1 \left[\frac{Z(1+k)^2 - kR_L - 2kZ(1+k)}{R_L + 2Z(1+k)} \right] \quad \text{..............} (12)$$

Now let the coupling coefficient $k = 1$, then

$$V_{1G} = I_1 Z \left[\frac{-kR_L}{R_L + 2Z(1+k)} \right] \approx -\frac{I_1 R_L}{4} \quad \text{for } Z >> R_L.$$

When the transformer is matched, $R_L = 4R_g$, so that $V_{1G} = I_1 R_g = -V_{2G}$, ... (13)

the load is balanced with respect to ground.

From (13) it is clear that the centre point of R_L is at ground potential, and can therefore be grounded physically, resulting in Fig. A.12.(a).

A.6.5 **Hybrid circuits - Fig. A.13 to Fig. A.16 [15]**

Various hybrid circuits are developed from the basic form using the transformer configurations discussed previously. The drawings are very nearly self-explanatory. In all hybrids in which all four arms are single ended, it has been found necessary to use two cores when ferrite toroids are used, in order to get proper magnetising currents [15].

Figure A.13.(a) indicates the Ruthroff basic hybrid and Figure A.13.(b) the unsymmetrical version of the same, with equal conjugate impedances.

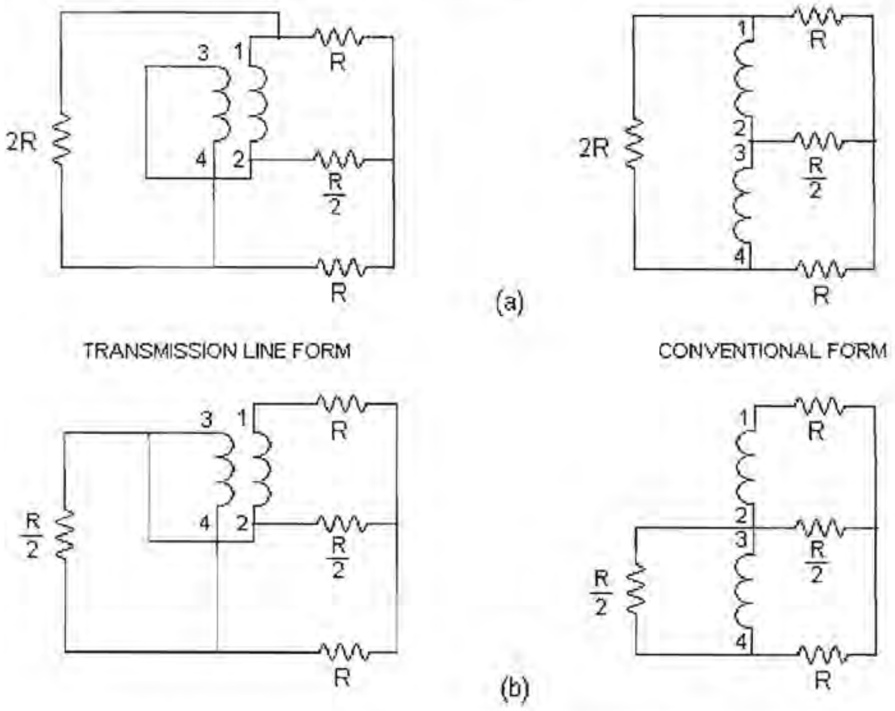


Fig. A.13. (a) Basic hybrid. (b) Unsymmetrical hybrid

Fig. A.14 indicates the Guanella symmetrical hybrid with conjugate impedances [16]. This hybrid is used very effectively as an antenna balun, specially in HF applications [20].

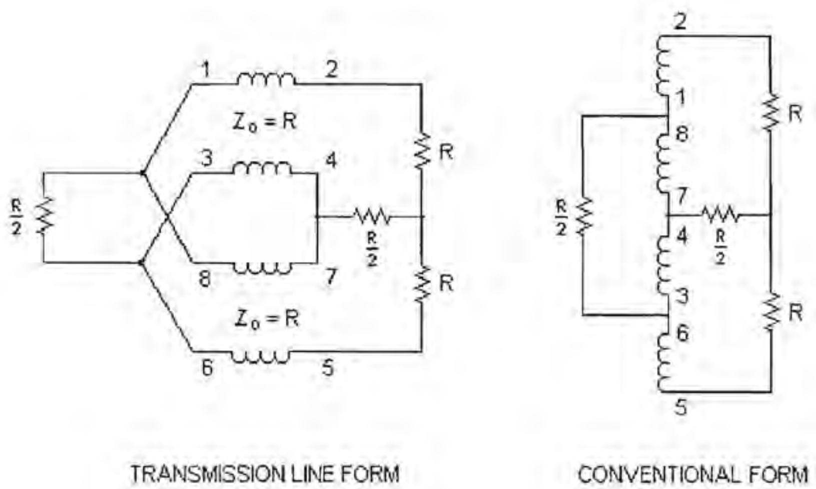
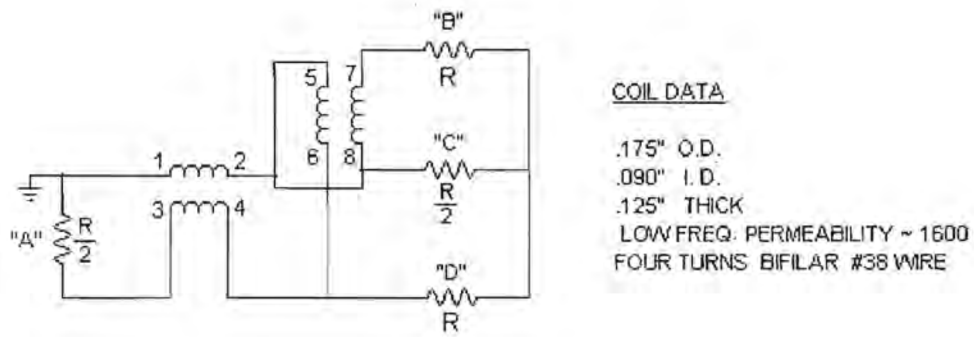


Fig. A.14. The Guanella hybrid

Two hybrids have been measured and data included here [15]. The response of a hybrid of the type shown in Fig. A.15 is given with the figure of the same. For this measurement R = 150-ohms. In order to measure the hybrid in a 75-ohm circuit, arms B and D were measured with 75-ohm resistors in series with the 75-ohm measuring gear. This accounts for 3-dB additional loss measured. Apart from these conditions arms B and D have a 3-dB insertion loss with respect to arm C.



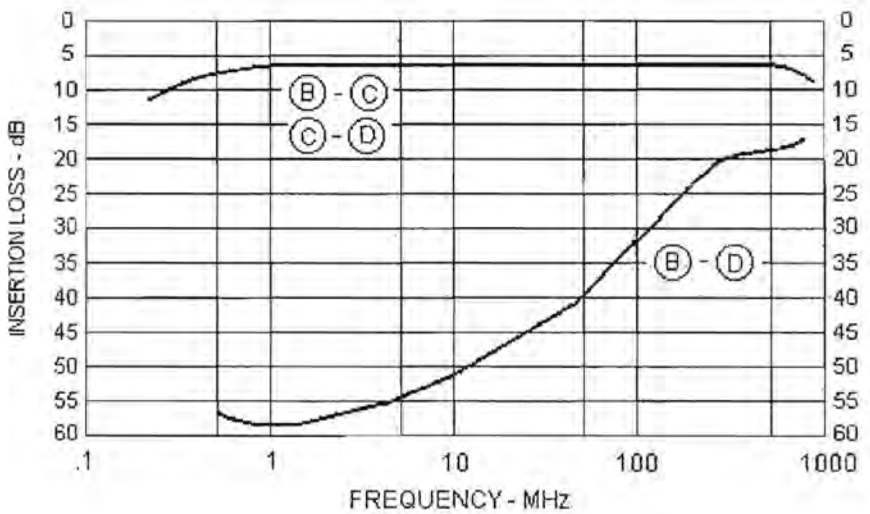


Fig. A.15.(a) Hybrid with equal conjugate impedances. (b) Insertion loss vs frequency [15]

As a power combiner, two signals fed into ports C and D yield their vector sum a port B.
As a power divider, a signal fed into port B divides into two equal amplitude, equal phase, isolated output signals at ports C and D. A signal fed into port A divides into equal amplitude, 180° out of phase, isolated signals at ports C and D.

The transmission characteristics of Fig. A.16 is shown with the same. The hybrid takes the form of a resistance bridge, with all arms single ended. This hybrid, with equal impedance loads, has been matched using the technique described previously in para. A.6.1 for the reversing transformer in Fig. A.8.(b). The results of this matching are included in the figure. [15]. The hybrid has a 3dB loss, dissipating in the resistive elements of the circuit, in addition to transformer loss.

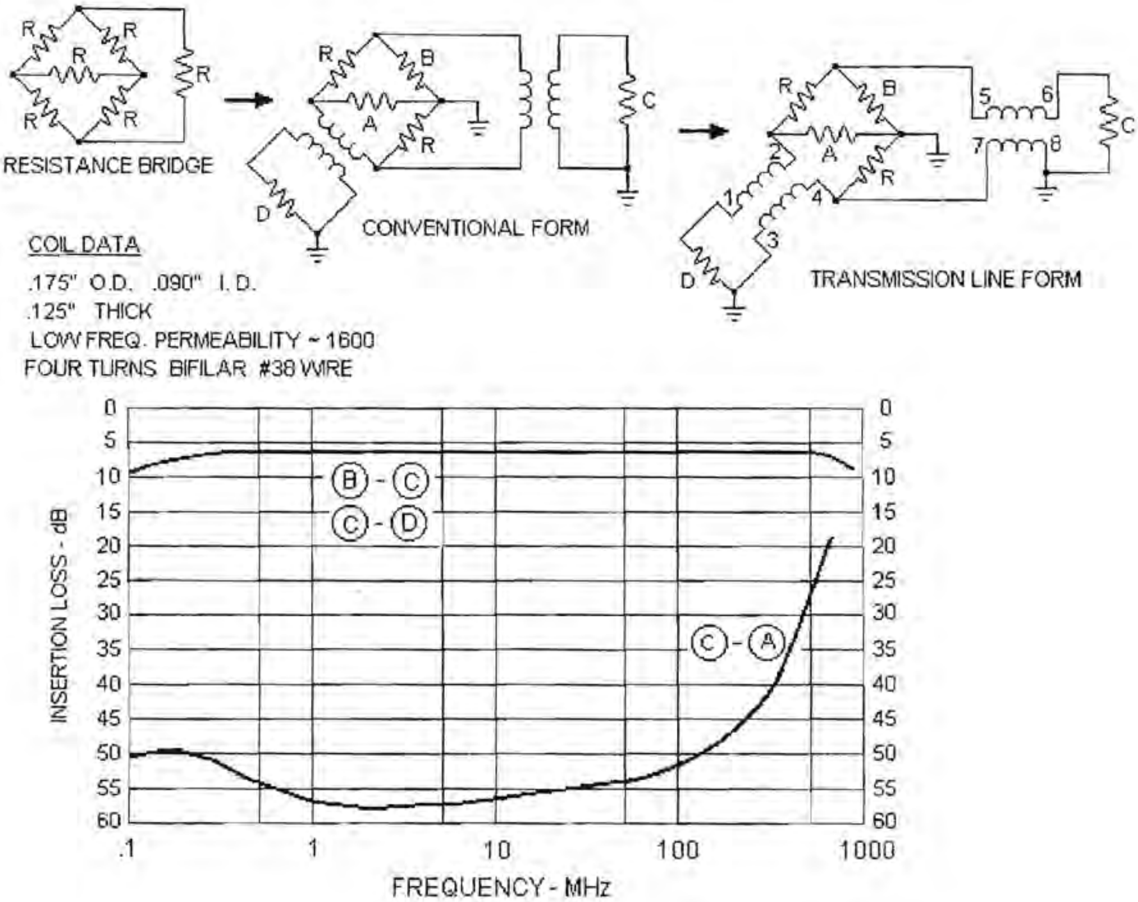


Figure A.16. (a) Matched resistance hybrid. (b) Insertion loss vs frequency [15].

When considering hybrids for use as power dividers or combiners, it is interesting to note that resistors used within the matched circuit in Fig. A.16 have the same insertion loss as the circuit depicted in Figure A.15, but with improved return loss at higher frequencies.

Annexure B

ACCURACY OF THE NOISE MEASUREMENTS

The measurement gear used to perform the noise measurements was the Hewlett Packard HP 8593E Spectrum Analyser with HP 85714A Scalar Measurement Personality option [21].

When resolving signals of equal or different amplitudes in terms of their power levels, one must consider the characteristics of the measuring gear before comments may be made about the accuracy of the measurement.

B.1 Resolving signals of equal amplitude using the resolution bandwidth function

In responding to a continuous wave signal, a swept-tuned spectrum analyser like the HP 8593E traces out the shape of the spectrum analyser's intermediate frequency (IF) filters. As we change the filter bandwidth, we change the width of the displayed response. If a wider filter is used and two equal-amplitude input signals are close enough in frequency, then the two signals appear as one. Thus the signal resolution is determined by the IF filters inside the spectrum analyser.

The resolution bandwidth (RES BW) function selects an IF filter setting for a measurement. Resolution bandwidth is defined as the 3-dB bandwidth of the filter. The 3-dB bandwidth tells us how close together equal amplitude signals can be and still be distinguished from one another.

Generally, to resolve two signals of equal amplitude, the resolution bandwidth must be less than or equal to the frequency separation of the two signals. If the bandwidth is equal to the separation a dip of approximately 3-dB is seen between the peaks of the two equal signals, and it becomes clear that more than one signal is present. See Figure B.2.

In order to keep the spectrum analyser calibrated, sweep time is automatically set to a value that is inversely proportional to the square of the resolution bandwidth. So, if the resolution factor is decreased by a factor of 10, the sweep time is increased by a factor of 100 when sweep time and bandwidth settings are coupled. (Sweep time is proportional to $1/BW^2$).

For fastest measurement times, use the widest resolution bandwidth that still permits discrimination of all desired signals. The HP8593E allows you to select from 30-Hz to 3-MHz resolution bandwidth in a 1, 3, 10 sequence, plus 5-MHz for maximum flexibility.

Example: Resolve two signals of equal amplitude with a frequency separation of 100-kHz.

B.1.1. To obtain two signals with a 100-kHz separation, connect the calibration signal and a signal generator through a suitable coupler (see Figure 4.6, p.39) to the spectrum analyser input as shown in Figure B.1. (If available, two signal generators may be used.)

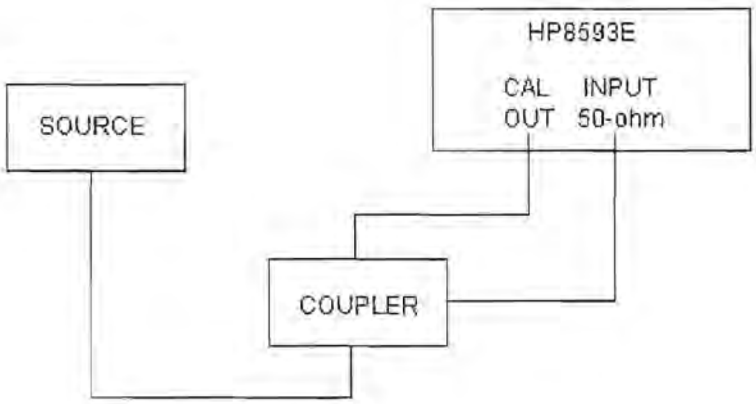


Figure B.1. Set-up for obtaining two signals

B.1.2 If using the 300-MHz calibration signal, set the frequency of the generator 100-kHz greater than the calibration signal, (i.e. 300,1-MHz). The amplitude of both signals should be approximately -20-dBm.

B.1.3 On the spectrum analyser, press [PRESET]. Set the centre frequency to 300-MHz, the span to 2-MHz, and the resolution bandwidth to 300-kHz by pressing [FREQUENCY] 300 [MHz], [SPAN] 2 [MHz], then [BW] 300 [kHz]. A single peak is visible.

B.1.4 Since the resolution bandwidth must be less than or equal to the frequency separation of the two signals, a resolution bandwidth of 100-kHz must be used. Change the resolution bandwidth to 100-kHz by pressing [BW] 100 [kHz]. Two signals are now visible as in Figure B.2. The rotating knob or step keys may now be used to further reduce the resolution bandwidth to better resolve the signals.

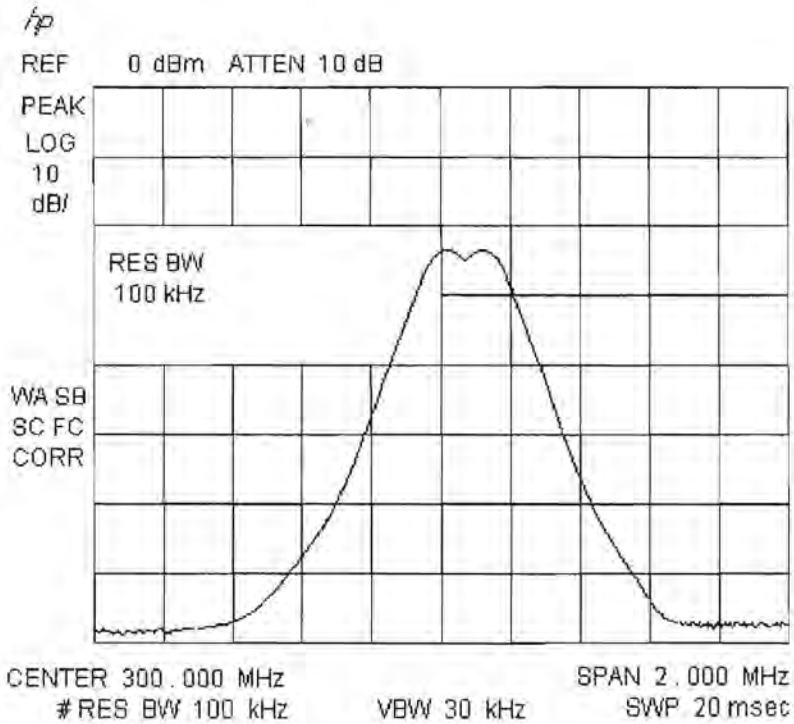


Figure B.2. Resolving signals of equal amplitude

As the resolution bandwidth is decreased, resolution of the individual signals is improved and the sweep time is increased. For the fastest measurement times, use the widest possible resolution bandwidth. Under preset conditions, the resolution bandwidth is coupled (or linked) to span.

If the resolution bandwidth has been changed from the coupled value, a "#" mark appears next to RES BW in the lower-left corner of the screen, indicating that the resolution bandwidth is uncoupled.

B.2 Resolving small signals hidden by large signals

When dealing with the resolution of signals that are not equal in amplitude, one must consider the shape of the IF filter as well as its 3-dB bandwidth. The shape of the filter is defined by the shape factor, which is the ratio of the 60-dB bandwidth to the 3-dB bandwidth. (Generally, the IF filters in the HP4593E spectrum analyser have shape factors of 15:1 or less)

If a small signal is too close to a larger signal, the smaller signal can be hidden by the skirt of the larger signal. To view the smaller signal, or to resolve the large signal in terms of its amplitude, one must select a resolution bandwidth noticeably smaller than the frequency separation, such that k is less than a. See Figure B.3. The separation between the two signals must be greater than half the filter width of the larger signal, at the amplitude level of the smaller signal.

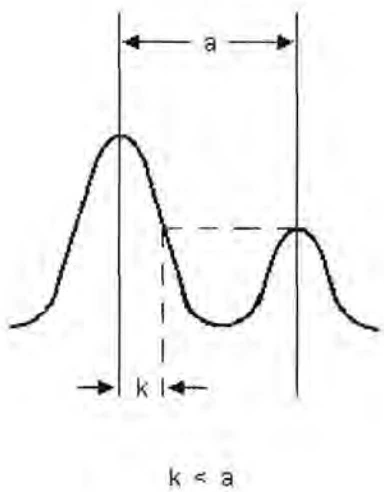


Figure B.3. Resolution bandwidth requirements for resolving small signals.

Example Resolve two input signals with a frequency separation of 200-kHz and an amplitude separation of 60-dB.

B.2.1. To obtain two signals with a 200-kHz separation, connect the equipment as shown in Figure B.1.

B.2.2. Set the centre frequency to 300-MHz and the span to 2-MHz; press [FREQUENCY] 300 [MHz], then [SPAN] 2 [MHz].

B.2.3. Set the source to 300,2-MHz, so that the signal is 200-kHz higher than the calibration signal. Set the amplitude of the signal to -80-dBm (60-dBm below the calibration signal)

B.2.4. Set the 300-MHz signal to the reference level by pressing [PEAK SEARCH], and [MKR - >], then /MARKER ->REF LVL/.

If a 10-kHz filter with a typical shape factor of 15:1 is used, the filter will have a bandwidth of 150-kHz at the 60-dB point. The half-bandwidth (75-kHz) is narrower than the frequency separation, so the input signals will be resolved.

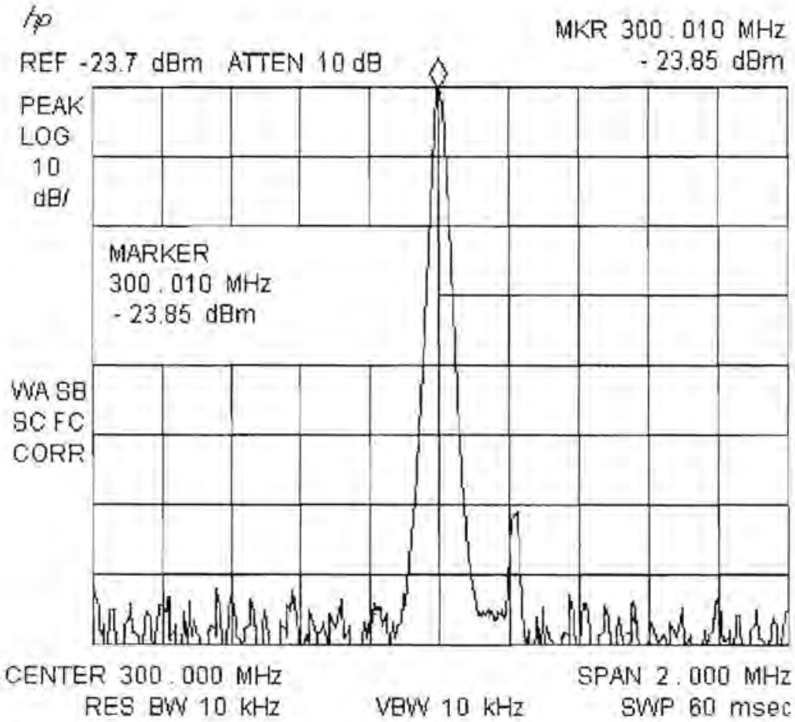


Figure B.4. Signal resolution with a 10-kHz resolution bandwidth

If a 30-kHz filter is used, the 60-dB bandwidth will be 450-kHz. Since the half-bandwidth (225-kHz) is wider than the frequency separation, the signals most likely will not be resolved. See Figure B.5. (To determine resolution capability for intermediate values of amplitude level differences, consider the filter skirts between the 3-dB and 60-dB points to be approximately straight. In this case, we simply used the 60-dB value.)

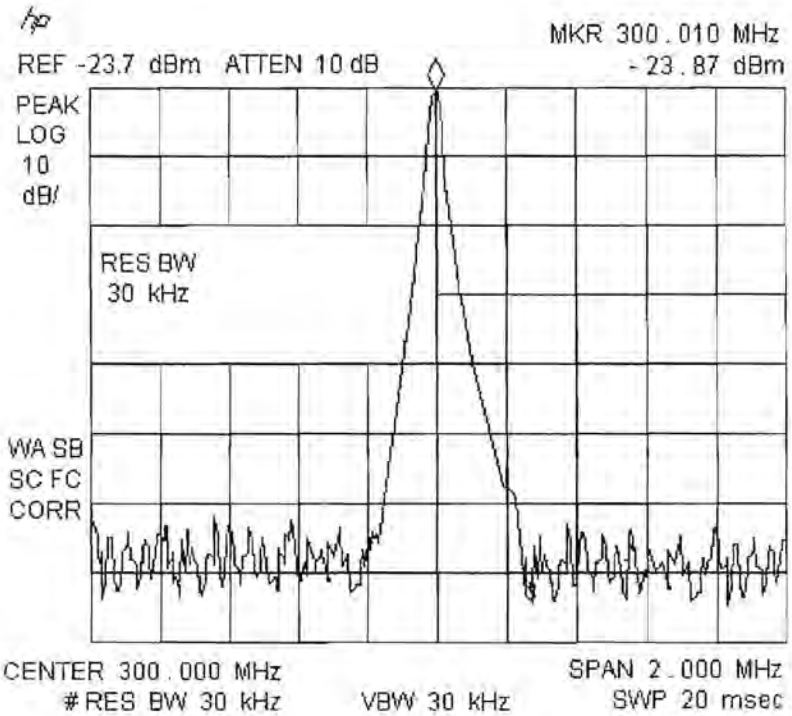


Figure B.5. Signal resolution with a 30-kHz resolution bandwidth

B.3 Measuring low-level signals using Attenuation, Video Bandwidth and Video Averaging

Spectrum analyser sensitivity is the ability to measure low-level signals. It is limited by the noise generated inside the spectrum analyser. The spectrum analyser input attenuator and bandwidth settings affect the sensitivity by changing the signal-to-noise ratio.

The attenuator affects the level of a signal passing through the instrument, whereas the bandwidth settings affects the level of internal noise without affecting the signal. If a signal is very close to the noise floor, reducing input attenuation brings the signal out of the noise. Reducing the attenuation to 0-dB maximises signal power in the spectrum analyser.

When, after adjusting the attenuation and resolution bandwidth, a signal is still near the noise, visibility may be improved by using video-bandwidth and video-averaging functions. Video-bandwidth function allows for the video-filter control, which is useful for noise measurements and observation of low-level signals near the noise floor. The video filter is a post-detection low-pass filter that evens the displayed trace.

When signal responses near the noise level of the spectrum analyser are visibly masked by the noise, the video-filter may be narrowed to smooth this noise and improve visibility of low-level signals. (reducing video bandwidth requires slower sweep times to keep the instrument calibrated.)

Video averaging is a digital process in which each trace point is averaged with the previous trace-point average. Selecting video averaging changes the detection mode from peak mode to sample mode. The result observed is a sudden decrease in the displayed noise level. The sample mode displays the instantaneous value of the signal at the end of the time or frequency interval represented by each display point, rather than the value of the peak during the interval. Sample mode is not used to measure signal amplitudes accurately because it may be incapable to find the true peak of the signal.

Video averaging clarifies low-level signals in wide bandwidths by averaging the signal and/or the noise. During averaging, the current sample appears at the left side of the graticule. Changes in active function settings, such as the centre frequency or reference level, will restart the sampling. The sampling will also restart if video averaging is turned off and on again. Once the set number of sweeps has been completed, the spectrum analyser continues to provide a running average based on this set number of sweeps.