N-Way (Star Junction) Transmitter Combining

Theory and Practice

With Examples using Motorola T1500 Cavities

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1. Overview

Transmitter combining is often used where 2 or more repeaters share a common site and it isn't practical to have an individual antenna for each repeater. In this case, two or more transmitters share a common antenna and the receivers use a companion master receive antenna system.

With a dedicated receive antenna, proper filtering, and a high quality LNA, it is possible to enhance site receive performance for all of the combined systems. This can have benefits where many users are equipped with portables and it helps to provide balanced talk-in and talk-out coverage despite the slightly higher insertion loss of a combiner compared to a duplexer.

There are several methods of combining transmitters. In the UHF band, where transmitter spacing of 400 kHz or more is common, filter-ferrite, N-way combiners are often used. These are sometimes referred to as star junction combiners. *(See Appendix 1)*

An important consideration in designing an N-way combiner is determining the proper length of the cables to match impedances for maximum power transfer to the antenna.

This document will present the theory of the N-way combining method and will provide an example using Motorola T1500 cavities.

In Figure 1, a two branch combiner is shown using T1500 cavities. E_g is the signal from a transmitter in question, Z_L is the antenna for that transmitter, and Z' is the equivalent impedance of the remainder of the combining network.
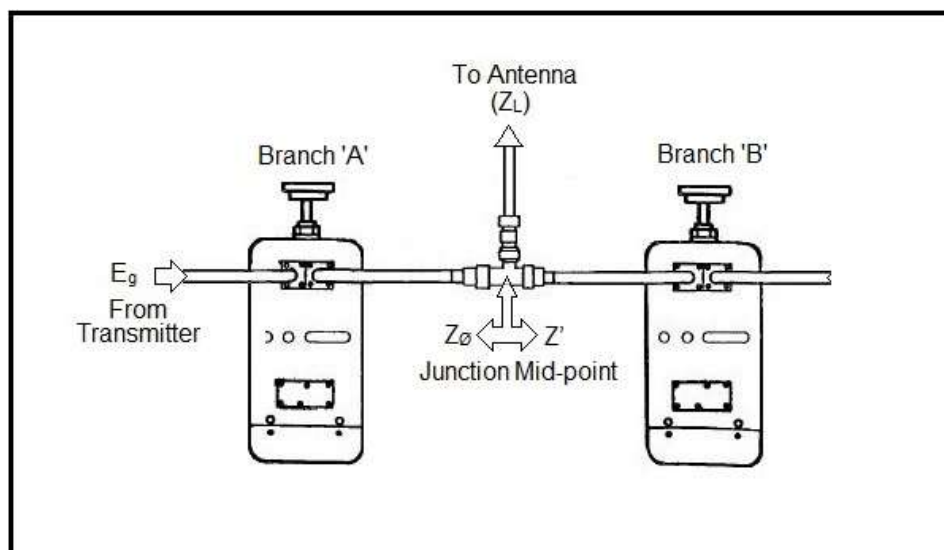

Figure 1

The goal of a combiner is to transfer the energy from each transmitter to the antenna with as little loss as possible. Intuitively, this will take place if Z' is very high at the junction mid-point *(See Appendix 2)*

2.   The Need for Critical Length Cables

When cavities are employed in a combiner, each cavity will approach $Z_\emptyset$ at its resonant frequency - i.e. the frequency of the transmitter connected to that branch. Ideally, each of the other branches will appear to be a high impedance at the junction mid-point or common point where each branch meets the antenna port. Electrical lengths are measured from that point.

Transmission line theory shows that an impedance connected to the end of a transmission line will be exactly repeated at intervals of 18Ø°. At any other length of transmission line, the impedance will be 'transformed' to some other value.

In order to cause the input impedance of a cavity to appear as a high impedance at the frequency of the other branch(s), the length of the transmission line used to connect the cavity to the junction is critical.

The common belief that these cables are one quarter wave long is based on the assumption that the off-resonance input impedance of a bandpass cavity is $\emptyset \pm j\emptyset\Omega$. The actual cavity input impedance must be known so that the correct length of transmission line can be determined.

Since cavity impedance is not published by cavity manufacturers, the cavity's input impedance could be calculated providing that the insertion loss and coupling factor between the loop and cavity were known. (*See Appendix 3*) Unfortunately, those parameters are also usually not available

Accordingly, a cavity's input impedance may be measured. This measurement can performed in a number of ways (*e.g. impedance bridge, slotted line*) but today it is easy with a network analyzer.

3.   The Network Analyzer Impedance Display

Before measuring the cavity's impedance on the network analyzer, a quick review of Smith chart basics will facilitate understanding the network analyzer's display.

The Smith chart uses polar format to graph complex impedance and other related parameters. (*See Appendix 4*) It has many applications but it is not necessary to master all of them to understand a Smith chart display on a network analyzer.
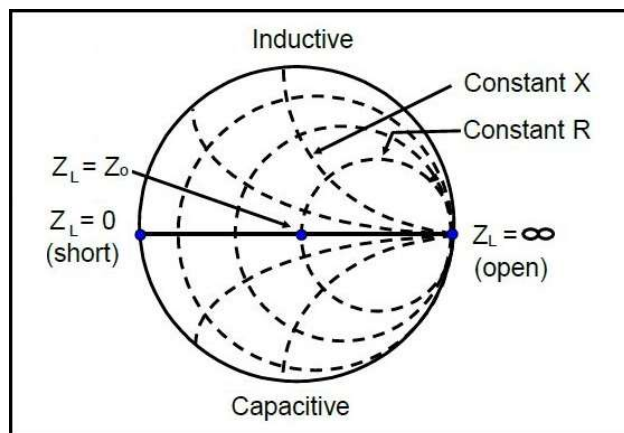
As shown in Figure 2, the central axis of the Smith chart is a horizontal line representing resistance using a log scale with zero on the left and infinity on the right.

The outer boundary of the Smith chart is a circle that circumscribes the resistance line. The center of the circle and the center of the resistance line coincide. That mid-point on the resistance line represents the normalized $Z_\emptyset$, in this case 50 ohms.

Figure 2

Any point located on this line represents pure resistance (i.e. R ±jØ). The outer circle (also known as the reactance axis) because it intersects the resistance line at zero ohms, has a value of R=Ø. Thus any point that lies on this circle is pure reactance (i.e. Ø ± jX).

The chart also contains resistance circles and reactance arcs. An impedance point (R ±jX) is plotted on the chart at the intersection of the resistance circle that equals R and the reactance arc that equals X. Points above the resistance line represent inductive reactance and points below the resistance line represent capacitive reactance.

4. Measuring the Cavity Input Impedance
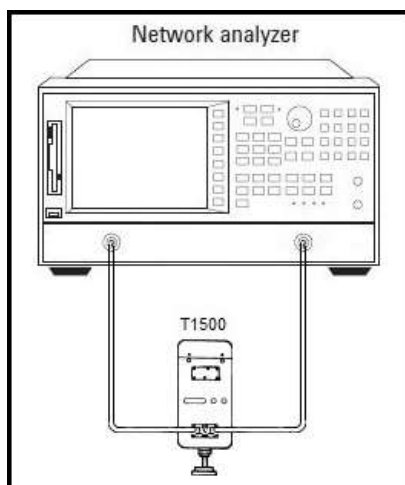
Figure 3

Figure 3 illustrates the method used with a vector network analyzer to measure the transmission (frequency response), reflection (return loss), and input impedance of a cavity. Shown being measured is a Motorola T1500 bandpass cavity.

To provide accurate impedance readings, the test setup must take into account the electrical length of any adapters used and the coaxial portion of the coupling loop assemblies.

In other words, the measurement reference plane (*See Appendix5*) must coincide with the point where the loop assembly transitions from coaxial construction to inductance.

When the vector network analyzer is configured to make reflection measurements (S11), the results can be plotted as a log-magnitude display of return loss vs. frequency as shown in Figure 4 or in Smith format showing impedance vs. frequency as illustrated in Figure 5.
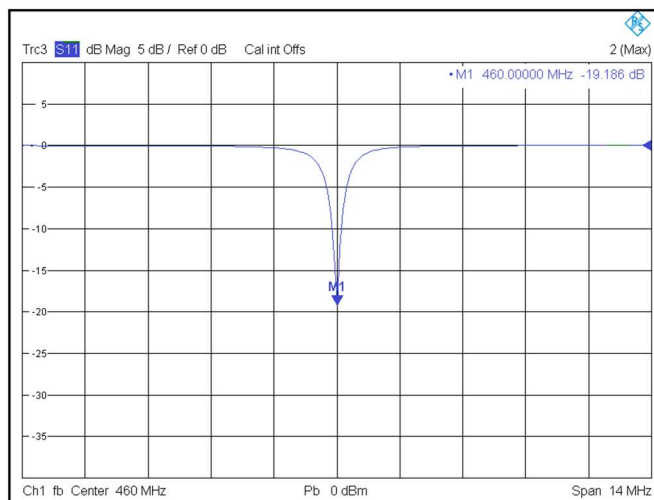
Figure 4
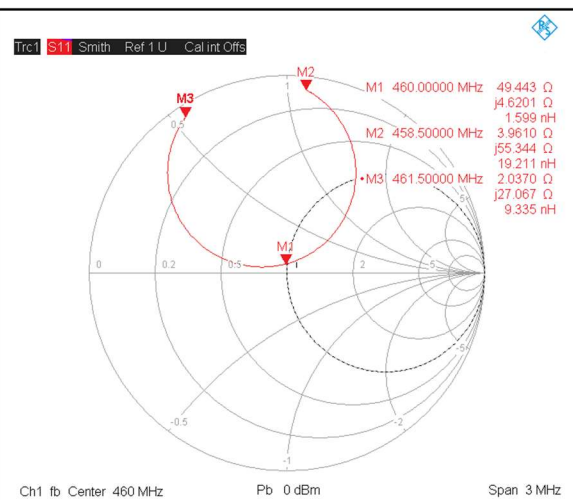
Figure 5

Note that the return loss of the T1500 bandpass cavity is symmetrical about a vertical axis line that runs through the point representing the tuned resonant frequency. Therefore, the return loss is the same at points equally removed in frequency from the resonant point.

Thus it follows that on the Smith format display, points equally removed in frequency from the resonant point will also have identical return loss because the same S11 data is simply being presented in a different format.

The distance from the center of the Smith chart to any impedance point is directly proportional to the magnitude of the reflection coefficient (Γ)[1]. Since VSWR = $(1 + \Gamma)/(1 - \Gamma)$, circles concentric about $Z_\emptyset$ will represent constant VSWR (and return loss). This means that impedance points with equal return loss must lie on the same constant VSWR circle.

When plotted in Smith format, the cavity reflection response (impedance locus) appears as a circle[2]. With the analyzer span of ±1.5 MHz used in Figure 5, the circle is incomplete. A wider span would cause the ends of the response curve to converge toward being a complete circle.

If the circle were complete, it would be internally tangent to the outer circle of the Smith chart (the reactance axis where R =$\emptyset$) and that point would be the self-reactance of the loop. ($Z = \emptyset \pm jX$) With an incomplete circle, that point can be determined by taking the impedance values of the end points of the response curve and finding the mid-point.

This is consistent with the cavity input impedance equation shown in Appendix 3.

The reactance of the loop alone can be measured with a network analyzer as shown in Figure 6.

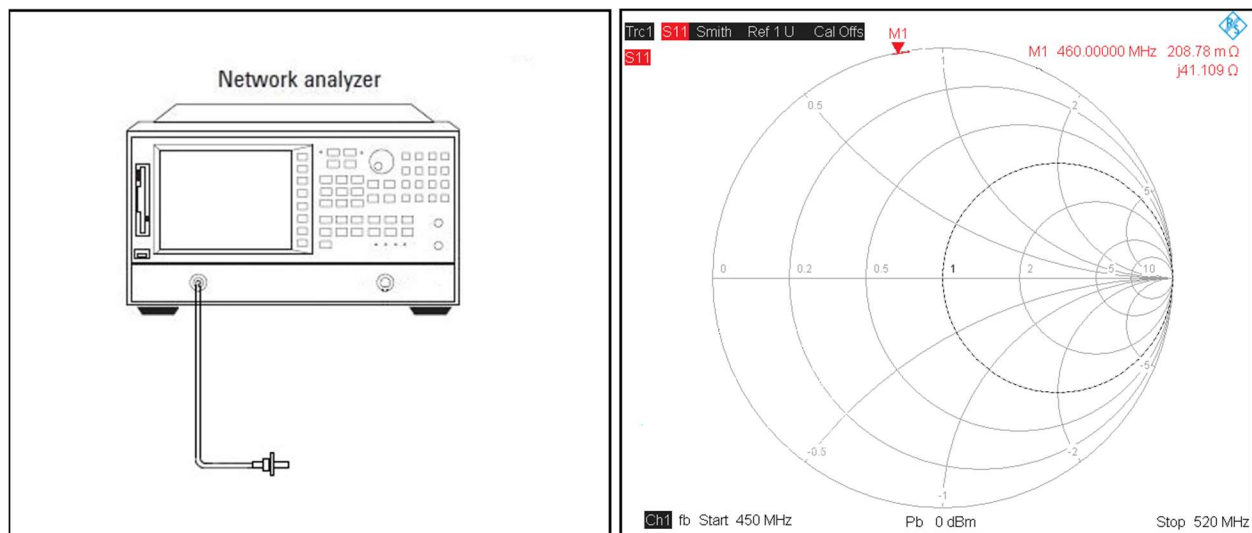|	
Network analyzer	Trc1 S11 Smith Ref 1 U Cal Offs M1
	M1 460.00000 MHz 208.78 mΩ j41.109 Ω
	Ch1 fb Start 450 MHz Pb 0 dBm Stop 520 MHz

Figure 6   Figure 7

In Figure 7, measurement of a T1500 bandpass loop shows a reactance of 41 ohms at 46$\emptyset$ MHz, the exact value stated by Motorola[3] (In impedance terminology $Z = \emptyset + j41\Omega$)
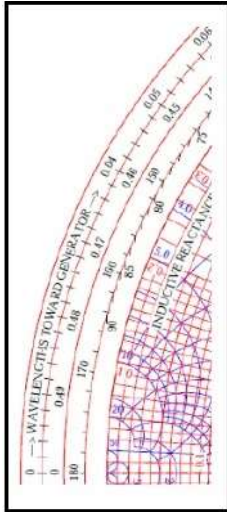
5. The Smith Chart and Rotation
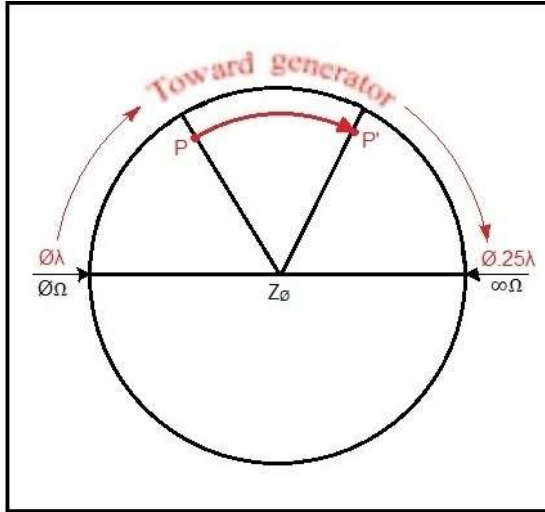

Figure 8

Figure 9

On a printed Smith chart form, there are several calibrated scales surrounding the perimeter of the chart itself. (See Appendix 4) The outermost scale shown in Figure 8, wavelengths toward generator, is useful for impedance transformation[4].

Note that the scale starts at Ø wavelengths at the Ø ohm point on the resistance axis and continues around to Ø.25 wavelengths at the infinite end of the resistance line.

As with transmission line theory, this shows that a 1/4 wavelength transmission line will transform a short to an open.[5]

In figure 9, an impedance is plotted at point P. If a length 'L' of transmission line is inserted between the impedance and the measurement point, the impedance will be transformed to some other value. (*For the purposes of this document, the short lengths of transmission line used are assumed to be lossless*) Using the Smith Chart to show the transformation, first a radial line is drawn from the center of the chart through point P continuing until it intersects the wavelengths toward generator scale.

From that point, continue around the wavelengths toward generator scale by an amount equal to the wavelength of transmission line L. At that point, if a radial line is drawn back to the center of the chart, P', the transformed impedance of point P, will lie on this new line.

Since a length of transmission line transforms impedance but not VSWR, when point P is transformed to point P' the new point must lie where the new radial line intersects the constant VSWR circle that point P lies on. Consequently, point P appears to "rotate" along the constant VSWR circle.[6]
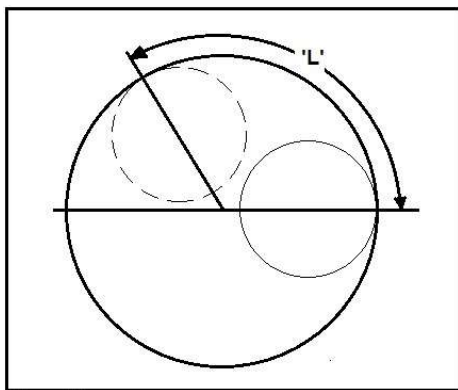
Figure 10

This will also be true for all the points of an impedance vs. frequency curve so that the entire curve is 'rotated' to a new position.

An example of impedance rotation is shown in Figure 10. Note that the length of transmission line in this example has been chosen to rotate the impedance circle so that it is tangent to the outer circle at the 0.25λ location.

The result of this rotation is that the points on the impedance curve proximate to the point of tangency are translated to high impedance.

6.   Determining the Critical Cable Length

It was shown previously that the point of tangency occurs for frequencies equally far removed from the cavity's resonant frequency and this point is the impedance that must be translated to a high impedance at the combining junction.
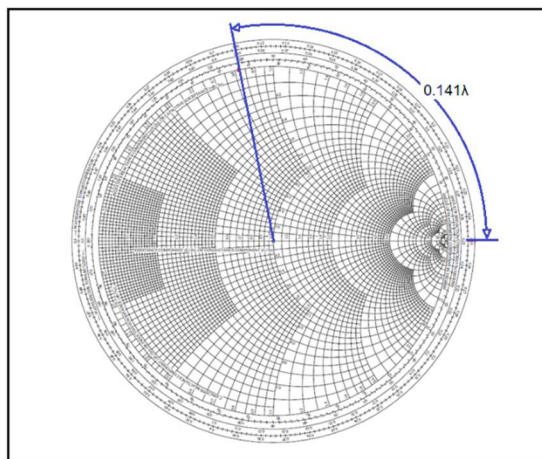
The length of transmission line to translate the cavity to a high impedance can be determined by plotting the cavity's input impedance on the Smith chart and rotating the impedance circle by the length 'L' which is the electrical length of the line required.

For a T1500 tuned to 460 MHz, the cavity's input impedance has been determined to be $\emptyset + j41\Omega$. In Figure 11, that point has been plotted on a Smith chart and the 'wavelengths toward generator' scale has been used to determine that the electrical length necessary is 0.141λ[7].

Figure 11

The wavelength at 460 MHz is 25.67" (in air). A cable 0.141λ would have an electrical length of 3.6". As this is not physically practical, a cable having an electrical length of 0.641λ will perform the same impedance transformation and has an electrical length of 16.45".
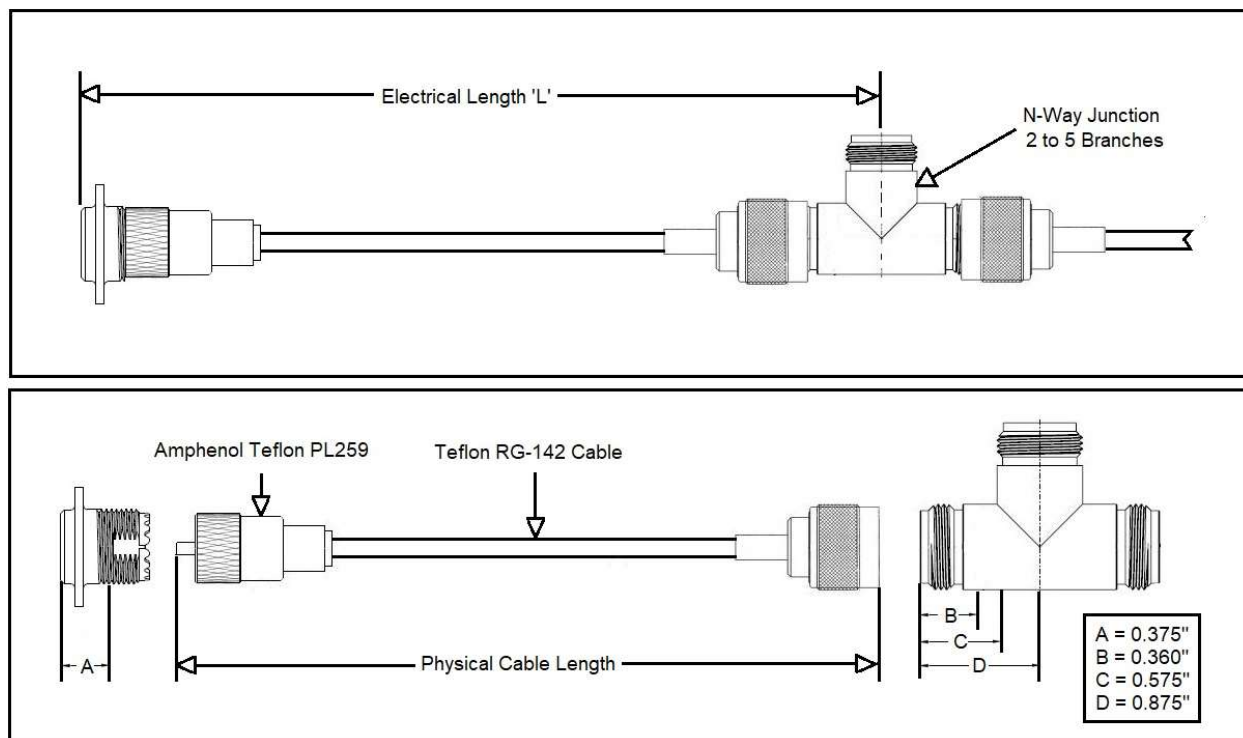
Figure 12

The top half of Figure 12 illustrates that the electrical length of cable 'L' includes the coaxial portion of the coupling loop assembly and the length to the mid-point of the junction.
The three physical lengths are shown the lower half of Figure 12 for the case where the combining junction is a type N tee. Using the known physical lengths of the SO-239 and the type N tee, the electrical lengths of each can be determined. Subtracting them from the electrical length 'L' will provide the electrical length of the cable assembly which can then be converted to the correct physical length.

In Motorola's T1500 documentation[8], all cable lengths are measured from the tip of the PL-259 connector. Therefore, the length of the tip (Ø.375") must be subtracted from the length of the SO-239 to determine the remaining length up to the point where coaxial construction ends and the loop inductance begins. That remaining portion is Ø.375" (A). This portion has Teflon dielectric so the length in air is Ø.54". [Ø.375 ÷ Ø.695 = Ø.54]

For a type N tee, the length to the mid-point has two portions. First, there is a section in air from the N connector reference plane to the start of the Teflon dielectric (C) – (B) which is Ø.215" [Ø.575 - Ø.36Ø]

Second, there is a section with Teflon dielectric. The physical length of 1/2 of the tee is Ø.875"(D), the Teflon start is at Ø.575, Ø.875 - Ø.575 = Ø.3". That electrical length is Ø.432" [Ø.3 ÷ Ø.695 = Ø.432]

Thus L + Ø.215 + Ø.432 + Ø.54 = 16.45". Solving for L = 15.263" in air. Converting the length L in air to Teflon gives 10.61" for an RG-142 cable assembly from the UHF tip to N reference plane.

7.   Implementing a Practical Combiner

Figure 13

With the critical cable length determined, a practical combiner can be constructed as illustrated in Figure 13.

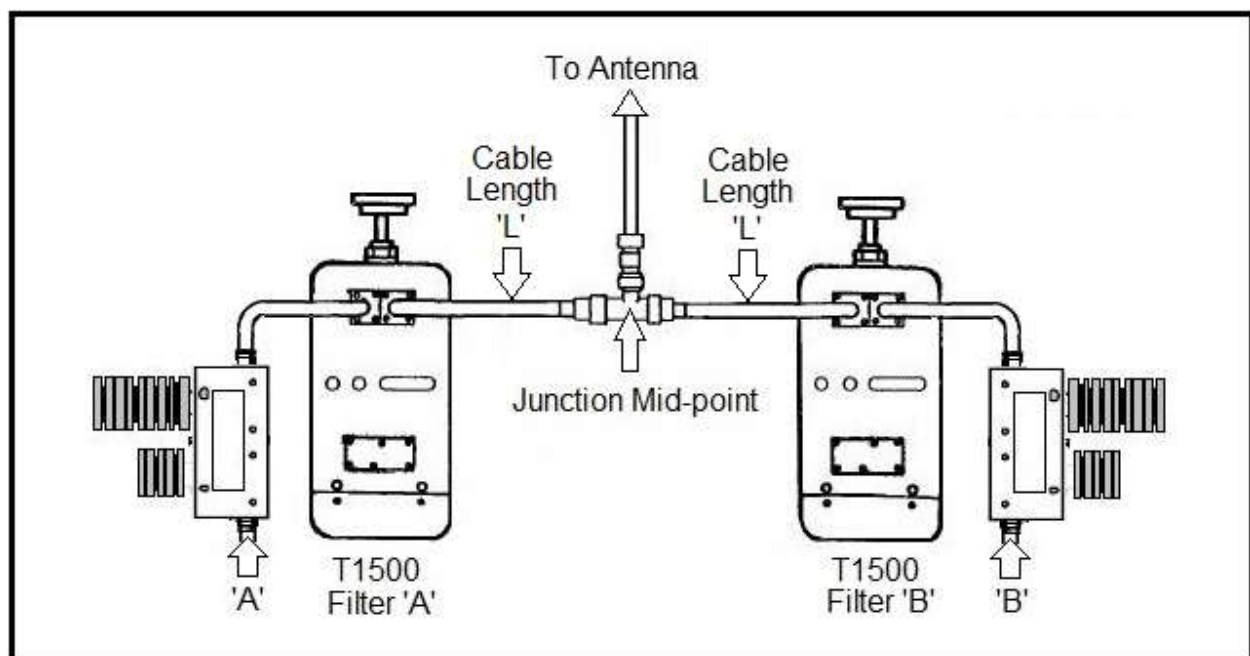A practical combiner implementation includes isolators in each branch. These prevent the attenuated signals of other branches from reaching the transmitter's PA where they could mix and create IM products. The isolators also protect the transmitters from any reflected power that could originate at the antenna port due to a miss-match or fault in the antenna system.

Only the cables between the combining junction and each cavity must be critical lengths. Neither the length of the cavity to isolator cables nor the isolator to transmitter cables is critical but these should be as short as practical to minimize losses. These cables must be double shielded such as RG-142.

Determining if a combining solution is appropriate for a specific application requires consideration of the frequency separation between the transmitters to be combined and the branch insertion loss. These are related as will be shown.

The branch loss of an N-way cavity-ferrite combiner is comprised of the cavity loss, bridging loss, isolator loss, and cable losses.

Cavity loss is based on the loop setting. In the example case using the Motorola T1500 cavities, the loops are set to the 1 dB position.

Bridging loss is caused by the other cavities connected to the junction. If the critical length cables were able to transform the other cavity branches to ∞ impedance, there would be virtually no bridging loss. However, only frequencies very far away from cavity resonance are transformed close to ∞. Frequencies nearest to $Z_\emptyset$ remain at very nearly the same impedance regardless of the length of the cable.

Intermediate frequencies on the cavity's impedance curve between the cavity's resonant point and the point of tangency with the R = Ø outer circle on the Smith chart will be transformed toward high impedance. The closer a point is to R = Ø, the closer its transformed impedance will be to ∞.

Logically, the bridging loss caused by a branch will depend on how high its transformed impedance is. When the impedance transformation is performed with a network analyzer, the instrument's markers will show the before and after impedance of intermediate frequencies.

For example, Figure 14 shows the input impedance vs frequency of a T1500 cavity plotted from 458.5 MHz to 461.5 MHz (C.F. = 46Ø.Ø MHz). The cavity's input impedance of Ø + j41Ω has been identified with marker 4 and a radial line has been drawn from that point to $Z_\emptyset$.

In Figure 15, the impedance transformation has been performed by inserting a transmission line length of Ø.141λ so that the analyzer display shows the desired rotation. The electrical length

can be inserted in the analyzer or externally using a mechanical device known as a 'line stretcher' or 'phase shifter' such as the FXR/Microlab SR-15N.
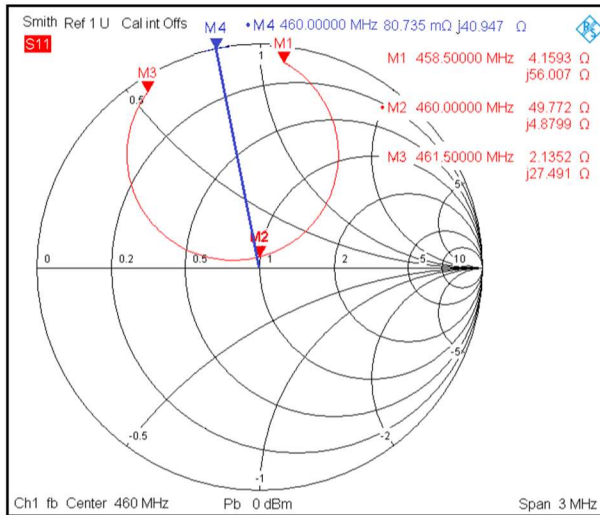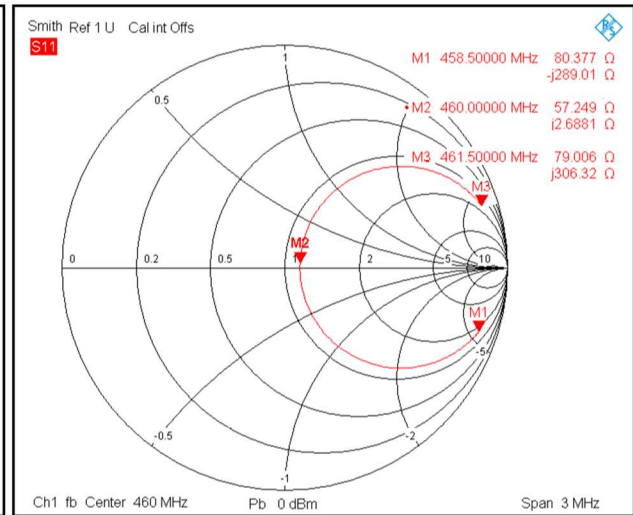
Figure 14   Figure 15

In Figure 15, note that the 460 MHz resonant point remains close to $Z_\emptyset$ and the cavity's input impedance, $\emptyset + j41\Omega$, has been transformed to ∞. The input impedance at 458.5 MHz has been transformed from 56.15Ω $\angle86°$ to 299Ω $\angle$-74.5° and the input impedance at 461.5 MHz has been transformed from 27.58Ω $\angle78.2°$ to 316Ω $\angle75.5°$. *(See Appendix 6)*

Although the transformed impedances are relatively high in the 300Ω range, when bridged across $Z_\emptyset$ at the junction, a branch with this impedance will absorb some power and thus cause bridging loss. Clearly, frequencies closer to $Z_\emptyset$ will be transformed to impedances that will cause higher bridging loss.
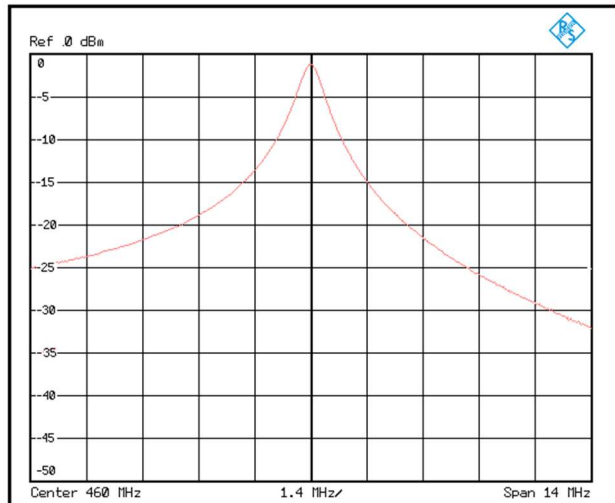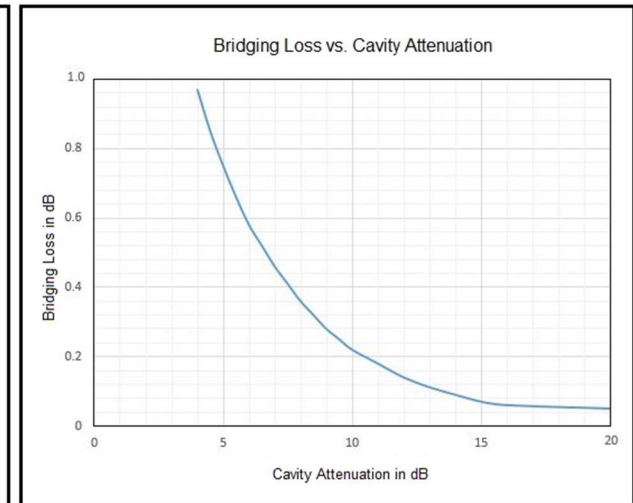
Figure 16   Figure 17

The loss for each branch could be computed:

$$Bridging\ Loss = -10log_{10}(\frac{1}{N}\sum_{i=2}^{N+1} S_{i1})$$

However, a graphical method facilitates the process.

In order to determine the bridging loss of a branch, the attenuation for the cavity must be obtained from selectivity curves using the frequency separation between the branch being considered and each of the other branches. The selectivity curve for the Motorola T-1500[9] is shown in Figure 16. A more accurate value can be obtained by measuring the actual loss with a network analyzer.

Bridging loss data from Motorola[10] has been graphed in Figure 17. For a given branch, the bridging loss must be calculated for each of the other branches. These are added to give the total bridging loss for the branch being studied.

Adding the bridging loss, isolator loss, and cable loss yields the total branch loss. Typical branch loss for a two port combiner using T1500 cavities is shown in Figure 18.

If closer transmitter spacing is necessary, the branch loss can be reduced through the use of cavities with higher Q as shown in figure 19. *(See Appendix 7)*

Figure 18

Figure 19

For transmitter to transmitter frequency spacing of 500 kHz or more, a combiner using T1500 cavities should have a branch insertion loss of 3 dB or less. If closer spacing, such as 300 kHz is necessary, 10" cavities can be used to keep the branch insertion loss in the 3 dB range.

9. N-Way Combiner Alignment

Combiner manufacturers generally provide alignment instructions specific to their product. Those should be followed in most cases, however, for an unknown or constructed combiner, the following procedure is recommended.

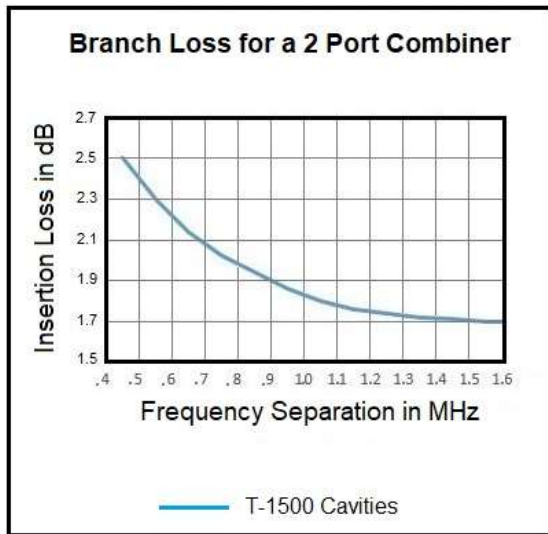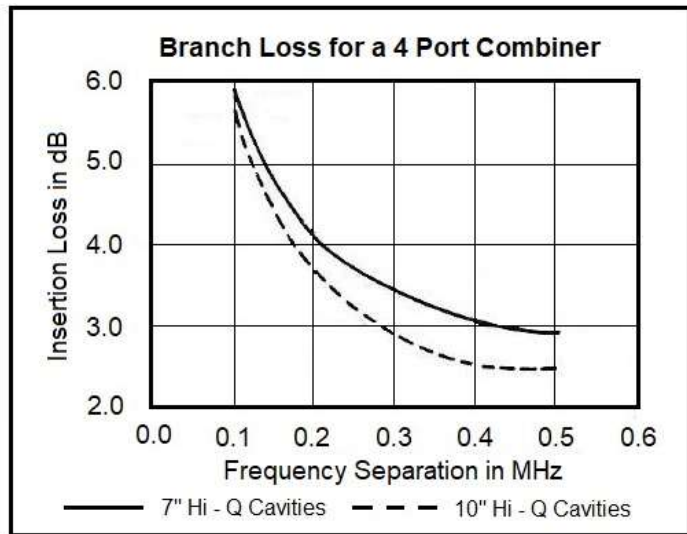Start with the isolators and perform a complete alignment of each using the manufacturer's instructions. EMR has a document that describes the process using a network analyzer.

http://www.repeater-builder.com/antenna/pdf/emr-dual-isolator-tuning-instructions.pdf

Label each isolator with the combiner port number and frequency. Next tune each cavity using the setup shown in Figure 3 on page 6. Label each cavity with its port number and frequency. Now assemble the complete combiner.

Connect a network analyzer port one to the first transmitter port and the network analyzer port two to the combiner antenna port. Measure S21 at that port's frequency. The cavity response curve should be centered and the branch insertion loss should not exceed the value calculated for that port. Repeat this procedure for each combiner port.
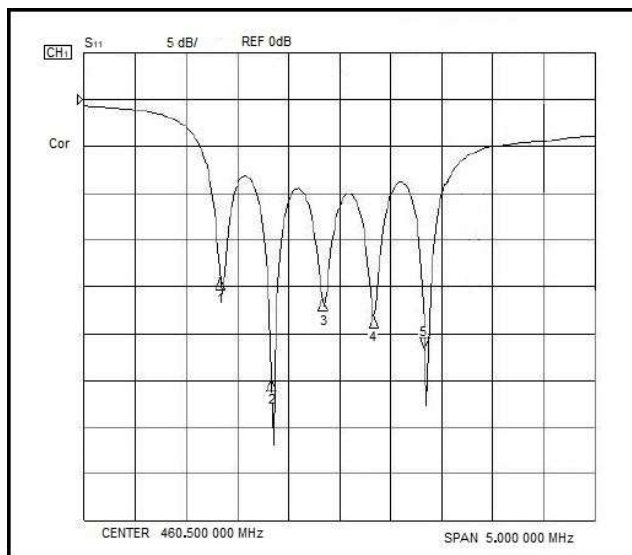
Now configure the analyzer with a span wide enough to show all the combiner frequencies and activate a marker at each transmit frequency. Perform an S22 measurement of the combiner's output return loss (ORL).

The analyzer display should be similar to Figure 20 which shows the ORL for a five channel UHF combiner.

If each cavity's bandpass is correctly aligned, the return loss dips should precisely coincide with the markers. If any do not, connect the analyzer's port one cable to that frequency's isolator, configure the display to show S21

Figure 20

and S22 concurrently, and optimize the cavity alignment. It is desirable that the ORL for each port is greater than 13 dB and that the ports are relatively the same across the range of frequencies.

Next, with the combiner installed and the transmitters connected, attach a wattmeter and load to the antenna port. Key each transmitter individually and observe the forward power. It should match the power calculated for that port based on the transmitter's output power and the branch loss. (e.g. a 50 watt transmitter connected to a port with 3 dB branch loss should read 25 watts)

It is possible at this point to make a final minor tuning adjustment with the transmitter keyed. It is preferable to reduce the output of high power stations (e.g. 100 watt) to about 25 watts.

Finally, when installing a system utilizing any transmitter combiner, it is essential to follow best practice. With systems employing transmit combining, there are multiple high level transmit signals present concurrently in the transmit transmission line and antenna. This increases the possibility of IM products.

Only the finest components should be used, preferably ones rated for low PIM. There should be no braided cable anywhere beyond the common junction. Low PIM installation practices should be followed. 7/16" DIN connectors are preferred and all connectors should be properly installed and torqued to manufacturer's specifications.

10.   Summary

The optimum type of combiner for a particular application depends on several factors. The N-way, cavity-ferrite combiner is ideal for applications of two to five transmitters if there is sufficient separation of the frequencies being combined.

Transmitter to transmitter frequency spacing trade-offs have been presented.

Design principles such as the determination of the input impedance of a cavity and critical length cables have been covered.

Application considerations have been presented so that an appropriate combiner may be selected or constructed.

References are provided to enable further research of critical points.

Appendix 1   Examples of N-Way Junctions

Figure 21

Figure 20 shows a three way junction with SO239 UHF connectors for the three transmitter branches and a type N connector for the common antenna port.

Figure 22

Figure 21 shows a four way junction with all type N-F connectors.

Figure 23

Figure 22 shows a five way junction with all type N-F connectors.

Appendix 2   Transmitter Combining Theory

In figure 1 on page 4, energy from the transmitter connected to branch A passes through the branch A cavity filter which is tuned to that transmitter's frequency and arrives at the junction mid-point.

From the mid-point, the energy 'sees' two possible paths. The path to a well matched antenna will appear to be $Z_{\emptyset}$. The path through branch B will be some other impedance. Ideally, all of the energy will continue on to the antenna. The following discussion illustrates how the branch B impedance is optimized toward that objective.
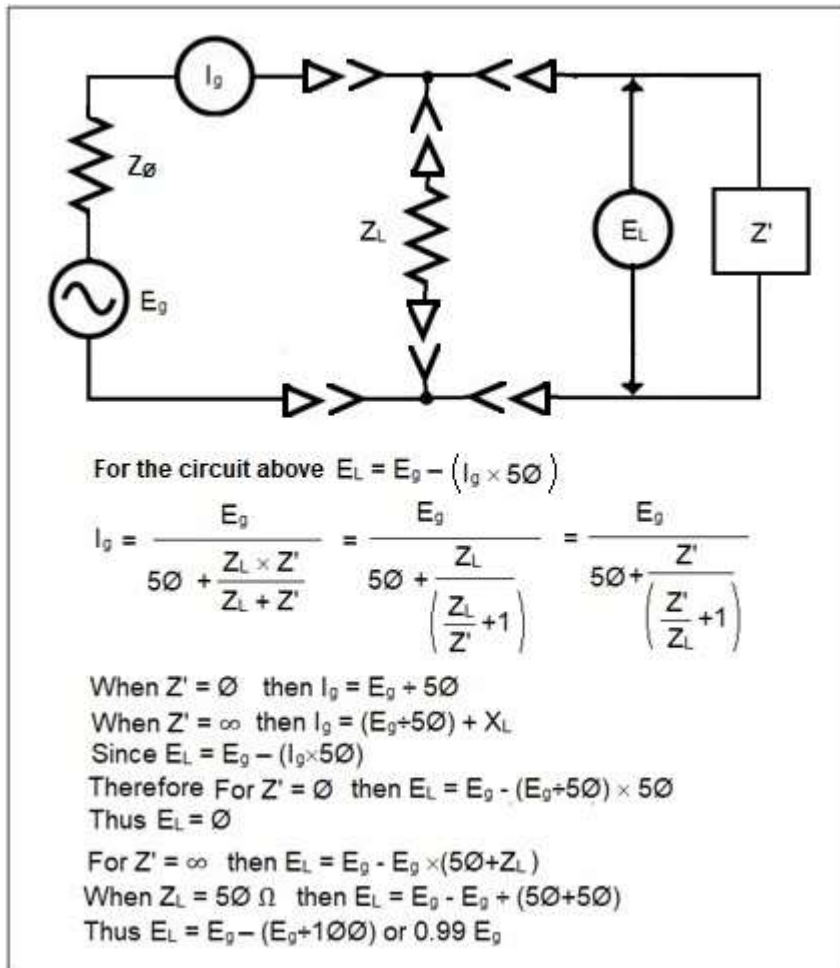
An equivalent circuit for the combining network[11] is illustrated at left.

As with Figure 1, E_g is the signal from a transmitter which has a characteristic impedance $Z_{\emptyset} = 50\Omega$, Z_L is the antenna for that transmitter, and Z' is the equivalent impedance of the remainder of the combining network.

The equation illustrates how the impedance of the rest of the combining network affects the voltage (and corresponding energy transfer) across the desired load (Z_L).

For the circuit above $E_L = E_g - \left(I_g \times 5\emptyset \right)$

$$I_g = \frac{E_g}{5\emptyset + \dfrac{Z_L \times Z'}{Z_L + Z'}} = \frac{E_g}{5\emptyset + \dfrac{Z_L}{\left(\dfrac{Z_L}{Z'} + 1 \right)}} = \frac{E_g}{5\emptyset + \dfrac{Z'}{\left(\dfrac{Z'}{Z_L} + 1 \right)}}$$

When Z' = Ø   then $I_g = E_g \div 5\emptyset$
When Z' = ∞   then $I_g = (E_g \div 5\emptyset) + X_L$
Since $E_L = E_g - (I_g \times 5\emptyset)$
Therefore For Z' = Ø   then $E_L = E_g - (E_g \div 5\emptyset) \times 5\emptyset$
Thus $E_L = \emptyset$

For Z' = ∞   then $E_L = E_g - E_g \times (5\emptyset + Z_L)$
When $Z_L = 5\emptyset \, \Omega$ then $E_L = E_g - E_g \div (5\emptyset + 5\emptyset)$
Thus $E_L = E_g - (E_g \div 1\emptyset\emptyset)$ or $0.99 E_g$

Thus it has been shown that maximum power is delivered to the load when Z' is made as large as possible.

Appendix 3 The Input Impedance Equation for a Cavity

The input impedance equation for a cavity[12] is:

$$Z = jX_L + \frac{Z_{\emptyset}\beta}{1 + j2(\beta + 1)\, Q_L\, \delta}$$

Where

Z = Input Impedance of the Cavity
X_L = Reactance of the Coupling Loop
$Z_{\emptyset}$ = Characteristic Impedance of the Transmission Line
β = Coupling Factor Between the Loop and the Cavity
f = measurement frequency (where the equation is solved)
$f_{\emptyset}$ = cavity resonant frequency
Δf = Difference between 3 dB down points on the cavity response curve
Q_L = The Loaded Q of the cavity = $f_{\emptyset} \div \Delta f$
$\delta = (f - f_{\emptyset}) \div f_{\emptyset}$

On page 7 it was stated that: "If the circle were complete, it would be internally tangent to the outer circle of the Smith chart (the reactance axis where R =Ø) and that point would be the self-reactance of the loop."
$Z = \emptyset \pm jX$

This is consistent with the cavity input impedance equation.

In the equation, δ is principally determined by the offset between the tuned resonant frequency $f_{\emptyset}$ and the measurement frequency f.
$\delta = (f - f_{\emptyset}) \div f_{\emptyset}$

As the offset between the resonant frequency $f_{\emptyset}$ and the measurement frequency f is increased, the right side of the equation approaches $Z = \emptyset + j\emptyset$.

Then the cavity Z is a pure reactance and that reactance is the reactance of the loop alone
$Z_{in}(f) = \emptyset + jX_L)$[13]

NORMALIZED IMPEDANCE AND ADMITTANCE COORDINATES

RADIALLY SCALED PARAMETERS
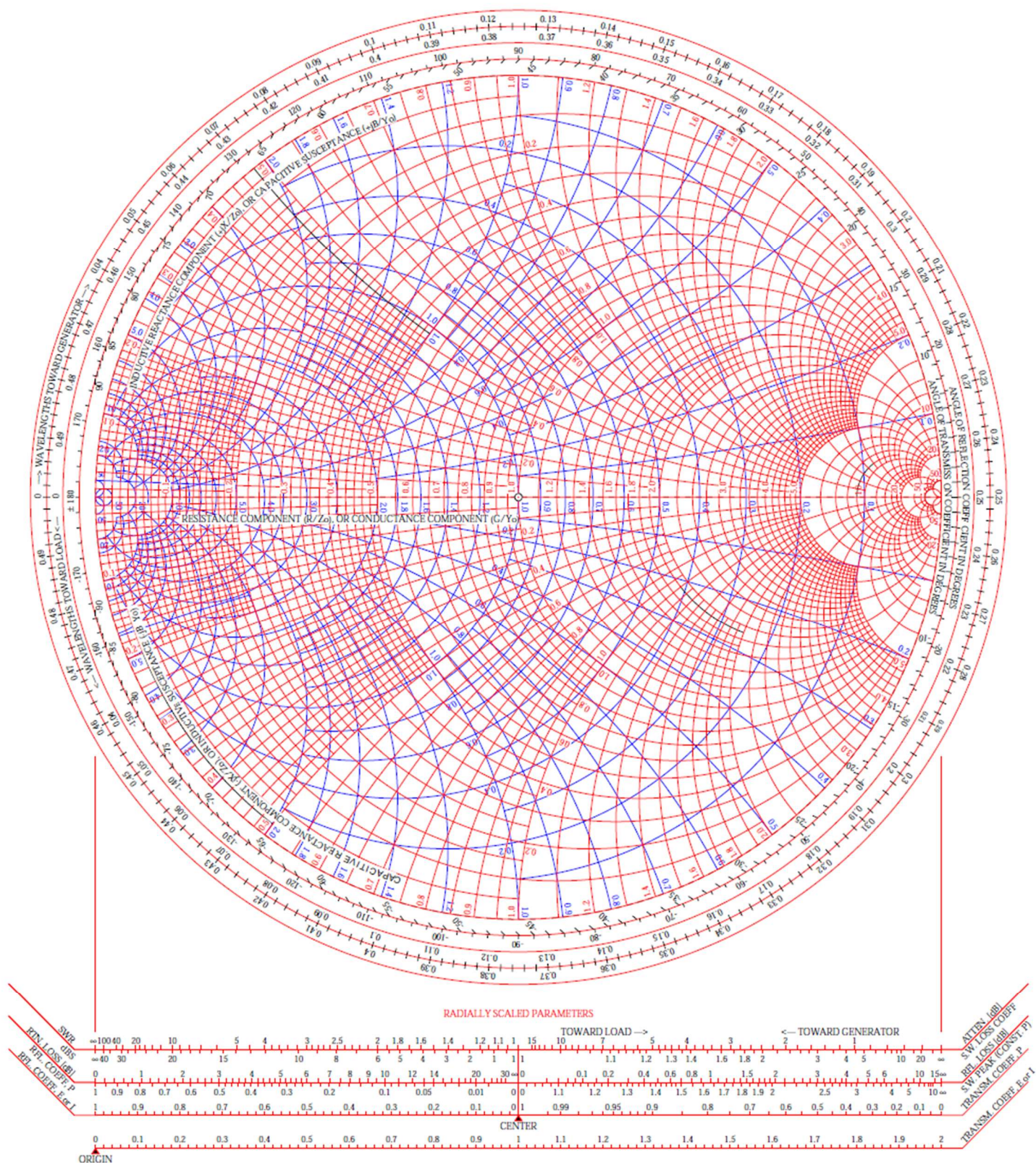
Appendix 5 – Reference Planes

One concept that must be mastered when measuring impedance is the issue of reference planes.

Every coaxial connector specification identifies the reference plane for that connector type. This is the physical point that becomes the transition point between connectors of a mated pair.
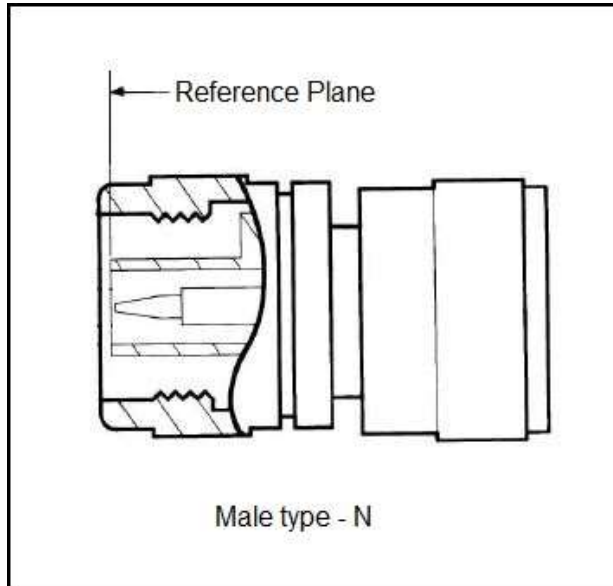
Figure 23

The reference plane is also the electrical point of transition between connectors of a mated pair. Figure 23 shows the reference plane of a male type – N connector.

Before taking measurements with a network analyzer, the user executes a calibration at the 'working' end of the test port extension cables using ether a mechanical calibration kit or an electronic calibration module.

The network analyzer uses stored calibration coefficients combined with the calibration measurements to establish a calibration reference plane at the physical reference plane of the test port connector.

The calibration reference plane becomes the measurement reference plane. All measurements taken by the network analyzer are corrected to this physical and electrical point. An impedance measurement is accurate at that point only if the impedance entity itself is located there. If the actual impedance is located beyond that point – physically contained within the DUT – the impedance 'seen' by the network analyzer will have been transformed due to some length of transmission line that exists between the measurement reference plane and the actual impedance.

The objective when measuring the input impedance of a cavity for the purpose of determining the critical length of a cable is to measure the impedance of the coupling loop at the loop location, not the impedance seen some electrical distance away[See Note]. Fortunately, the network analyzer can relocate the measurement reference plane to the correct point through the use of electrical delay or port extension (offset).

In order to configure the network analyzer to relocate the measurement reference plane, it is necessary to determine the electrical length between the calibration reference plane and the T1500 loop. Obviously, this must include the electrical length of the UG83-U adapter. In addition, the loop inductance starts where the SO-239 transitions from coaxial construction to the loop. (See Figure 24)
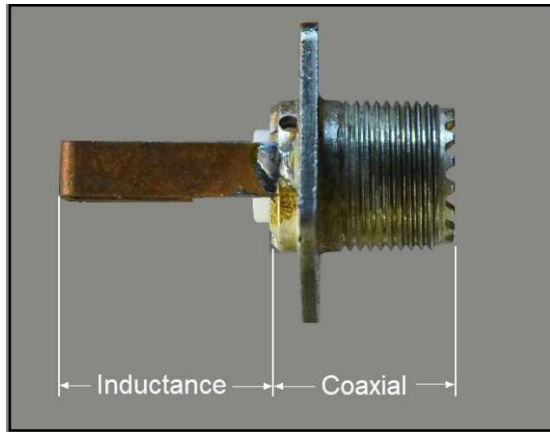
Within the SO-239 there will be some length of coaxial construction that will become part of the electrical length. This is the length from the coupling loop inductance to the tip of the mated PL-259 connector. This definition is used because the cable measurements stated by Motorola are conducted to the tip of the PL-259 connector.

The electrical length of the mated pair of the UG83-U and the SO-239 is the length from the reference plane of the female N side of the UG83-U adapter to the end of the coaxial portion of the SO-239.

Figure 24

The electrical length can be determined by measuring the physical length of a mated pair consisting of a UG83-U adapter and an SO-239 with Teflon dielectric, and then using the physical dimensions to compute the electrical length. (See Figure 25)
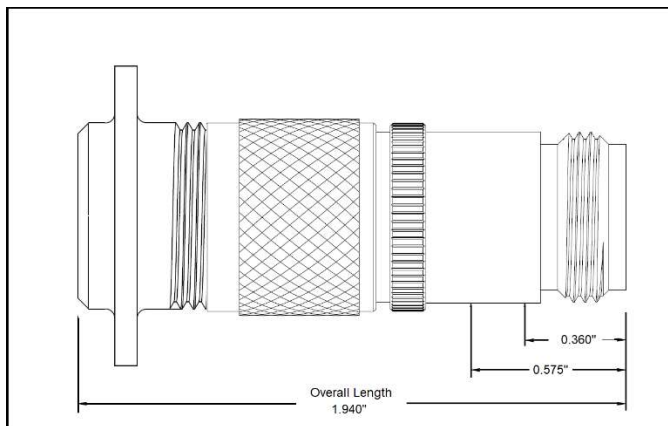
The electrical length of the mated pair consists of two portions:

First, there is a section in air from the N connector reference plane to the start of the Teflon dielectric which is: 0.575" – 0.360" = 0.215"

Second, there is a section with Teflon dielectric. Part of this dielectric exists in the adapter and part is within the portion of the SO-239 occupied by the PL-259 tip.

Figure 25

This is: 1.940" – 0.575" = 1.365" which is equivalent to an electrical length in air of 1.964". The combined electrical length in air is 1.964" + 0.215" = 2.179" or 55.3 mm electrical offset.

Note: One could measure the cavity's input impedance as it exists at the reference plane of the cavity's connector. If the cavity were equipped with type N connectors, the measurement reference plane could remain at the calibration reference plane. An impedance rotation from that point to a common junction would determine the correct length of the critical cable.

For cavities fitted with UHF connectors, an adapter is necessary between the test cable and the cavity. Since the electrical length of the adapter must be accounted for, it is logical to extend the measurement reference plane to the loop itself. This is the point used by Motorola in 1976.

By measuring at the actual loop, a network analyzer impedance curve (Figure 5 on page 6) exactly matches the Motorola Smith chart from 1976 (See Appendix 8). Also, the cavity's input impedance ($Z = Ø + j41$) and the critical cable electrical length ($Ø.141\lambda$) match exactly.

Appendix 6 – Complex Impedance – Rectangular and Polar Notation

COMPLEX IMPEDANCE
Rectangular and Polar Notation

$Z = R \pm jX$
$Z = |Z| \angle \Theta$

$Z = 2.14 + j27.5$
$Z = 27.58 \angle 78.2°$

X_L |Z|

Ø

X_c

R

Θ
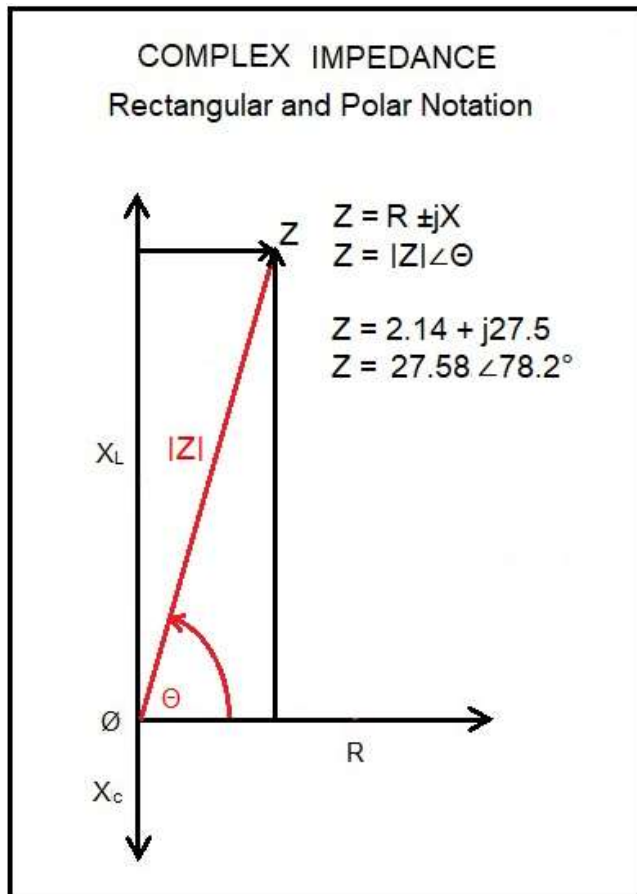
Impedance is expressed as a complex number because it consists of both a "real" component: resistance (R), and an "imaginary" component: reactance. (X)

When graphed using rectangular coordinates, the R values are placed on the horizontal axis and the reactance values are located on the vertical axis.

Any point in a system of rectangular coordinates can be located by an ordered pair which represent the horizontal and vertical distances from the origin.

Because all R values are assumed to be positive, only the two quadrants to the right of the origin are used when graphing impedance.

Inductive reactance is graphed using positive locations on the vertical axis and capacitive reactance is graphed using negative locations.

The rectangular notation form for expressing impedance is: $Z = R \pm jX$

It is also possible to draw a vector from the coordinate origin to any point. This vector is described by its length and the angle between the origin and the vector. This method of locating a point is called polar notation

The polar notation form for expressing impedance is $Z = |Z| \angle \Theta$ where $|Z|$ is the length of the impedance vector and Θ is the angle between the impedance vector and the horizontal axis.

Using basic right triangle math ($A^2 + B^2 = C^2$), it is easy to convert from one type of notation to the other. For example $|Z|$ = the square root of $R^2 + X^2$.

In Figure 14 on page 12, the cavity's input impedance at 458.5 MHz is shown as Z = 4.16 j56 . Converted to polar notation, $|Z| = 56.15\Omega \angle 86°$

In figure 15 on page 12, after rotation by 0.141λ, the input impedance at 458.5 MHz is shown as Z = 80.4 j289. Converted to polar notation, $|Z| = 299\Omega \angle -74.5°$

When considering bridging loss, using the magnitude of the impedance vector clearly shows the difference between a transformed impedance point and $Z_\emptyset$

Appendix 7 – Bandpass Cavity Considerations for Combiners

The N-way combining method relies on the selectivity of the resonant cavity filters to provide isolation between the branches at the common junction. As has been shown in the branch insertion loss discussion on page 11, an off-resonance branch is transformed to a higher impedance at the common junction by the critical length cable.

The farther around the impedance curve the off-resonance frequency is, the higher the transformed impedance will be and the lower the corresponding bridging loss will be. How far around the curve a frequency will be directly corresponds to the cavity isolation at that frequency.

Cavity isolation vs. frequency (or selectivity) is directly related to the cavity's 'Q'. Cavity Q depends on several factors. Two of the most important are the RF resistance of the materials used in the cavity's construction and the volume of the cavity. A larger cavity has a larger volume and correspondingly has a higher Q than a smaller cavity for the same coupling factor.

Thus, the size of the cavities used becomes an important consideration in the selection or construction of a combiner. A combiner using larger cavities will have less branch insertion loss for a given transmitter to transmitter frequency spacing than one using smaller cavities but the larger cavities increase the cost and space requirements of a combiner.

For example, Figure 27 shows the selectivity of a Motorola T1500 and Figure 28 shows the selectivity of a Telewave TWPC 10" cavity.

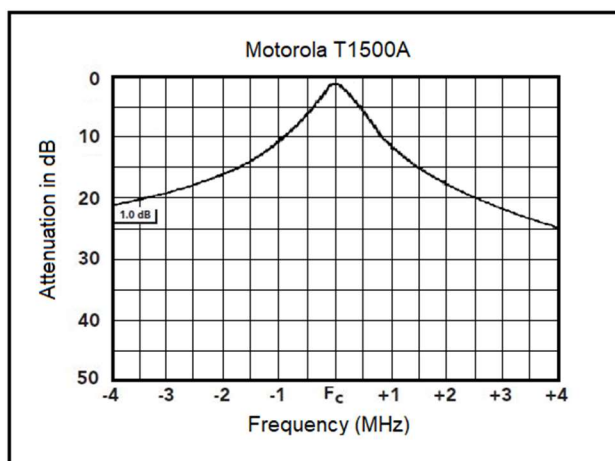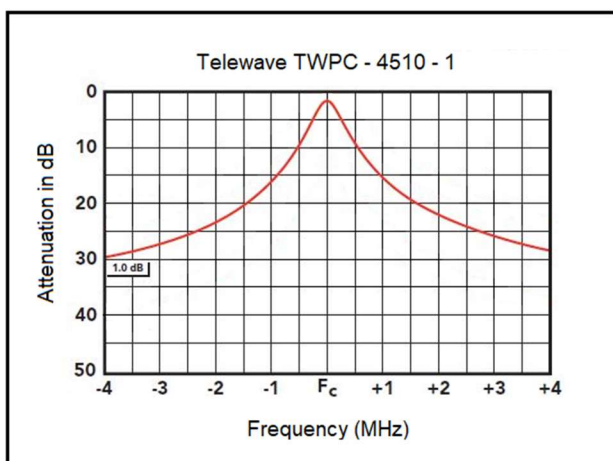Figure 27   Figure 28

At 500 kHz spacing the T1500 has about 6 dB of attenuation whereas the 10" cavity has 10 dB. Using the bridging loss curve (Figure 17 on page 12) and the branch insertion loss curves on page 13, the suitability of a combiner to a particular application can be evaluated.

Appendix 8 - Motorola T1500 Smith Chart from 1976

IMPEDANCE OR ADMITTANCE COORDINATES

$Z_o = 50\,\Omega$

RADIALLY SCALED PARAMETERS
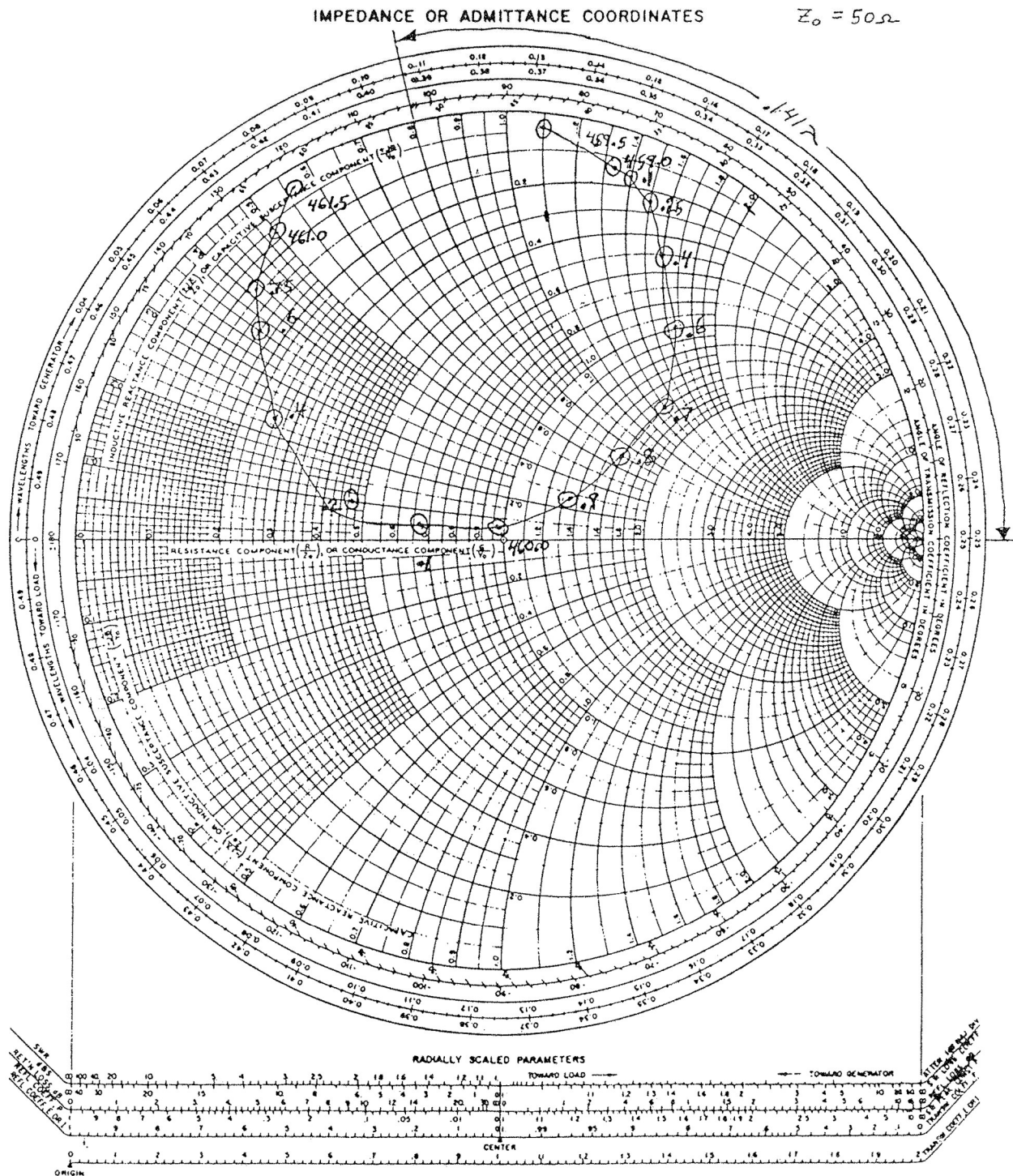
References

[1] … RF Engineering, Basic Concepts: The Smith Chart Fritz Caspers, CAS, Aarhus, June 2010 page 12

[2] … Handbook of Microwave Component Measurements with Advanced VNA Techniques, Joel P. Dunsmore © 2012 John Wiley & Sons, Ltd, 5.4.1 Resonator Responses on a Smith Chart

[3] … Motorola, National Engineering Services Manual R60, 1989, page 1-31

[4] … RF Engineering, Basic Concepts: The Smith Chart Fritz Caspers, CAS, Aarhus, June 2010, page 18 impedance transformation by transmission lines

[5] … Ibid. page 20 $\lambda/4$ line transformations

[6] … Transmission Lines and E.M. Waves, Prof R.K. Shevgaonkar, I.I.T., Lecture-8 page 3

[7] … Motorola, National Engineering Services Manual R60, 1989, page 1-31

[8] … Filters and Duplexers, T1500A Series, Motorola p.n. 68P81102E96-G, 1-25-1983

[9] … Ibid.

[10] ... Motorola, National Engineering Services Manual R60, 1989, page 1-31

[11] ... Ibid., page 1-29

[12] …Ibid., page 1-30

[13] …Ibid., page 1-30