

Managing Interstation Interference

Revised Second Edition



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Managing Interstation Interference
with
Coaxial Stubs and Filters

Second Edition
Revised and Updated

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1.0 DEFINING THE PROBLEM

1.1 Basics and Terminology

Whenever two or more transceivers are used in close proximity there is some level of interference involved. This level can vary from practically no problem to actually burning up components in the receiving radio. The purpose of this book is to identify and quantify the various parameters that create the interference and to show methods that will reduce or eliminate it.

To understand the problem it is necessary to learn about the imperfections in modern transmitters and receivers. Transmitters produce substantial output at multiples of the operating frequency. There are noise sources in them that will be radiated along with the desired signal. Receivers can only tolerate signals only below certain levels. High signal levels cause nonlinearities that make reception difficult or impossible, or that physically damage the receiver.

It's also necessary to understand something about the use of dB, dBm, dBm/Hz, etc., in order to discuss the problem intelligently. This is where we begin.

1.1.1 Understanding the various forms of dB that the RF engineer uses

Using dB allows us to talk about very large differences in power or Voltage levels with numbers that are easy to comprehend. For example: the maximum power output of a legal transmitter in the USA is 1500 watts and the noise floor of a modern receiver is 0.04 microvolts. The received power at this Voltage level into 50 ohms is 0.00000000000000003 watts or 3×10^{-17} watts. These numbers are not easy to deal with. RF engineers refer to power levels as dB above or below 1 milliwatt in a 50 ohm system and call the result dBm. Thus, 1 milliwatt is 0 dBm. If we convert the above power levels to dBm, we obtain +62 dBm for transmitter output and -135 dBm for receiver noise floor. The equations:

$$1500 \text{ watts in dBm} = 10 \log (1500 / .001) = +62$$

$$3 \times 10^{-17} \text{ watts in dBm} = 10 \log (3 \times 10^{-17} / .001) = -135$$

For narrow band signals such as CW, these numbers are easy to calculate from the amount of power read on a wattmeter. For example:

How many dBm is a 100 watt transmitter producing?

$$100 \text{ watts in dBm} = 10 \log (100/0.001) = 10 \log(100,000) = 10 \times 5 = +50\text{dBm}$$

A receiver noise floor measurement can be made with a true rms AC meter and a signal generator. The signal generator is connected to the receiver input and is set to the received frequency with its output set to zero. The rms meter reading at the receiver audio output is set to a convenient value on the scale by adjusting the radio volume control. The generator output is then adjusted for 3 dB increase

in audio output on the rms meter. The noise floor is then equal to the signal generator output power. Let us assume this level is -135 dBm and the receiver bandwidth is 500 Hz.

The level of noise introduced into a receiver output will depend upon the noise power and the bandwidth of the receiver. The noise may exist over a large bandwidth, but we are interested only in the noise power within our receiver bandwidth. By taking as a standard a very narrow slice of noise, we can obtain a number that is useful in predicting interference effects. The common bandwidth reference is 1 Hz. Noise power then is defined as noise density in terms of dBm/Hz. If we measure -135 dBm of noise out of a receiver that has 500 Hz bandwidth the power in dBm/Hz would be:

$$\text{Noise power} = -135 \text{ dBm} - 10 \log 500 = -135 \text{ dBm} - 27 \text{ dB} = -162 \text{ dBm/Hz}$$

Where the -27 dB is due to the ratio of bandwidths from 1 Hz to 500 Hz.

Another form commonly used is dBc, where c represents carrier. A number expressed in dBc is dB related to the carrier. Sideband power is generally expressed in dBc. The amplitude of noise sidebands of a transmitted carrier would also be expressed in dBc.

1.2 Power Levels in the Amateur Station Transmitter

Even the best transmitters produce undesired products along with a main signal. These products can produce excessive interference in other radios that are being used nearby unless steps are taken to reduce them. Harmonics occur at integer multiples of the transmitted frequency. The 2nd and 3rd harmonics are usually the worst. For example, a 7 MHz transmitter will produce harmonic signals at 14 and 21 MHz with sufficient energy to cause interference while trying to listen on those bands. Two forms of random noise are also generated in a transmitter. These are phase noise and wide-band transmitted noise, and are described below.

1.2.1 Harmonics

Transmitters produce harmonics, which can cause interference to receivers monitoring other amateur bands. The FCC regulations require these harmonics to be more than 40 dB down from the transmitted carrier or -40 dBc. When full legal power is being used, harmonics can be radiated at substantial power levels. Forty dB below 1500 watts is 150 milliWatts or +22 dBm. These numbers can be taken as the worst case. Amplifiers with pi-L output networks may have harmonic levels considerably below -40 dBc.

1.2.2 Phase Noise

The frequency generating portion of a transmitter contains one or more oscillators, usually part of a frequency synthesizer. The solid state devices used to produce the oscillators are not noise free, so some random noise is present in the device currents. This noise modulates the generated waveform and creates small amounts of phase shift. These phase perturbations can be seen on a spectrum analyzer as a broadening of the generated waveform, or noise sidebands. These noise sidebands are called phase noise. The noise sideband level is generally highest close to the carrier frequency. Further from the carrier frequency the noise is reduced until at some offset it reaches a plateau called the noise floor. Modern radios have a phase noise floor of about -140 dBm/Hz. (See note 1.) At an offset of 10 kHz it's about -138 and at 2 kHz offset it's about -120 dBm/Hz. The closer to the actual carrier that we tune, the stronger the phase noise becomes. This noise is present only when the transmitter is actually sending. The synthesizer is used in both transmitting and receiving, so phase noise also a receiving problem.

Note 1. To add realism to the calculations which follow, many of the radio characteristics listed have been taken from laboratory measurements of the Elecraft K3, performed by the ARRL. The author, on his K3, has measured some of the numbers that were not measured by the ARRL.

1.2.3 Wide Band Noise

Solid-state transmitters use wideband amplifiers from the last mixer through the power amplifier at the 100 watt level. The low level amplifiers in this chain have some internal noise, as does any real amplifier. This noise is amplified up to and through the final power amplifier. This amplified wide-band noise will be present when the transmitter is activated, even before the carrier is turned on. The power amplifier is usually followed by a low pass filter that helps remove harmonics and also reduces the noise output for bands above the band in use.

As an example, assume a clean, noise free signal is heterodyned to the operating frequency by a noise free injection source. The typical output level of the mixer might be -10 dBm. The gain required to produce 100 watts out of a transceiver would be 60 dB. Let's assume that the amplifier producing the gain has a noise figure of 4 dB. The input noise of such an amplifier is

$$-174 \text{ dBm/Hz} + 4 \text{ dB} = -170 \text{ dBm/Hz.}$$

Where -174 dBm/Hz is the noise in a 50 ohm resistor at 300 degrees Kelvin.

At the transceiver output this becomes -110 dBm/Hz. This level will increase by 27 dB for a receiver bandwidth of 500 Hz. Now the noise level becomes -83 dBm and related to the 100 watt carrier power it is

$$-83 - (+50) = -133 \text{ dBc.}$$

1.3 Maximum Receiver Levels

If a large enough signal is presented at a receiver input, damage to internal components will result. There are also limits to how large a signal that a receiver can tolerate before it refuses to allow simultaneous reception of normal signals. Transmitter harmonics and noise can interfere with reception if their levels are high enough. Let's examine these various problem areas and assign specific levels below which normal reception can take place.

1.3.1 Receiver Damage

Examining the components that make up receiver input circuits will show what power levels are safe. Input attenuators commonly use resistors rated at 0.1 watt. Switching diodes used to select input filters are usually rated at 0.2 watts. To apply a safety factor of 2 to the resistor ratings would result in 0.05 watts, or 50 milliWatts, or +17 dBm, being the maximum safe power. This is very safe, because it is highly unlikely that any one component will absorb most or all of the received power at the antenna input of the radio. The worst case would be in the attenuator resistors when the attenuation is set at maximum.

1.3.2 Receiver Blocking

When high level signals are applied to a receiver input there is a level, somewhat lower than the damage point, at which reception is blocked or disabled. This level is a function of receiver design and varies depending upon the manufacturer and model. A top of the line modern receiver has a blocking level around +5 dBm for a signal offset of 20 kHz and 0 dBm for 5 kHz spacing. The blocking level is usually defined as the 1 dB compression point. This is where the output of the receiver departs from linearity by 1 dB. Reception is impossible, but no damage results.

1.3.3 Intermodulation Distortion (IMD)

Intermodulation occurs when mixing action takes place between two or more signals to create new signals. Usually these are third-order products of the type $2F_1 - F_2$ and $2F_2 - F_1$. For example: F_1 at 7020 mixes with F_2 at 7025 to form intermod products at 7030 and 7015. The minimum level at which this happens in a modern radio is about -27 dBm. (Note that more and stronger IMD products are created above this level.) This is less than 10 millivolts and would read just under S9 +50 on an S meter. Receiving is possible with lots of interference from cross products. There are an infinite number of intermodulation products, but the odd order products fall closest to the receiving frequency and are usually the most bothersome. The "order" of the product is the sum of the harmonics of the two frequencies involved.

1.3.4 Noise Floor

All transistors and integrated circuits used in the manufacture of receiver amplifiers and mixers have internal noise sources. This internal noise creates a

limit, or floor, in signal level that can be received. The typical noise floor in today's radios is about -135 dBm as measured in a 500 Hz bandwidth.

1.4 The Power Level Chart

Figure 1 shows the various power levels of importance for analyzing interference problems in the amateur station. The dBm scale from +70 to -150 represents power levels from above maximum legal to below the best receiver sensitivity. Along the right side of the scale various important power levels are indicated. By taking the difference between two dBm levels of interest, the isolation in dB can be determined. For example: to reduce the +62 dBm transmitter output to that level which will not damage a receiver, 45 dB of isolation is required. To reduce the transmitted level to 10 dB below S9 that will minimize interference, 147 dB of isolation is needed. We will look into the various methods for obtaining isolation in later sections.

Power Levels in the Amateur Station

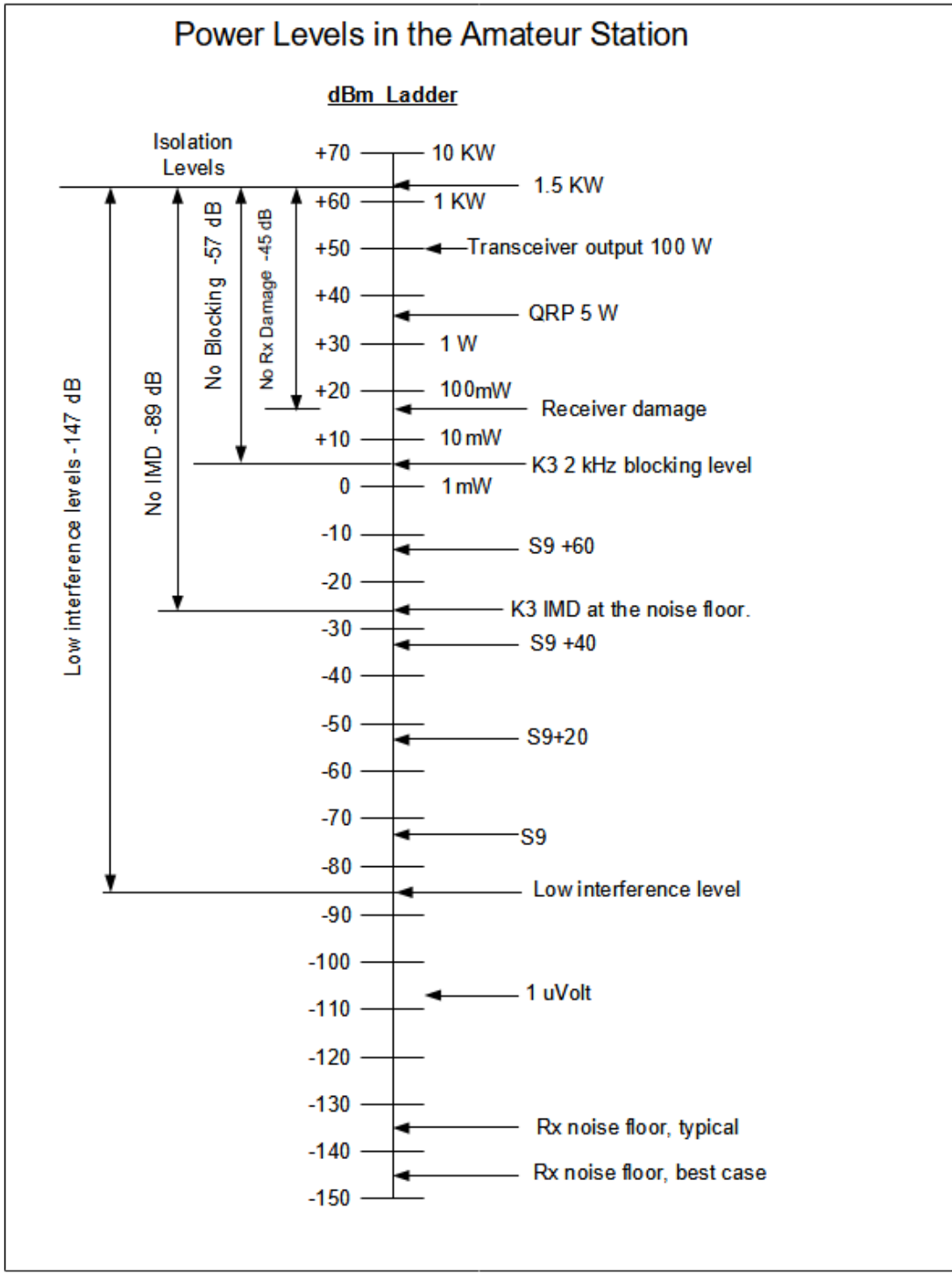


Figure 1 The dBm ladder.

2.0 FACTORS AFFECTING ISOLATION BETWEEN RADIOS

2.1 Basics

Several factors combine to determine how much power is presented to a receiver operated near a transmitter:

Antenna gain and spacing
Transmitter power output
Antenna orientation and polarization
Receiver antenna efficiency at the transmitted frequency

Obviously, more spacing between antennas will reduce the power picked up by the receiving antenna. Also, reducing the transmit power will reduce the interfering power at the receiver. If antennas are rotary, they may be oriented in such a manner that coupling is minimum. This usually occurs when the antenna ends are pointed at one another. Using cross-polarized antennas can also reduce coupling. Some stations with multiple transmitters and Yagi antennas use vertical antennas for spotting purposes. Using monoband antennas will reduce coupling due to the lower efficiency of Yagis when receiving off frequency signals. Multiband antennas are not the best choice when trying to isolate a receiver from a nearby transmitter.

2.2 Typical Numbers.

Figure 2 shows a two radio configuration using multiband antennas. One radio is transmitting at 1500 watts on 40 meters and the second is receiving on 20 meters. The path loss between the antennas is 35 dB. The tables show the levels of the various transmitted components. The receiver is using a 500 Hz bandwidth filter and we are assuming the antenna efficiency is the same on both bands.

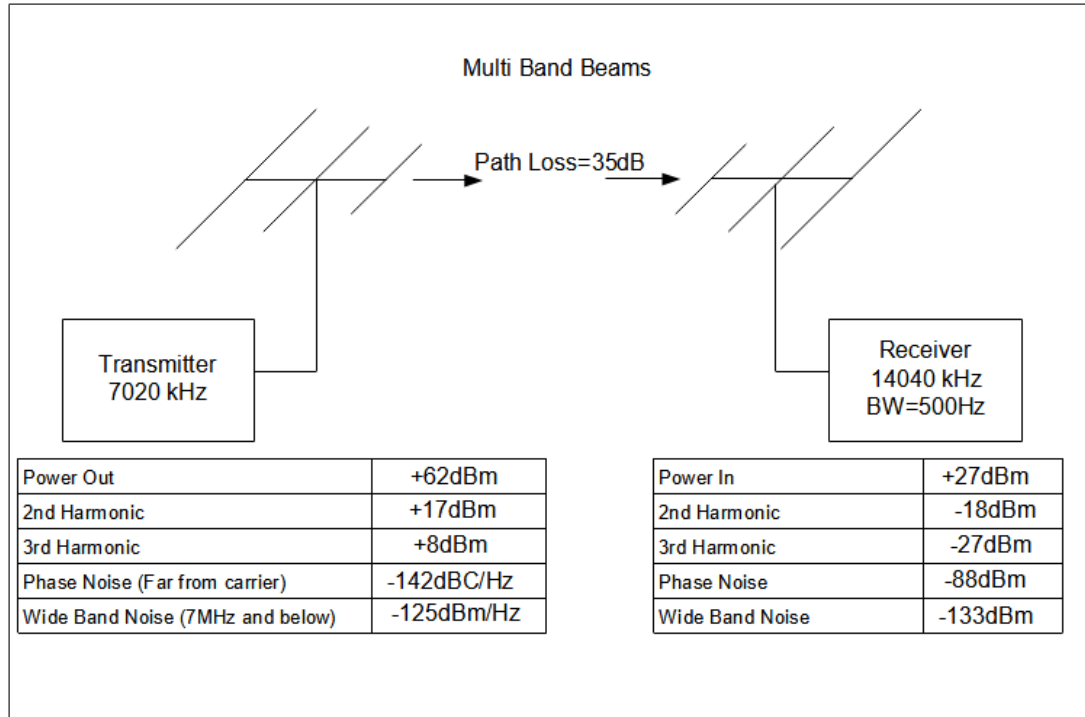


Figure 2 Transmit and receive levels in a typical 2 radio station.

The 7020 kHz carrier power radiated is +62 dBm and the receiver's antenna gets +27 dBm after the 35 dB path loss. From the dBm ladder in **Figure 1** we see that +27 dBm exceeds the damage level we have set for the receiver. To make it safe, we need an additional 10 dB of isolation. We are ignoring any gain that the antennas may add here.

The second harmonic comes in at -18 dBm which exceeds the level above which IMD products are created by about 14 dB. The third harmonic comes in at -27 dBm, which is just at the IMD creating level.

The phase noise (-88 dBm) and wideband noise (-133 dBm) each exceed the receiver noise floor. Let's see how we got those numbers:

Transmitted phase noise = -142 dBc/Hz (K3 measured value)
 Transmitted power carrier level = +62 dBm
 Transmitted phase noise power density = +62 dBm -142dBC/Hz = -80 dBm/Hz
 Received phase noise power density = -80 dBm/Hz -35 dB = -115 dBm/Hz
 Received phase noise power in 500 Hz band = -115 dBm/Hz +27 dB = -88 dBm

Transmitted wideband noise density = -125 dBm/Hz (measured value)
 Received wideband noise density = -125 dBm/Hz -35 dB = -160 dBm/Hz
 Received wideband noise power in 500 Hz band = -160 dBm/Hz + 27 dB = -133 dBm

Each of these noise factors will cause considerable interference to a radio with a -135 dBm noise floor except for the -133 dBm wide-band transmitted noise.

2.3 Antenna Isolation Measurements

Now that we have established some transmitted parameters and some receiver requirements for various levels of operation, we can look at the isolation measurements taken under real station conditions. At W2VJN there are two towers with a variety of antennas on each. Isolation measurements were made with a lab quality RF Voltmeter. The station transmitter was used with a Bird Wattmeter as the signal source. First we will look at a simple installation.

2.3.1 Small Tower with a Tribander and Wires.

Let's look at a small tower setup with a tri-bander and some wires. **Figure 3** shows the antenna configuration.

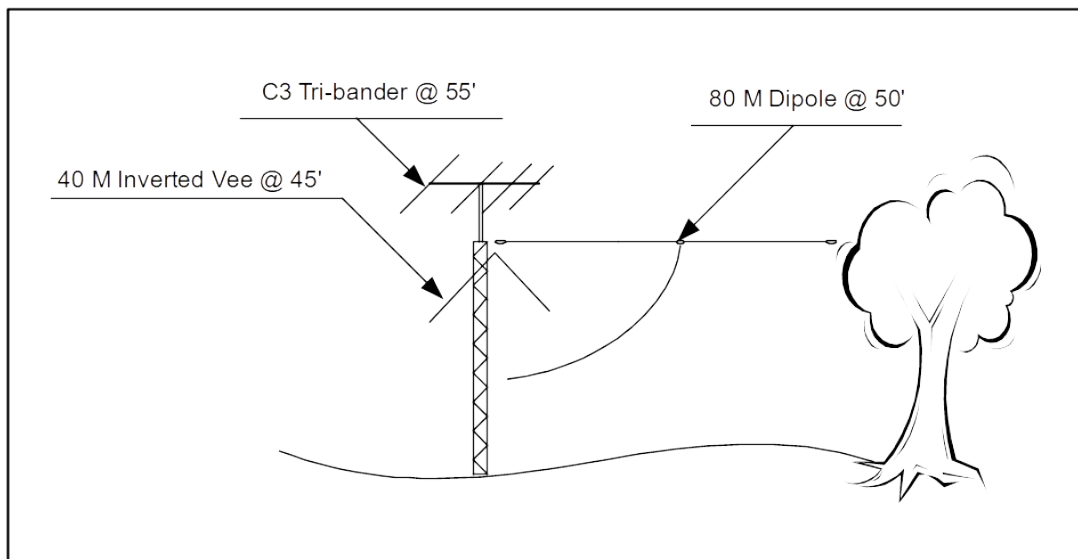


Figure 3 The small tower.

The tower is a 50 foot crank up. On top is a C3 tri-bander on a 5' mast. A 40 meter inverted vee is supported below the top of the tower at 45'. An 80 meter dipole runs from the tower top to a tree. The tri-bander is aligned with the 40 meter vee and is at right angles to the 80 meter dipole for the following measurements.

Table 1. Isolation in dB between the tri-bander and wires

TX Band	10	15	20	40	80
RX Band					
10	-	-	-	-35	-32
15	-	-	-	-35	-32
20	-	-	-	-35	-32
40	-36	-23	-30	-	-34
80	-36	-40	-38	-27	-

Table 1 indicates the possible combinations of one antenna transmitting and a second receiving. The top line is the transmitting frequency band and the left side column is the receiving frequency. The numbers refer to the isolation or attenuation in dB from the transmitter fundamental frequency to the receiver input. The tri-bander is used on 20, 15 and 10 meters while the 40 and 80 meter dipoles are used on their respective bands. Looking at the first data row, -35 and -32 under 40 and 80 indicates the dB isolation between the 40 and 80 meter dipoles and the tri-bander. That is, when transmitting on 40 meters with the inverted vee, the tri-bander will pick up 40 meter energy, which is 35 dB down from the transmitted power. Looking at the 40 meter RX row we see that transmitting into the tri-bander will inject 15 meter RF into the 40 meter antenna, which is 23 dB down from the radiated power, etc.

The value of 23 dB in the example is the worst case isolation between any combinations of antennas, so let's see what level of power will get to the receiver. If the transmitter is putting out 1500 watts, then 23 dB isolation will inject 7.5 watts into the second receiver if no precautions are taken. This is sure to cause extensive damage. Using dBm, 1500 watts is about +62 dBm, and subtracting 23 dB gives +39 dBm. From the dBm ladder in **Figure 1** we can see that we need about 22 dB additional isolation to prevent receiver damage. Even with 150 watt transmitters we will need 12 dB additional isolation to prevent receiver damage. So running low power does not guarantee safety. *Only QRP levels would be safe with no additional precautions.*

Since these numbers are measured under specific conditions at one location, an additional safety factor should be added for unknowns and variations in antennas. Ten dB would not be too extravagant.

2.3.2 A Larger Tower with Many Antennas

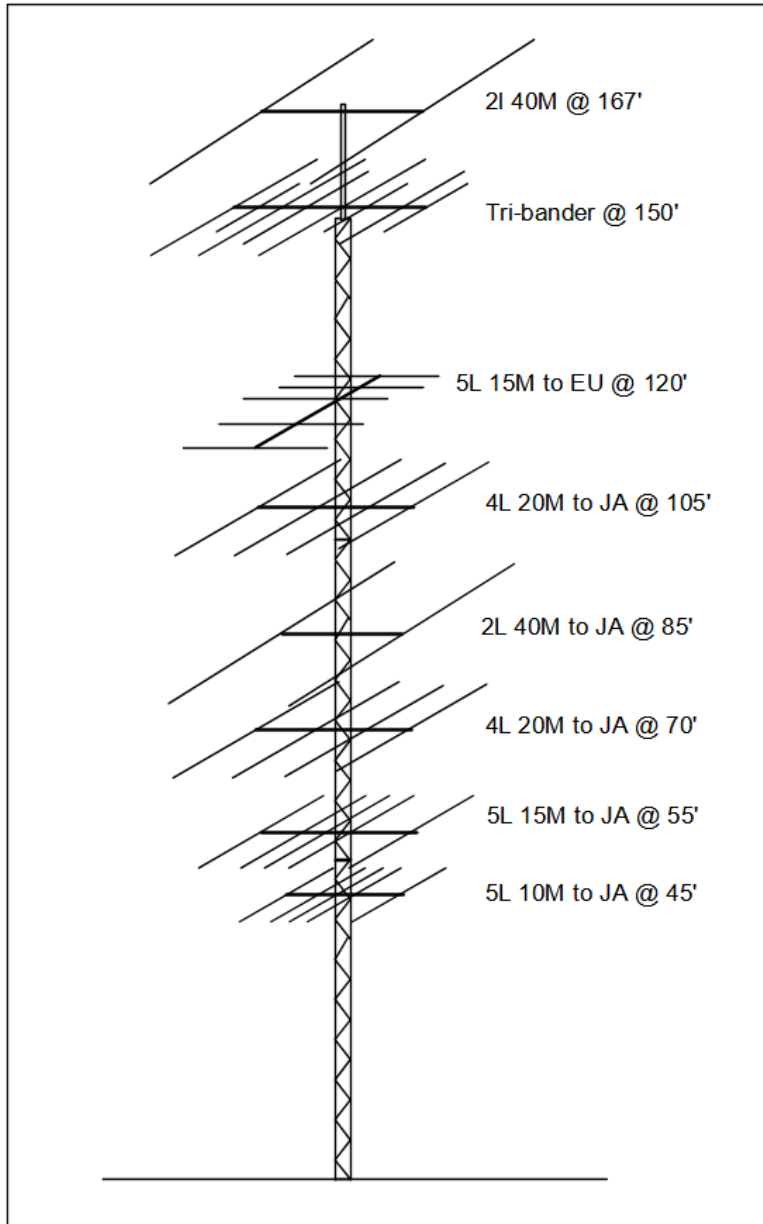


Figure 4 The large tower.

Figure 4 shows an example of Yagis on a 150 foot tower. Rotating on top are a 2 element 40 meter Yagi and a large tri-bander. They are spaced at 17 feet apart. There are 6 Yagis side mounted below the top of the tower. All except the upper 5 element 15 meter beam are fixed on JA. The upper 15 is pointed at Europe. Table 2 shows isolation measurements made between the tri-bander and the other antennas. The tri-bander was the transmitting antenna and the power coupled into the various antennas was measured. In each case the tri-bander was rotated until maximum coupled power was noted.

Table 2 Isolation in dB with tri-bander transmitting.

Tx	20	15	10
Rx			
40R	-49	-35	-43
40JA	-68	-71	-67
20Stk	-62	-56	-64
15EU	-55	-47	-64
15JA	-59	-67	-70
10JA	-75	-72	-53

The minimum isolation occurs between the tri-bander transmitting on 15 meters and the 40 meter beam directly above it. As the table shows, the isolation is just 35 dB. If 1500 watts is put into the tri-bander at 21 MHz, the 40 meter beam will send 0.43 watts, or +27 dBm to the second receiver. This is unacceptably high and additional isolation must be provided with filters and/or stubs.

There are some interesting numbers in the Table. For example, when the tri-bander is transmitting on 20, the 4 over 4 stack is picking up 0 dBm, or just 1 milliWatt. This means that two radios could be operating on 20 meters at the same time and no additional filters or stubs would be needed for protection even with full legal power being used. Of course, we would want some additional isolation to help prevent overload of the second receiver. Other combinations that will safely work are:

1. Tri-bander transmitting on 20 meters and second radio using the 10 meter JA antenna
2. Tri-bander transmitting on 15 meters and second radio using the 15 meter JA antenna.
3. Tri-bander transmitting on 10 meters and second radio using the 15 meter JA antenna.

There are also 6 wire antennas on the tower, 2 inverted-vees for 80 meters at right angles and four $\frac{1}{2}$ wave slopers for 160 meters. Isolation when the tri-bander is transmitting is shown in the Table 3. In each case the tri-bander has been rotated to the point of minimum isolation.

Table 3. Isolation between tri-bander and wires.

Tx	20	15	10
Rx			
160	>-80	-73	>-80
80EU	-54	-45	-60
80JA	-50	-46	-57

Coupling to the 160 meter slopers is minimal. This is probably due to the antenna centers being so far away from the tri-bander. The 80 meter inverted-vee centers are quite close to the tri-bander. Coupling to the 80 meter inverted-vees results in received powers in the +16 or +17 dBm range.

2.3.3 A Two Tower Station

Figure 5 shows the two tower setup at W2VJN. The spacing is 350 feet and the small tower is northwest of the big tower. Thus all of the fixed JA antennas on the big tower point almost directly at the small tower. When any rotary antenna is pointed at Europe, it is ends-on to the other tower. Note that the wire antennas on the larger tower are not shown.

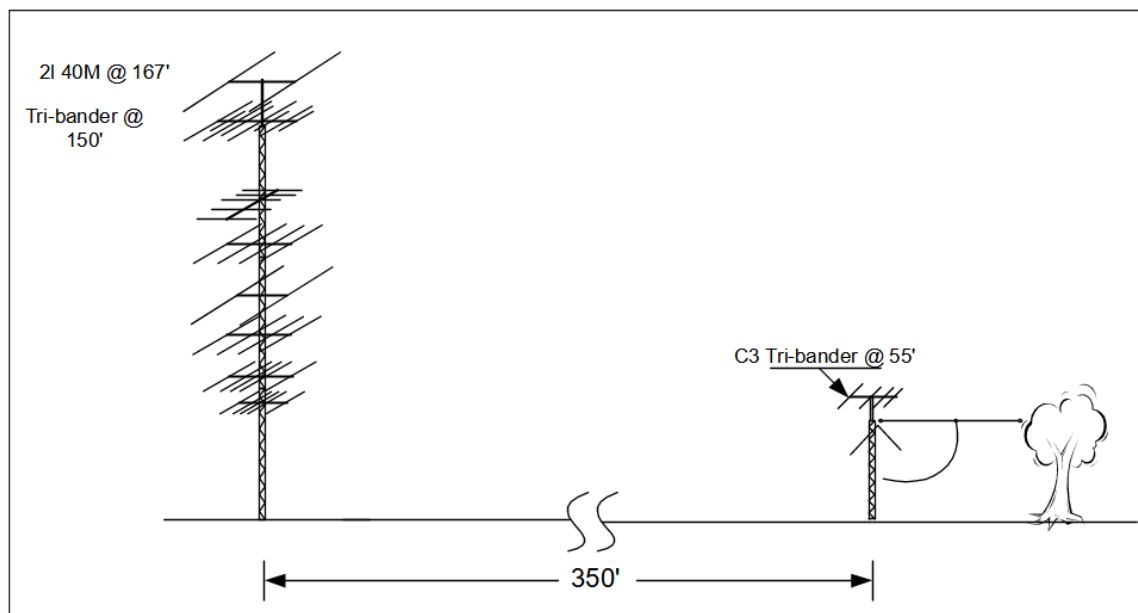


Figure 5 The two tower setup.

Table 4 lists the isolation between towers. The column on the left lists the big tower antennas and the top row lists the small tower antennas. The big tower is transmitting and the small is receiving. Rotary antennas are facing each other. Remembering that 45 dB is our minimum allowable isolation, we can see several combinations that are trouble. There are some marginal conditions which are so close to 45 dB that they can't be ignored. The worst case is the 4 over 4, 20 meter stack looking right at the C3. These towers are spaced 350 feet apart and yet there is only 25 dB isolation for this case. This will put 5 watts into the receiving radio and is clearly excessive. Even the X9 at a much higher point will put over 1.5 watts into the receiving radio. Caution is required when antennas face each other, even when fairly far apart.

Table 4. Isolation between towers.

		Rx Antenna	80M Dipole	40M Inv V	C3 Tri-bander
Tx Band	Tx Antenna				
160	NW Sloper	-52	-44	-48	
160	NE Sloper	-70	-63	-67	
160	SE Sloper	-68	-61	-65	
160	SW Sloper	-71	-51	-60	
80	EU Inv V	-43	-66	-53	
80	JA Inv V	-60	-58	-66	
40	JA 2L	-59	-36	-53	
40	Rotary 2L	-56	-36	-53	
20	JA 4/4	57	-45	-25*	
20	Rotary X9	-57	-50	-30	
15	JA 5L	-59	-44	-31	
15	EU 5L	-70	-69	-51	
15	Rotary X9	-59	-49	-35	
10	JA 5L	-70	-48	-35	
10	Rotary X9	-67	-50	-45	

*Worst case.

Figure 6 shows a sweep frequency response made with the tri-bander antennas pointing at each other. The generator was driving the X9 and the spectrum analyzer was monitoring the C3. (Note: There are some differences in the value between the sweep and Table 4 due to calibration methods.)

2.4 The Worst Case Conditions

Looking through the isolation tables for the examples given, we can see that the worst case minimum numbers are 23 to 27 dB. We have previously noted that 45 dB is the absolute minimum tolerable isolation which will prevent radio damage. We have noted also that 147 dB of isolation is needed to provide a two radio setup with minimal interference. Thus the additional isolation ranges from a minimum of 22 dB to a maximum of 124 dB. This should cover the range of requirements for most situations. Next we'll look at the options available for improvement.

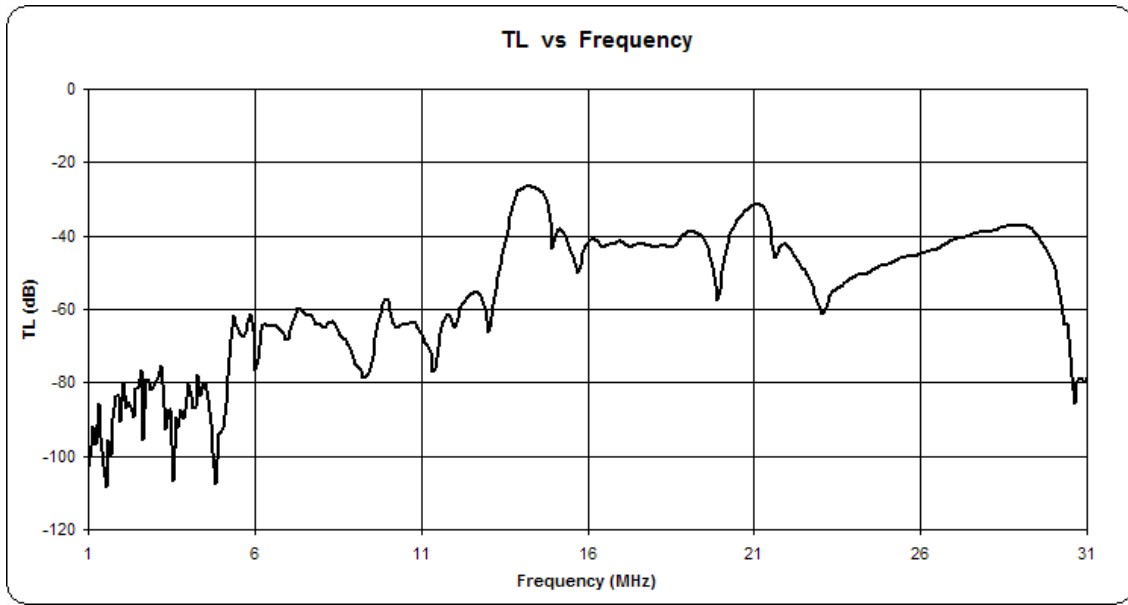


Figure 6. Tribanders pointing at each other. TL refers to transmission loss.

3.0 COAXIAL STUBS

3.1 Basics

A coaxial stub is a length of transmission line that is shorted or open at one end and connected to a circuit or another transmission line at the other end. Generally, in amateur applications the stub would be connected to a transmission line between a radio and an antenna. Stubs can act as inductors, capacitors or resonant circuits of series or parallel form, depending upon the wavelength and their length. Stubs may be used to reduce a transmitter harmonic output. Stubs also may be used to reduce receiver input signals at sub harmonics or harmonics of the desired signal. Stubs generally are cut to either $\frac{1}{4}$ or $\frac{1}{2}$ wavelength. Some special purpose stubs are cut to other fractions of a wavelength.

One wavelength may be calculated with the following formulas:

$$\text{One wavelength in feet} = V_p(983.6/\text{frequency in MHz})$$

$$\text{One wavelength in meters} = V_p(299.8/\text{frequency in MHz})$$

Where V_p = relative velocity of propagation for the cable in use

V_p is a function of the dielectric constant for the insulation between the inner and outer conductors of coaxial cable. The V_p for some common cables is listed in Table 5.

Table 5 Coax cable velocity factors

Cable	Vp
RG-58	0.665
RG-213	0.665
RG-8X	0.78
LMR-240	0.84
LMR-400	0.89
RG-142	0.695

Vp can be calculated for other dielectrics from the following formula:

$$V_p = 1/\sqrt{\epsilon}$$

Where ϵ = the cable insulation relative dielectric constant

Note that there can be some variation in Vp for the same cable but from a different lot or manufacturer. This is particularly true for foam types of insulation. Vp can also vary with frequency. If high precision is required, it is best to measure Vp for the cable in question.

Stubs are usually connected to a transmission line carrying power to and from an antenna. In some cases they are used to reduce the harmonic energy emanating from a transmitter. In other cases they are used to reduce the off frequency energy arriving at a receiver from another source. Operation relies on some fundamental properties of transmission lines. These principles are:

1. The input impedance of a $\frac{1}{2}$ wavelength line is the same at both ends. Thus if it is open-circuited at the far end, the input impedance is very high. If shorted, the input impedance is very low.
2. The input impedance of a $\frac{1}{4}$ wavelength line is opposite the impedance at the far end. If it is open at the far end, the input impedance is very low. If shorted, the input impedance is very high.

We can observe these characteristics in Table 6. The stubs are cut for the 40 meter band.

Table 6 40 Meter stub characteristics

BAND	1/4 SHORTED	WAVE 1/4 OPEN	WAVE 1/2 SHORTED	WAVE 1/2 OPEN	WAVE
80	NOT USED	NOT USED	OPEN CKT	SHORT CKT	
40	OPEN CKT *	SHORT CKT	SHORT CKT	OPEN CKT *	
20	SHORT CKT	OPEN CKT	SHORT CKT	OPEN CKT	
15	OPEN CKT	SHORT CKT	SHORT CKT	OPEN CKT	
10	SHORT CKT	OPEN CKT	SHORT CKT	OPEN CKT	

It can be seen that two types of stubs present an open circuit to the 40 meter transmission line. They are marked with an asterisk. These are the $\frac{1}{4}$ wave shorted and the $\frac{1}{2}$ wave open stubs. The other stubs would not be usable on 40 meters because they would short out the transmission line. The $\frac{1}{4}$ wave shorted stub presents a short to 20 meter and 10 meter harmonic energy and is useful for reducing transmitter harmonics from the 40 meter transmitter on those bands. The $\frac{1}{2}$ wave open stub also presents an open circuit to the 40 meter transmission line. As the table shows, it will reduce any 80 meter energy on the line. This typically would be used to protect a 40 meter receiver from an 80-meter transmitter.

Transmitter harmonics must be reduced at the transmitter. They cannot be filtered out at the receiver. Out of band signals can be filtered at the receiver, but if the transmitter harmonics fall in the listening band, nothing can be done at the receiver.

The $\frac{1}{4}$ wave shorted stub normally would be connected to the amplifier output. The $\frac{1}{2}$ wave open stub would not need to be connected to the amplifier output as it affects only received signals. It can be used at the transceiver output before the amplifier. This can sometimes be an advantage, as smaller stubs can be made with RG-8X or RG-58 which might heat up under some higher power conditions. The $\frac{1}{2}$ wave open stub also will reduce the transmitted wide-band noise on 80 meters when transmitting on 40.

Figure 8a shows the frequency response obtained with a 40 meter $\frac{1}{4}$ wave shorted stub. Note the null at 14 MHz. Additional nulls occur at all even harmonics of 7 MHz. Note the null at 28 MHz.

Figure 8b shows the response obtained with a 40 meter $\frac{1}{2}$ wave open stub. Note the first null at 3.5 MHz and additional nulls at multiples of 3x, 5x and 7x 3.5 MHz.

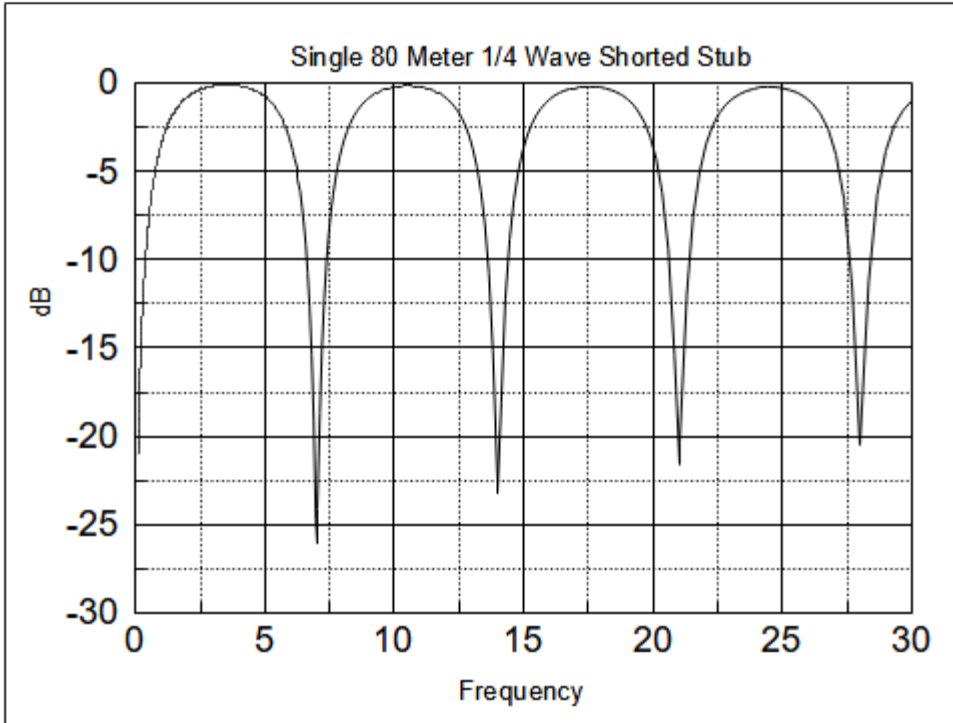


Figure 8a 7 MHz shorted $\frac{1}{4}$ wave stub.

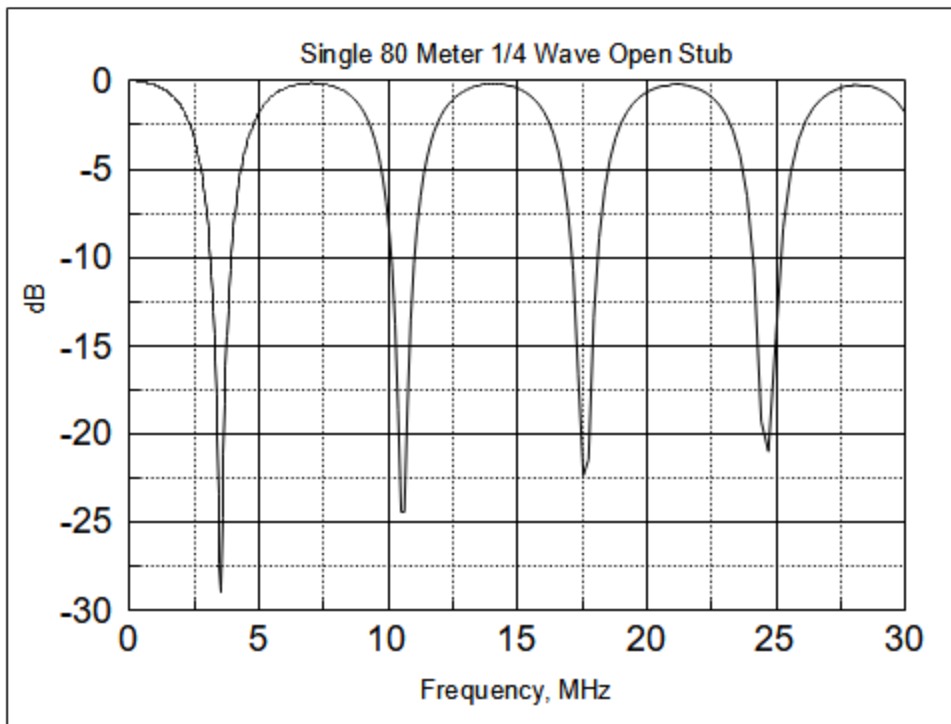


Figure 8b 7 MHz open $\frac{1}{2}$ wave stub.

The reactance seen at the input to a stub depends upon its length and whether it is shorted or open. A stub less than $\frac{1}{4}$ wave long presents a capacitive reactance at its input if it is open and inductive if shorted. As the stub is made

longer, the type of reactance reverses every $\frac{1}{4}$ wave. Table 7 shows the input reactance vs. length for stubs from 0 to 1 wavelength long.

Table 7 Stub reactance vs. length

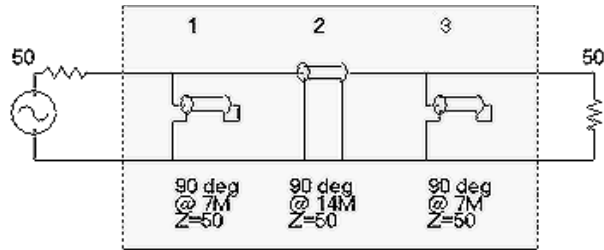
Length		Reactance type	
In Wavelengths	In Degrees	Open End	Shorted End
0 to 1/4	0 to 90	Capacitive	Inductive
1/4 to 1/2	90 to 180	Inductive	Capacitive
1/2 to 3/4	180 to 270	Capacitive	Inductive
3/4 to 1	270 to 360	Inductive	Capacitive

3.2 Using Elsie© to simulate stubs made with lossless coax.

Elsie is a filter program written by Jim Tonne, WB6BLD. (See the Sources section at the end of the book for references to this and other products.) It is very well conceived and makes filter design a trivial affair even for the inexperienced. While it will do complex inductor-capacitor filters, it also has lossless coax cable models. These can be strung together in various forms to simulate combinations of open and shorted stubs. So, the experimenting phase of a stub project can be mostly replaced by a few iterations done in Elsie on your computer. The resulting response curves will show no loss at the pass frequency and excess attenuation at the null frequency because the models are lossless. Its usefulness lies in the ability to show where those pass and null frequencies will occur for a given circuit design. Let's do a simple simulation of a pair of coupled shorted stubs.

Open the program by selecting **Manual parts entry**. Type 50 into the Input termination and Output termination windows.

Click on the shorted coax model as shown below as element number 1. Type in 90 (degrees) for **Line length**, 7 (MHz) for **Frequency** and 50 (ohms) for **Line impedance**, then click **Accept**. Next select the series coax cable model and type in 90 (degrees), 14 (MHz) and 50 (ohms) then click **Accept**. Last, select a second shorted coax model and repeat the 90 degrees, 7 MHz and 50 ohms as before. Click **Accept** and **Exit**. When you select **Schematic**, you should see what is shown in **Figure 9**. Select **Analysis** and set the various parameters for the desired graph type.



Family: Manual entry

Figure 9. Elsie schematic of example double stubs.

The plot obtained will look something like **Figure 10**.

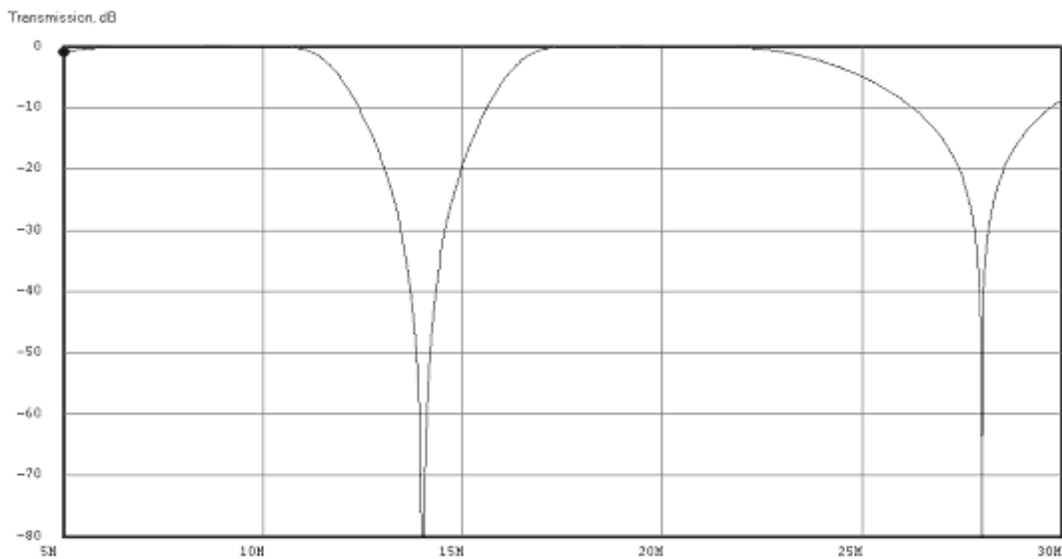


Figure 10 Elsie plot of double stubs.

Actual double 7 MHz stubs would have a few tenths of a dB loss at 7 MHz and the null would be about 70 dB at 14 MHz.

3.3 Making Stubs - Rough Cutting

Coax can be measured by stretching it out on a clean floor next to a measuring tape. Unless both are stretched very straight, the accuracy will be poor. It is easy to develop an error of a foot or more in a length of 46 feet when trying to cut an 80 meter stub. For this reason the rough cut should be made oversized by 5% or more.

Cable measuring meters that are available from Hykon Mfg. Co. allow accurate coaxial measurement to a fraction of an inch in any length. This company also markets various devices for coiling cable. **Figure 11** shows a homemade reel holder that the author uses. It holds a 500 foot spool of RG-213 or a 1000 foot spool of RG-8X.



Figure 11 The Hykon 1410 cable meter with a 500 foot roll of RG-213 mounted.

3.3.1 What Level of Accuracy is Required for the Final Trim?

The null frequency on the second harmonic is of primary importance. The cutting accuracy directly determines the null frequency. The accuracy required to cut the stub to within 3 dB of the best null is shown in Table 8. For example: if a ¼ wave, 14 MHz stub is being cut to null at 28 MHz, the perfect length would be 11' 8.16". This would give a null depth of about -32 dB. If we made an error of 0.66 inches, the null would be degraded at the design frequency by 3 dB or it would be -29 dB. The frequency of the null minimum would be off in frequency by about 130 kHz, or 0.4%.

Table 8 RG-213 cutting accuracy

Band	Length Deviation/3dB	Null Band	Null Frequency Error
1.8 MHz	1' 5"	3.6 MHz	+/-50 kHz
3.5 MHz	4.4"	7.0 MHz	+/-55 kHz
7.0 MHz	1.6"	14 MHz	+/-80 kHz
14 MHz	0.66"	28 MHz	+/-130 kHz

When multiple stubs are cascaded the maximum attenuation will be obtained if each stub is cut to the indicated accuracy or better. These numbers are typical for RG-213U and will vary somewhat for cable with a different velocity of propagation.

Table 9 shows the calculated lengths for $\frac{1}{4}$ wave length of RG-213. The actual length used for the rough cut by the author is also shown. Measurements are made with the Hykon cable meter and are quite accurate. The rough cut values represent the minimum length to guarantee a stub that is not too short for the bottom of the band.

Table 9 Rough cut lengths for RG-213

BAND	Calculated	Rough Cut
160	90' 10"	95'
80	46' 8"	47' 9"
40	23' 4"	24' 6"
20	11' 8"	12' 6"
15	7' 9"	8' 6"
10	5' 10"	6' 6"

3.3.2 *Tips on Installing PL-259 Connectors*

Installing PL-259 connectors can be done easily if the proper tools are used. A tool is available from various vendors that will cut the jacket, shield and insulation in one operation. The easiest connectors to install are the USA-made silver Teflon types with a gold plated center pin. They are available from many sources at a minimal cost. After the cable end is prepared, a small amount of grease applied to the end of the jacket will allow the connector to be threaded onto the cable easily. A 100 watt temperature controlled soldering iron with a $\frac{1}{4}$ inch tip will do a nice job on the soldering.

3.3.3 *Final Measurements*

There are two classes of measuring instruments generally used to measure the resonant frequency of a stub. These classes are the transmission type of instruments and the reflection types. They are also referred to as two port and single port instruments. In the transmission group, we have the spectrum analyzer with tracking generator, the network analyzer and the transmitter with receiver detector. In the reflection measuring class we have instruments such as the AIM 4170, the MFJ 259/269 series, the Autek RF-1 and others. Network analyzers usually can be set to do both types of measurements.

When a stub is used in any kind of system, it usually is installed as a parallel tap on a transmission line running from source to load. To reduce transmitter harmonics, for example, it is connected via a coaxial tee which is inserted in the antenna line. The tee then becomes part of the stub as it adds a small amount of length.

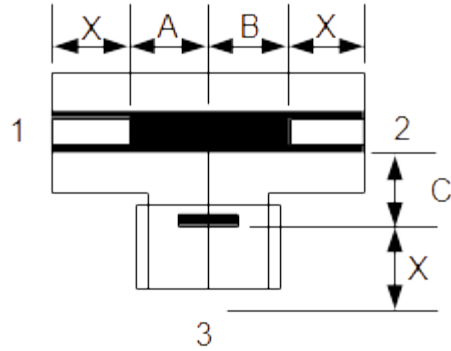


Figure 12. UHF tee dimensions

The figure above shows the cross section of an F-M-F UHF tee connector. Commonly, the stub is connected to side 1, the antenna to side 2 and the transmitter to side 3. The path from 2 to 3 is part of the antenna transmission line. The portions of the tee marked X are absorbed by the connectors fitted in operation and add nothing to the length of the coaxes in those connectors. The portions of the tee which add length to the connected cables are A, B and C. A is in series with the stub, while B and C are in series with the transmission line. **In order to trim the stub accurately, the tee must be in place on the measuring instrument.**

Let's examine the ways we measure stubs with the two types of instruments.

The transmission method is shown below.

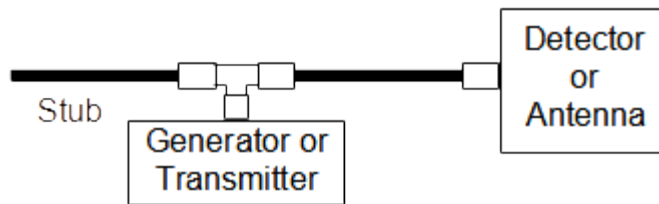


Figure 13. The transmission method

A coaxial tee is used to connect to the generator. One of the female sides of the tee connects to the stub being measured and the other female side of the tee goes to the detector. The male part of the tee connects to the generator. The connection through the tee adds length A to the stub length. The transmission line goes from point 3 to 2, and adds B+C. This setup is similar to the actual stub connections between a transmitter and antenna and gives accurate results. In practice, the generator is tuned until a null is shown on the detector.

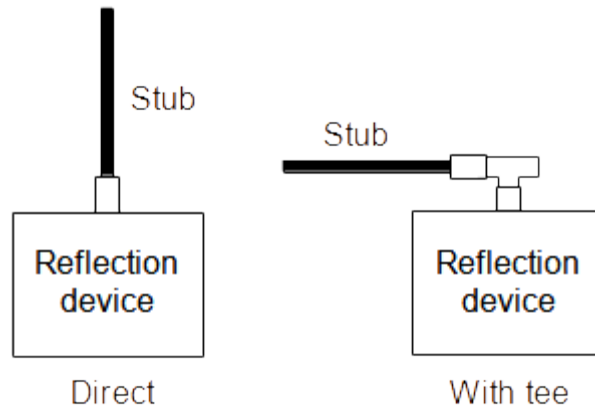


Figure 14. Reflection method

Referring to **Figure 12** again; the stub plus A will determine the resonant length while the transmission line goes through C + B. In **Figure 14**, if we choose the direct measurement method and cut the stub for resonance as determined when the phase goes to zero, it will be too long by B when connected through a tee. If we choose to measure with the tee in place, the stub will be too short by C when it is used on a transmission line. There is an error either way we try to measure it with a reflection tester.

This may seem like an insignificant dimension when cutting a stub for 80 meters, but it becomes very important at higher frequencies. Here are some actual measurements made on a $\frac{1}{4}$ wave open stub cut for 28.5 MHz.

Measured null frequency by transmission method = 28.725 MHz

Measured zero phase reading directly connected to a reflection instrument = 28.892 MHz.

Measured zero phase reading on a reflection instrument when measured with tee in place = 28.546 MHz.

We can see that the stub would be cut too long if placed directly on a reflection instrument or too short if the tee is inserted. **There is an error either with or without the tee.**

The error is about 0.6% at the target harmonic frequency. If the two readings are averaged, the net is 28.727 MHz or an error of about 0.02%. This is very close, but does not lend itself to a convenient method of cutting stubs.

How bad is this error? Look at **Figure 15**. It shows the example stub cut for 28.725 MHz. Line A shows where the null would have been if the stub was cut on a reflection instrument without the tee in place. Line B shows where the null would be if the stub was cut with the tee in place. The error in the null depth is about 4 dB. This is a lot of error to accept.

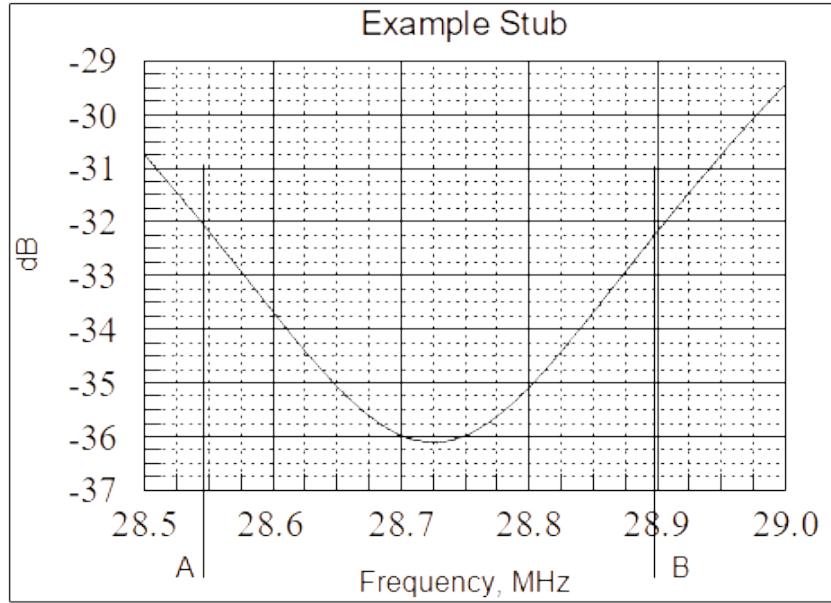


Figure 15. Cutting error example.

The measuring error vs. frequency is shown in **Figure 16**. It gets worse as we go higher in frequency and is not tolerable above 20 MHz , or so.

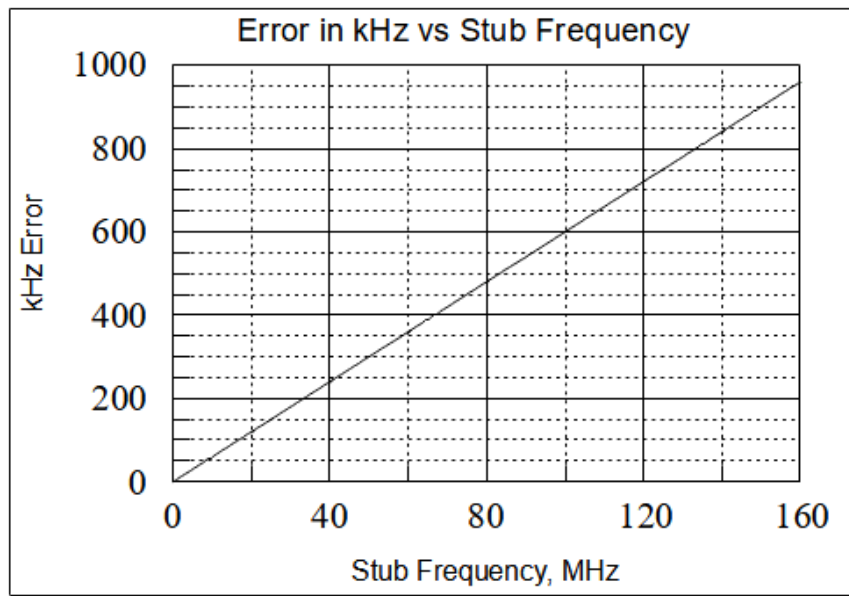


Figure 16. Cutting error vs. frequency.

Note that the reflection type instruments are fine for direct antenna measurements.

This kind of error made with reflection type instruments creates a lot more error in some other situations. Consider the stub switching arrangement made by NO9C shown schematically in **Figure 17**.

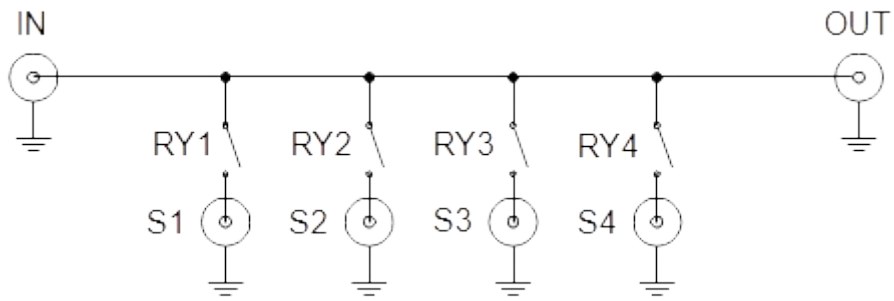


Figure 17. Stub switching box built by N9OC.

The IN and OUT connectors are bridged by a 50 ohm transmission line. Four relays, RY1-4, are connected to the transmission line with short wires. Four coax connectors are switched with the relays. Stubs are connected to the 4 coax connectors. This assembly is meant to connect between a transmitter and antenna in an SO2R station.

To cut the stubs, N9OC used an MFJ 259B (a reflection instrument) with each stub connected in sequence. To compensate for the relays, he measured the path length at 1.5 inches. This was subtracted from the measured stub length. This is an approximation, but is very reasonable. After he assembled the unit, he measured the stub resonances from the IN connector and they were all far lower than they were supposed to be.

In this instance, it is easy to see that the reflection instrument will give the wrong answer as it will add part of the common transmission line to the stub zero phase point. When this assembly is measured between IN and OUT with a transmission instrument, the nulls come out as planned. The stubs can be cut in place while connected to the switch box.

Clearly, stubs need to be measured in place when connected to a switching box.

Figure 18 is a photo of N9OC's switching box. Note how the input connects to the output across the box.

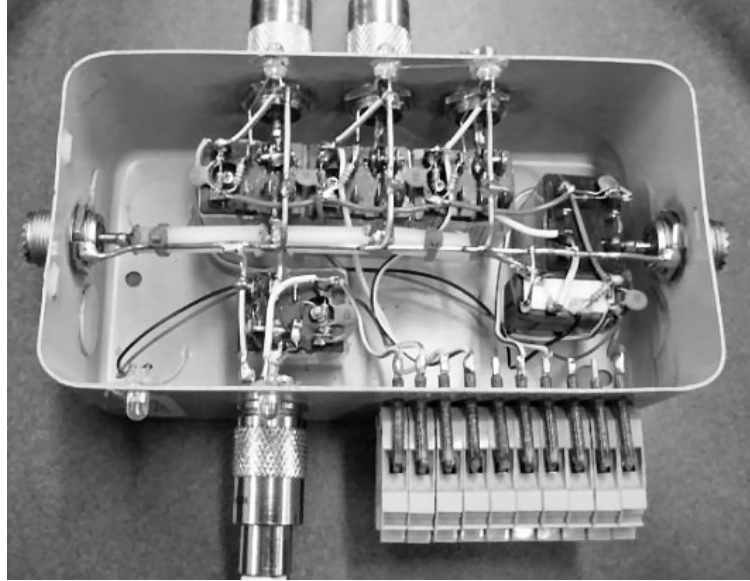


Figure 18. Photo of N90C's stub selector.

3.3.4 Terminating the coax.

The first step in applying a short circuit to the end of the coax after final cutting is to strip about $\frac{1}{2}$ inch of outer jacket. Take care not to cut through any shield wires.

Figure 19. Shorted and open terminations



Carefully spread out the shield wires to gain access to the inner insulation. Cut through the insulation all around the center conductor about $\frac{1}{4}$ inch in from the end. Strip this piece of insulation off. This leaves $\frac{1}{4}$ inch of center conductor exposed. Return the shield wires to their original place and squeeze the last $\frac{1}{4}$ inch down around the center conductor. Twist the end to tighten the shield around the center conductor. Apply a liberal amount of solder. **Figure 19** shows

a shorted end and an open end. They should be finished off by covering with shrink fit or, at a minimum, electrical tape.



Figure 20. After shrink wrapping.

Terminating an open stub is a bit easier. Cut around the outer jacket about $\frac{1}{4}$ inch in from the coax end. Cut deep enough to go through the

shield wires as well. Remove the jacket and shield wires and trim any stray wires.

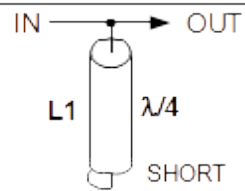
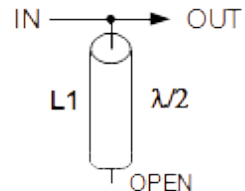
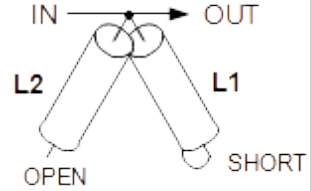
It's that simple to do each kind of termination. To complete the job, apply a short length of heat shrink tubing or a plastic cap.

These operations have shortened the stub by $\frac{1}{4}$ inch in each case. This can be compensated for in the cutting process by calculating how much lower in frequency to make the final cut before termination. For example: From table 8 we can calculate the frequency error this would cause on 14 MHz when cutting a 7 MHz stub. The error is proportional so $\frac{1}{4}$ inch would give $0.25/1.6 \times 80$ kHz or 12.5 kHz. So, cut the stub 12.5 kHz below the desired null frequency on 14 MHz.

An open ended termination can act as a very small antenna if the assembly is not mounted inside a shielded container. The radiation can be stopped with a small piece of shield braid over the open end. The end of the coax needs to be insulated with shrink wrap just over the inner insulation and extending a bit past the open center conductor. A piece of shield from a scrap of coax can cover up the unshielded part and it can be soldered to the stubs shield. The whole sandwich then can be double shrink wrapped. A piece of copper tubing can also be used with an end cap in place. If it is slipped over the coax for a few inches it will do a good shielding job.

3.4 Types of Stubs

There are three stub configurations in common use. Table 10 shows these types.

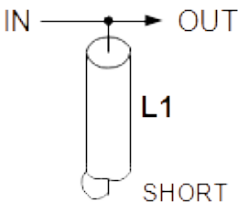
TABLE 10 TYPES OF STUBS, RG-213 CABLE				
TYPE	CONFIGURATION	NULL	DESIGN	F=Pass frequency, MHz Fn=Null frequency, MHz
1	 <p>IN → OUT L1 $\lambda/4$ SHORT</p>	2F, 4F, 6F, ETC	L1 in feet = $163.52/F$	
2	 <p>IN → OUT L1 $\lambda/2$ OPEN</p>	F/2, 3F/2, 5F/2, ETC	L1 in feet = $327.05/F$	
3	 <p>IN → OUT L2 L1 OPEN SHORT</p>	Fn	L1+L2 in feet = $163.52/F$ L2 in feet = $163.52/F_n$	

Arbitrary type numbers are assigned for the discussion. F is the pass frequency and Fn is the null frequency. The equations may be used for calculating the dimensions in feet.

3.4.1 Type 1 stubs.

Type 1 is the $\frac{1}{4}$ wavelength, shorted stub. The main use for Type 1 is transmitter harmonic reduction. It can provide 25 to 32 dB of harmonic null with very little loss at the fundamental frequency. RG-213 cable is safe to use for amateur power levels and higher as a Type 1 stub. They are useful for 160, 80, 40 and 20 meters in the contest station and also for reducing TVI from a 10 meter transmitter.

Table 11 shows characteristics and dimensions for some useful Type 1 stubs. 160 through 20 meter versions are listed. Note that the dimensions given are for the bottom of each band and must be modified to move up to the desired operating frequency. When rough cutting the cable, allow some extra to account for normal variations in the velocity of propagation and measurement error. Loss and null depth numbers are typical and depend somewhat on the particular make or run of cable.

TABLE 11					
APPLICATIONS OF THE TYPE 1 STUB					
					
REF #	PASS FREQUENCY	NULL FREQUENCY	LOSS	NULL	L1
1	1.800 MHz	3.600 MHz	0.3 dB	28 dB	90' 10"
2	3.500 MHz	7.000 MHz	0.2 dB	28 dB	46' 8.6"
3	7.000 MHz	14.000 MHz	0.2 dB	30 dB	23' 4.3"
4	14.000 MHz	28.000 MHz	0.2 dB	33 dB	11' 8"

Some plots of the Ref# 2 stub from Table 11 follow.

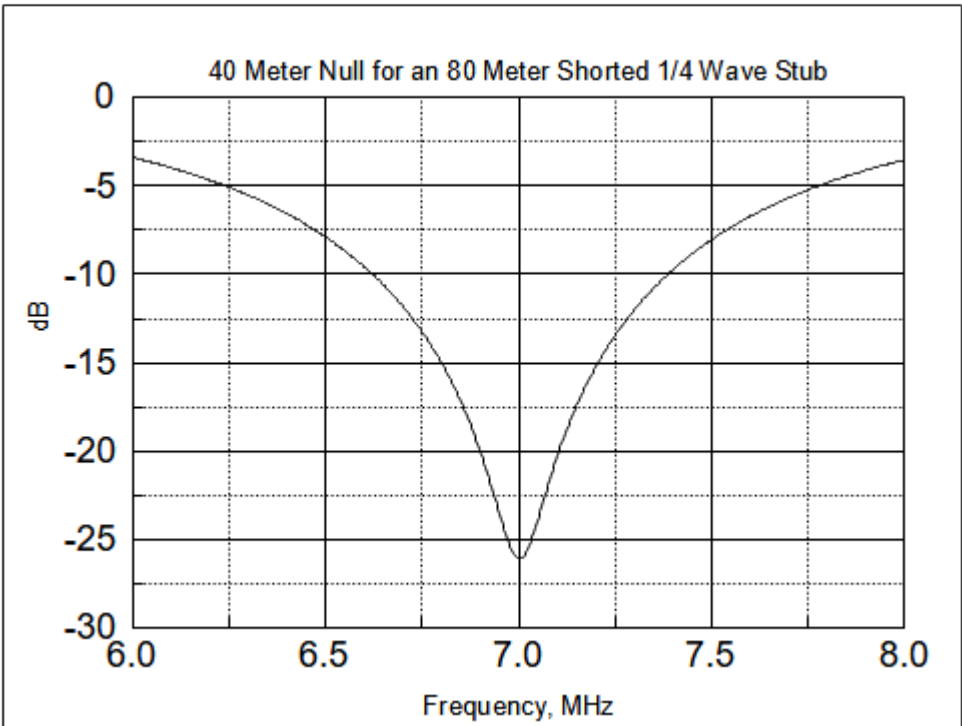


Figure 21. 40 meter null for an 80 meter shorted 1/4 wave stub.

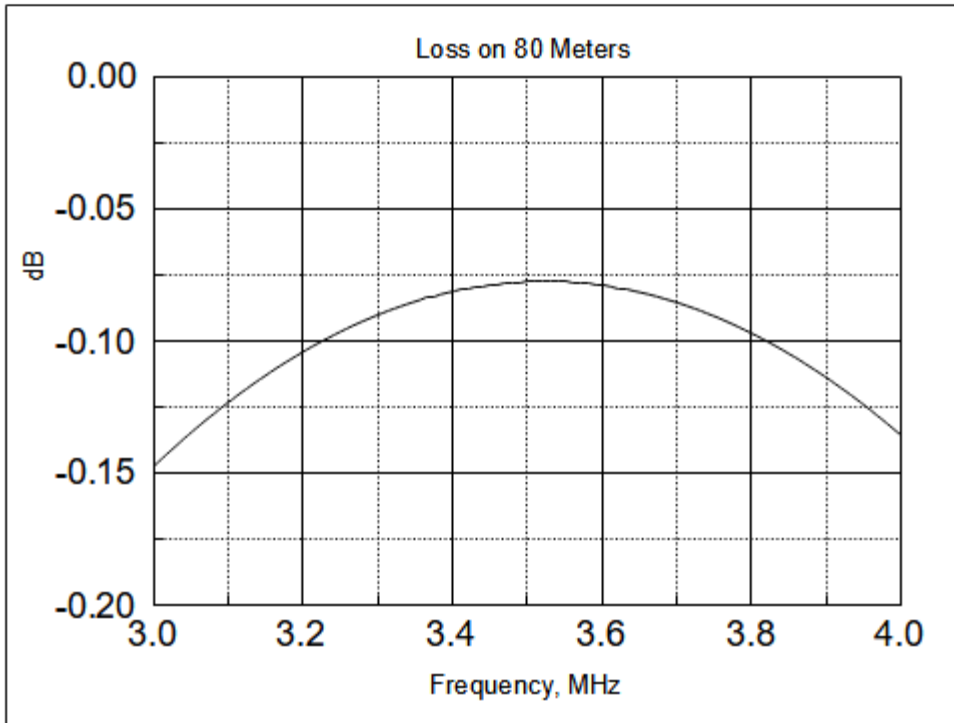
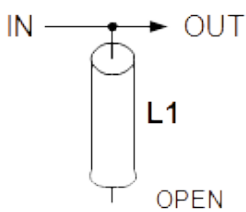


Figure 22. Loss vs. frequency for an 80 meter shorted $\frac{1}{4}$ wave stub.

Other stubs of similar design will exhibit similar characteristics. For the higher frequency bands, the nulls will be a bit deeper.

3.4.2 Type 2 stubs.

Type 2 is the $\frac{1}{2}$ wavelength, open ended stub. It can null the received energy from a transmitter operating at $\frac{1}{2}$ the listening frequency. It's useful for 80, 40, 20, and 10 meter sub-harmonics. It also can null 15 meters while listening on 20 or 10 meters. Note that it is used only to protect the receiver from excessive transmitter fundamental signals, so it may be used at the receiver input only. When there is enough transmitter signal to de-sense the receiver or cause inter-modulation effects, the Type 2 stub would be very useful. This brings up the question of whether smaller sized cable would be OK, and it will, except for the reduction in the null depth it will provides.

TABLE 12					
APPLICATIONS OF THE TYPE 2 STUB					
					
REF #	PASS FREQUENCY	NULL FREQUENCY	LOSS	NULL	L1
1	3.500 MHz	1.750 MHz	0.2 dB	29 dB	93' 5"
2	7.000 MHz	3.500 MHz	0.2 dB	30 dB	46' 8.6"
3	14.000 MHz	7.000 MHz	0.2 dB	32 dB	23' 4.3"
4	28.000 MHz	14.000 MHz	0.2 dB	33 dB	11' 8"

Some plots of the Ref# 3 stub from Table 12 follow.

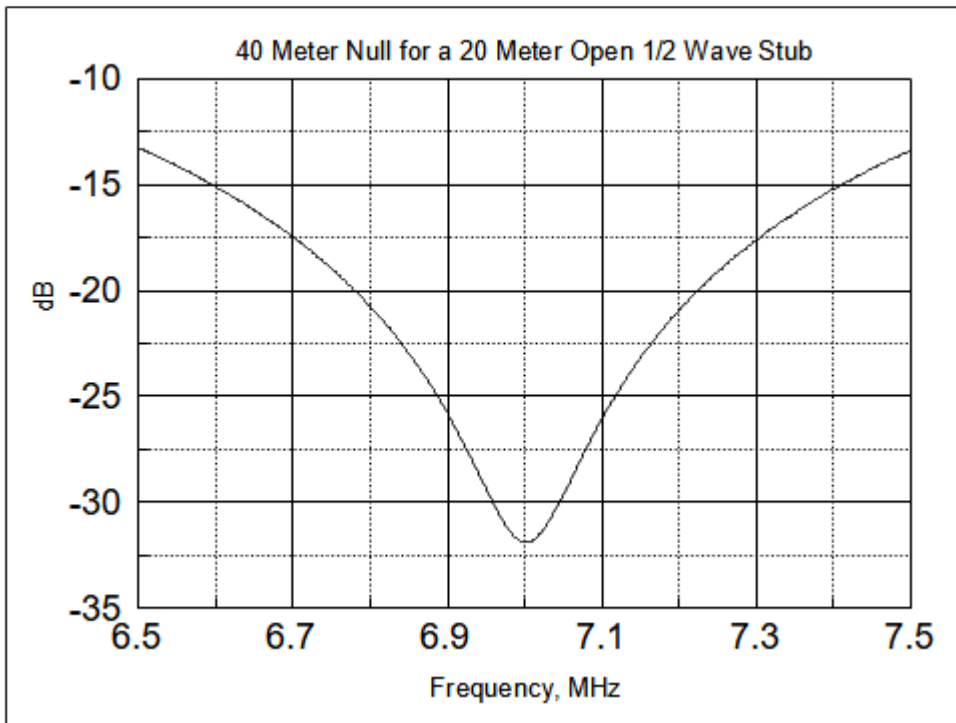


Figure 23. 40 meter null for a 20 meter open 1/2 wave stub.

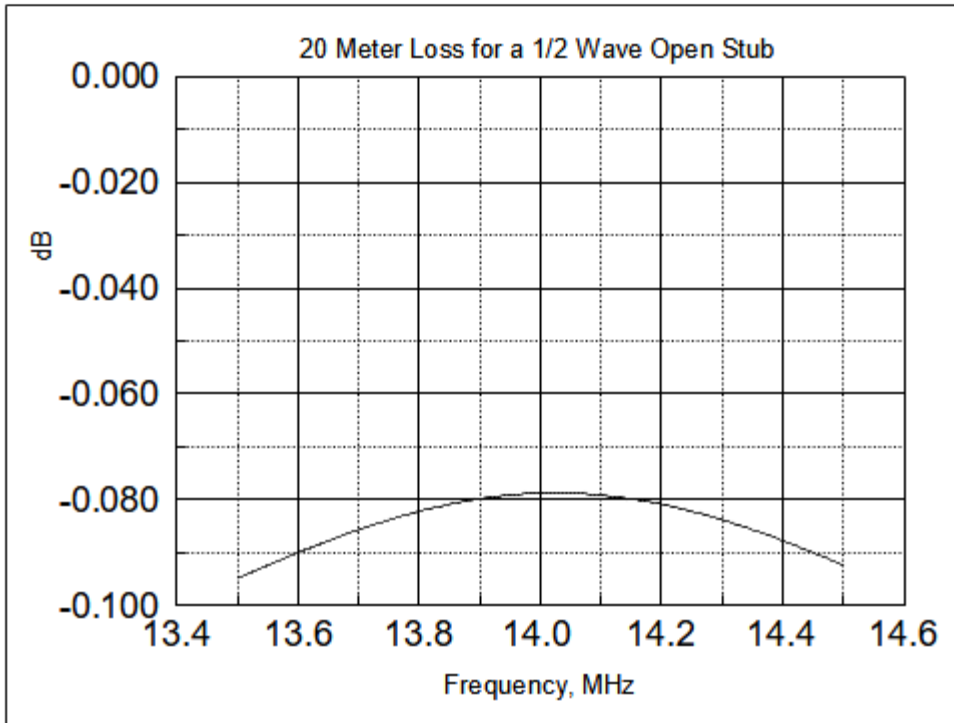


Figure 24. 20 meter loss for a 20 meter open $\frac{1}{2}$ wave stub.

3.4.3 Type 3 stubs

Type 3 stubs have been used for many years in amateur stations to null the 3rd harmonic of a 7 MHz transmitter. They provide over 35 dB attenuation in that capacity. Less known is the fact that they can null any frequency which is above the fundamental pass frequency. Another way to describe the type 3 is as a tapped coaxial resonator. The overall length of L1 + L2 is $\frac{1}{4}$ wave resonant at the pass frequency and the open part, L2, nulls at that frequency where it is $\frac{1}{4}$ wave long. There are many uses for this characteristic and we will examine some of them in detail.

The loss and null depth numbers are for RG 213. If a smaller cable or cable with more loss is used the loss at the operating frequency loss will be greater and the null will not be as deep. The reverse is true for larger cable or cable with less loss. The greater L1 is as a percentage of the overall length, the less will be the insertion loss. As L1 gets smaller and smaller, the loss increases until the design is useful for receiving only.

There is little to no danger of overheating a $\frac{1}{4}$ wavelength shorted stub from excessive loss or of flashover due to excessive voltage when using RG 213. This is true for power levels exceeding the legal level by several times. A bit more care is required when Type 3 tapped $\frac{1}{4}$ wave resonators are used as there is a voltage step up associated with them. The voltage pattern on a $\frac{1}{4}$ wave tapped resonator follows a $\frac{1}{4}$ sine wave. It will be zero at the shorted end and 100% at the open end. The shorted end is 0 degrees and the open end is 90 degrees. If we have a situation where the resonator is tapped in the middle, the

TABLE 13						
APPLICATIONS OF THE TYPE 3 STUB						
REF #	PASS FREQUENCY	NULL FREQUENCY	LOSS	NULL	L1	L2
1	1.830 MHz	3.525 MHz	0.30 dB	28 dB	42.967'	46.390'
2	3.525 MHz	3.800 MHz	5.0 dB	28 dB	3.357'	43.033'
3	7.025 MHz	7.200 MHz	14.5 dB	32 dB	0.565'	22.712'
4	7.0 MHz	14 MHz	0.13 dB	33 dB	11.680'	11.680'
4a	21 MHz	28 MHz	0.23 dB	30 dB	11.680'	11.680'
5	7 MHz	21 MHz	0.09 dB	37 dB	15.573'	7.787'
6	14 MHz	21 MHz	0.19 dB	35 dB	3.894'	7.787'
7	21 MHz	28 MHz	0.26 dB	35 dB	1.945'	5.840'
8	50 MHz	144 MHz	0.035 dB	42 dB	1.135'	2.135'
9	144 MHz	50 MHz	1.5 dB	38 dB	0.135'	3.270'

voltage at that point will be equal to the sin of 45 degrees times 100 or 70.7%. Suppose we are transmitting 1.5 kW and the voltage is 274 volts at the tap. Then the voltage at the open end will be $274 \times (100/70.7) = 387$ volts. This is true for a lossless cable. In practice, the voltage will be somewhat less and is a function of the cable loss. As we make L1 shorter and L2 longer, the voltage step up increases. Take the case of a tap at the 10% point or at 9 degrees. The open end would be $274 \times (100/15.6) = 1752$ for lossless cable. These voltage levels require that the cable assemblies be inside an enclosure for safety's sake. Additionally, if the end of the cable is not shielded it will radiate and possibly cause interference in the station.

3.4.3.1 Discussion of each Type 3 design from Table 13.

Ref# 1. There are two considerations when designing a stub. One is the effect on the transmitted signal and the other is the effect on the received signal. Looking at the numbers for this design, we see a small loss on the pass frequency (1.830 MHz), and a 24 dB null at 3525 kHz. A standard ¼ wave shorted Type 1 stub would present a small loss on frequency and a 24 dB null at 3660. **Figure 25** compares the two types of stubs. The tapped resonator from table 13 has a null depth of 18 dB at 3660 kHz.

If our main consideration is the effect on our transmitter second harmonic, we might automatically choose the $\frac{1}{4}$ wave shorted Type 1 stub. The improvement in harmonic reduction at 3660 kHz is about 6 dB. However, when we consider the reduction in received signal at 1830 kHz from a transmitter on 3525 kHz we see a reduction of 16 dB. So while we lose a small amount of transmitter harmonic reduction we gain a lot of receiver protection at 1830 kHz from a transmitter on 3525 kHz.

The tapped resonator requires a bit more work to assemble, but it gives a substantial improvement in performance. The amount of coax needed is the same in both cases.

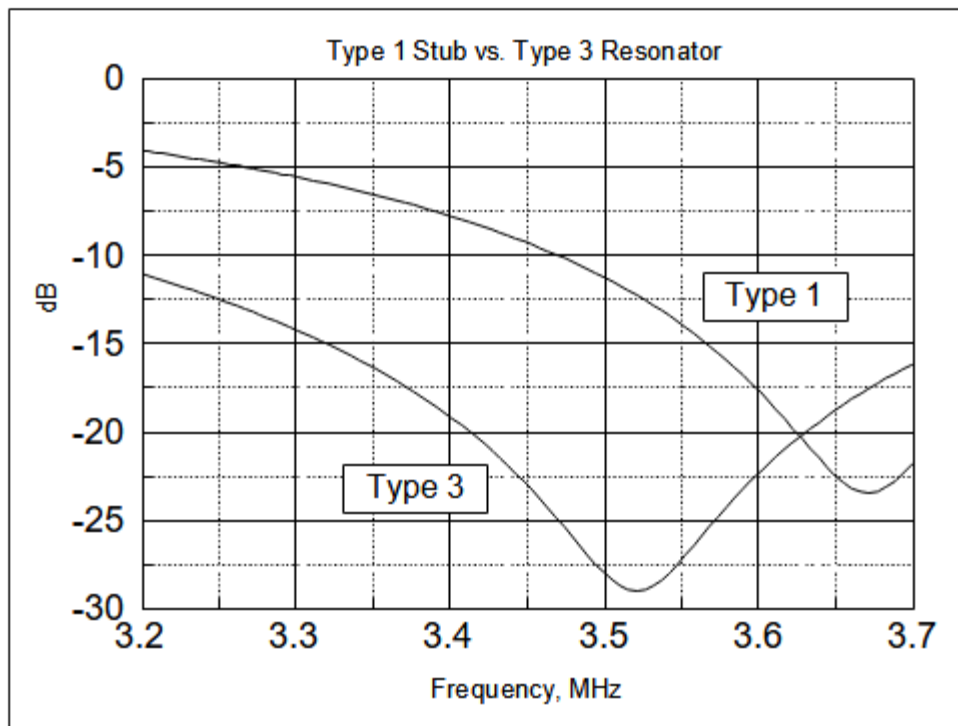


Figure 25. Comparing a Type 3 resonator with a Type 1, $\frac{1}{4}$ wave shorted stub.

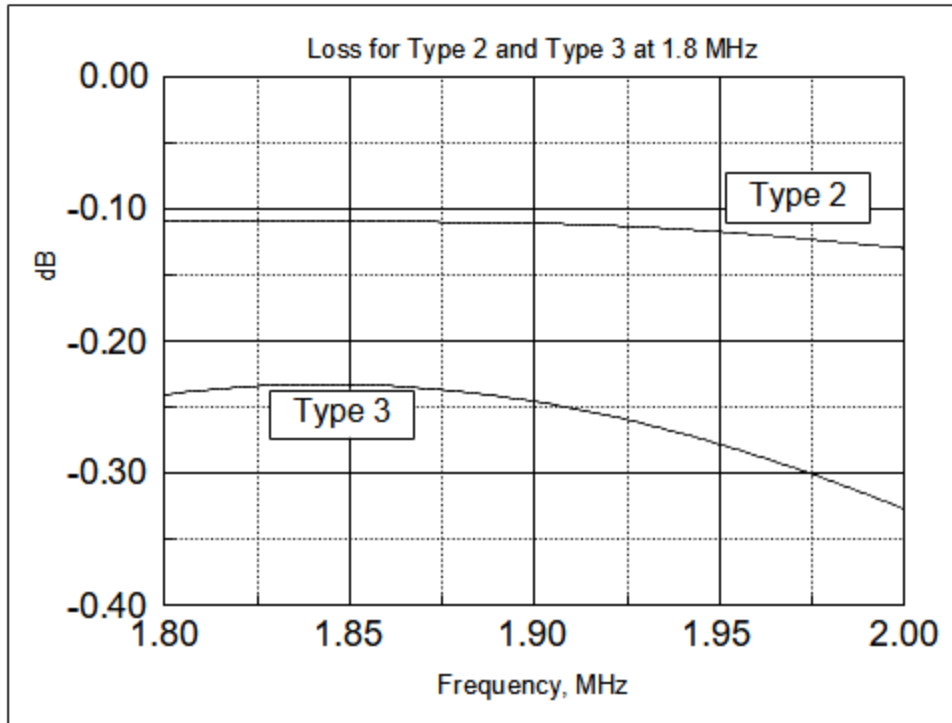


Figure 26. Losses for the Type 2 and 3 160 Meter stubs

Figure 26 compares the losses for the two types of stubs. While the Type 3 has over twice the loss of the Type 2, it is still very low. Of course, the loss in the type 3 depends on the position of the tap. As this tap is moved toward the low side, the loss increases.

Ref# 2. By selecting L2 to resonate at 3800 and then adding enough length to L1 to resonate at 3525, we can get an arrangement that may allow a CW radio to reject an SSB radio. Thus we can operate two radios in close proximity, one on CW and the other on SSB. With RG-213 coax cable, the loss at 3525 will be about 5 dB and the rejection at 3800 will be about 29 dB. These are measured numbers that can be obtained with real coax. **Figure 27** shows a measured plot for this design.

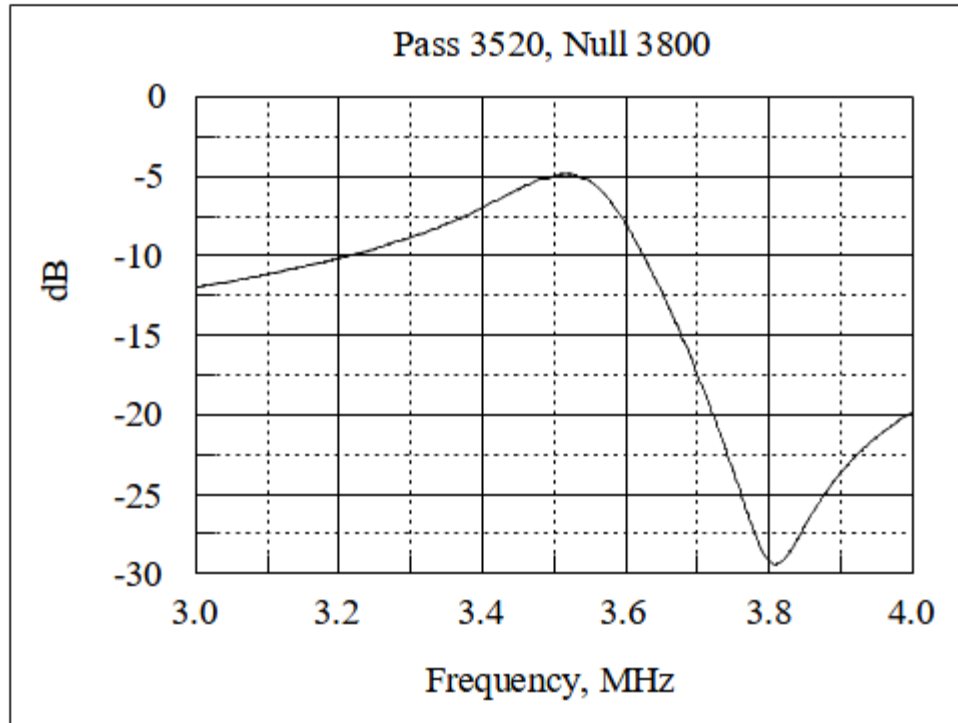


Figure 27. Type 3 stub that passes 80 CW and nulls 75 SSB.

Ref# 3. If the same method is used on 40 meters, the loss at the CW frequency will be excessive. There is much less percent difference in the two frequencies on 40 and other HF bands.

Ref# 4. If a $\frac{1}{4}$ wave resonator is tapped in the center, we find the second and fourth harmonics are nulled. If it is cut for 7 MHz, it will also pass 21 MHz. The null depths are as shown in Ref. #4 and 4a.

Ref# 5. If L1 is made $\frac{2}{3}$ of the length and L2 is the remaining $\frac{1}{3}$, the stub will null the third harmonic. The loss is very low and the null is quite deep. A 7 MHz design is shown.

Ref# 6. With L1 cut to $\frac{1}{3}$ and L2 to $\frac{2}{3}$, we have a design good for passing 14 while placing the null at 21MHz. This is an effective receiving filter for multi transmitter sites.

Ref# 7. With L1 cut to $\frac{1}{4}$ and L2 to $\frac{3}{4}$, this design passes 21 and nulls 28 MHz. This is also effective for multi transmitter sites.

Ref# 8 and 9. These are useful for the VHF bands. The loss for #9 is too high for transmitting as L1 is such a small part of the overall length.

3.4.3.2 Building and Testing a Ref# 7 Stub

Two pieces of coax need to be rough cut to start with. L1 will trim down to about 2 feet, so start with 2' 6". L2 will trim down to a little less than 6 feet, so start with 6' 6". Install a UHF plug (PL-259) on one end of each piece.

The transmission type of test set will give the best results. A UHF F-M-F tee will be used to make the measurements and to couple the two pieces together after they are cut. L2 should be connected to one of the female ends of the tee. The male end will be connected to one of the female ends on a second tee. The two free parts of the second tee will be put in line between the generator and detector of the transmission test set. Set the test set to the frequency of the desired null on 10 meters. In this case it is 28.1 MHz. Trim the end off of L2 until the null is exactly at 28.1 MHz. It will start off much lower. Use care to creep up on the best null in small steps so as to not overshoot.

Next we cut L1. There are two ways to do it.

The first uses a VSWR bridge. Attach L1 to the tee with L2. Connect the male of this tee to the second tee as before. Connect the male of the second tee to the VSWR bridge and put a good 50 ohm load on the other female port on the second tee. Set the frequency of the generator to the frequency on 15 meters which will be the desired operating frequency. Using a cable cutting tool, cut part way through the open end of L1 until the center conductor is shorted to the shield by the tools blade. Observe the VSWR. Move the cable cutter in steps to short L1 and watch the VSWR reading. It should be going down with each snip. As you get close to 1.0 VSWR, take smaller increments to creep up on the perfect match. Sweep the generator back and forth to determine the frequency of lowest VSWR. When the ideal length is reached, short the free end of L1.

If you have a good VHF transmission test set, the second method can be used. Connect L1 with the two tees as we had L2 connected earlier. Set the generator to the 2nd harmonic of the ¼ wave frequency of L1. In this case it would be:

$$F = 2 \times (163.524/1.945) = 168.1 \text{ MHz.}$$

Where: 163.524/L1 (ft) is the ¼ wave frequency for a shorted stub made with RG 213.

The minimum measured VSWR was at 21.05. Measured in the transmission test set, the null for L1 was at 164.1 MHz. In this case, that happens to be acceptable error. The ½ dB bandwidth as measured in the transmission test set was +/- 1 MHz, or 20 to 22 MHz. The cutting of L1 does not change the 28.1 MHz null, so a small error in L1 is not so important.

The resulting assembly nulls to -36 dB at 28.1 MHz and it has an insertion loss of 0.25 dB at 21 MHz.

The actual measured length of L1 was 1.979 feet or 1.7% longer than calculated.

The actual measured length of L2 was 5.792 feet or 0.8 shorter than calculated. The overall length was 7.771 feet or 0.2% shorter than calculated.

See **Figure 28** for a measured sweep.

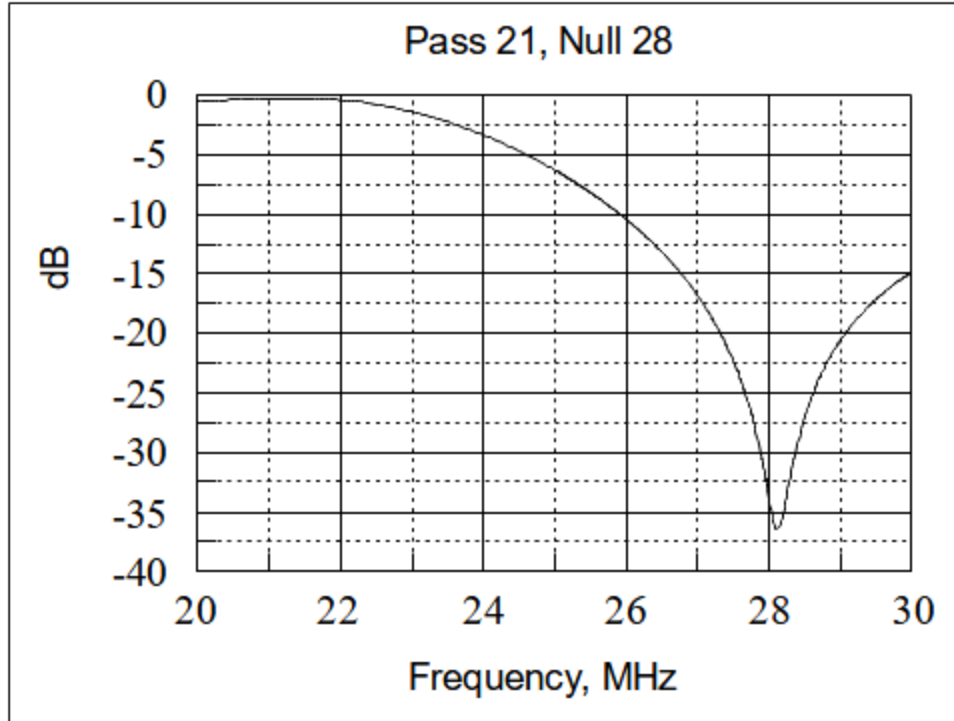


Figure 28. Measured response of a Ref. 7 resonator.

3.4.3.3 Placing the null below the pass frequency

Suppose we want to null 7 MHz and pass 18 MHz. First we need to divide the pass frequency by the desired null frequency. In this case the result is 2.57. For each integer above 1 obtained in the division we need to add a half wave at the higher frequency. In this case one additional half-wavelength is needed. The overall length of L1 + L2 must be an odd number of quarter waves. So we need to use $\frac{1}{4} + \frac{1}{2} = \frac{3}{4}$ waves.

$$L1 + L2 = 3(163.52/18) = 27.254'$$

Now, L2 needs to be $\frac{1}{4}$ wavelength at 7 MHz, so $L2 = 163.245/7 = 23.361'$ and therefore $L1 = 3.893$ resulting characteristic is shown in **Figure 29**.

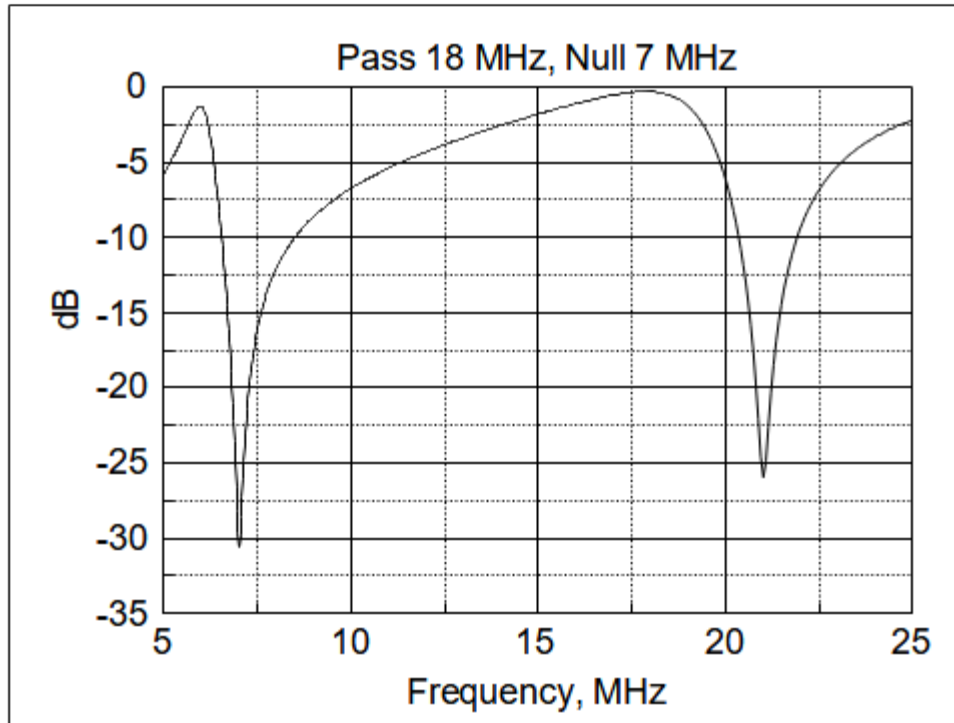


Figure 29. Passing 18 MHz and putting the null below at 7 MHz.

We can see that 6 MHz is also a pass frequency at $1/3$ of 18 and another null exists at 21 MHz as well as all odd harmonics of 7. The insertion loss at the desired pass frequency will be a bit lower than it is at 6 MHz, and is quite small.

3.4.3.4 Another characteristic of Type 3 stubs

Type 3 stubs can be described as coaxial resonators. Coaxial resonators are used at UHF and microwave frequencies to build filters and diplexers. Tuned circuits with capacitors and inductors are not practical at these frequencies because the values involved would be much too small to realize. Conversely, the techniques for building resonators at UHF are not practical at HF due to the physical size required. Resonators made with RG-213 have uses beyond what has already been described. The resonance effects can be used to serve as a type of band pass filter.

Consider a quarter wave transmission line, shorted on one end, open on the other with a tap that serves as both the input and output. (See **Figure 30.**) This is a filter with an input and a 50 ohm load at the same point and we are interested in its frequency response. **Figure 31** shows the swept response of such a resonator at its fundamental frequency with the in/out tap very close to the shorted end. Since the overall length is $1/4$ wave, or 90 degrees, we can say the tap is at 10 degrees. Note that the frequency response has a peak at the design frequency, 21 MHz, and a null slightly higher in frequency. Loss at resonance depends on the quality of the coax cable and is almost zero to a couple of dB for RG 213, depending upon the tap point position. Note that it rejects 14 MHz by about 14 dB and 7 MHz by about 17 dB. There is also a loss of some 15 dB at

28 MHz. Such tapped resonator filters were built in the 1980s by Ron Crouse at Dunestar for a few of the multi-multi contest stations. Low loss was achieved by making the resonator out of brass rod inside of a large sheet metal box. The rod was zig zagged around in the box between partitions to form a transmission line.

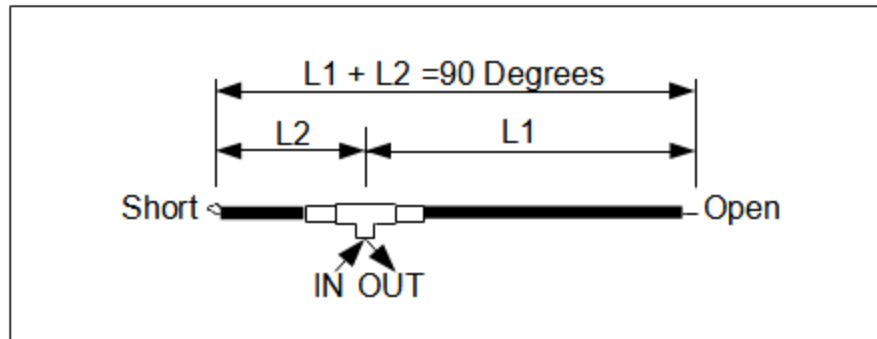


Figure 30. 90 Degree resonator

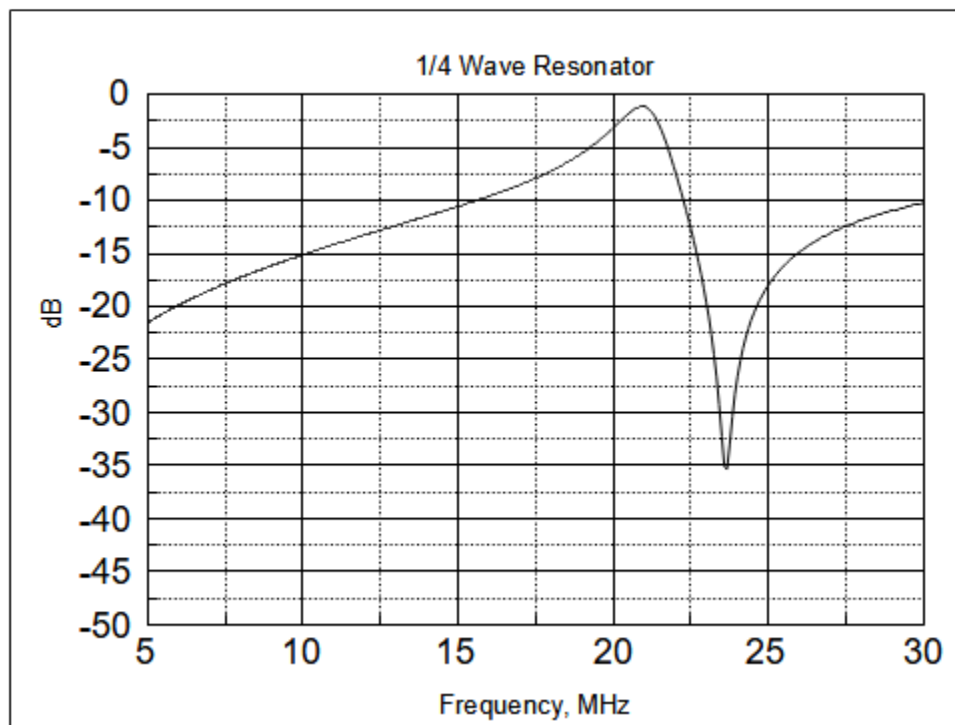


Figure 31. Quarter wave resonator frequency response with L2=10 degrees.

The notch just above the resonant frequency is due to the portion of the line above the feed point acting as an open circuit $\frac{1}{4}$ wave line. This provides a short at the feed point. In **Figure 31** this occurs at a frequency equal to $90/80 \times 21$ MHz, or about 23.6 MHz. i.e. The length of the top portion is $8/9$ ths of a quarter wave at 21 MHz. Moving the tap moves the null frequency and also changes the bandwidth of the filter. The lower we place the tap (i.e. closer to the short), the sharper the response, and the greater will be the loss at resonance.

Two of these 90 degree resonators can be coupled together to produce a sharper resonance with greater rejection below resonance. The optimum coupling cable length is 90 degrees, or $\frac{1}{4}$ wavelength. If the coupler is more or less than 90 degrees a double peaked response will result. **Figure 32** shows the response for a coupled pair. Note the attenuation at 14 MHz is now 27 dB and at 7 MHz it is 37 dB.

If a pair of these resonators is coupled with a $\frac{1}{4}$ wave line at the null frequency, the attenuation will be doubled plus an additional 6 dB. If other lengths are used for the coupling cable, the pass band will broaden somewhat.

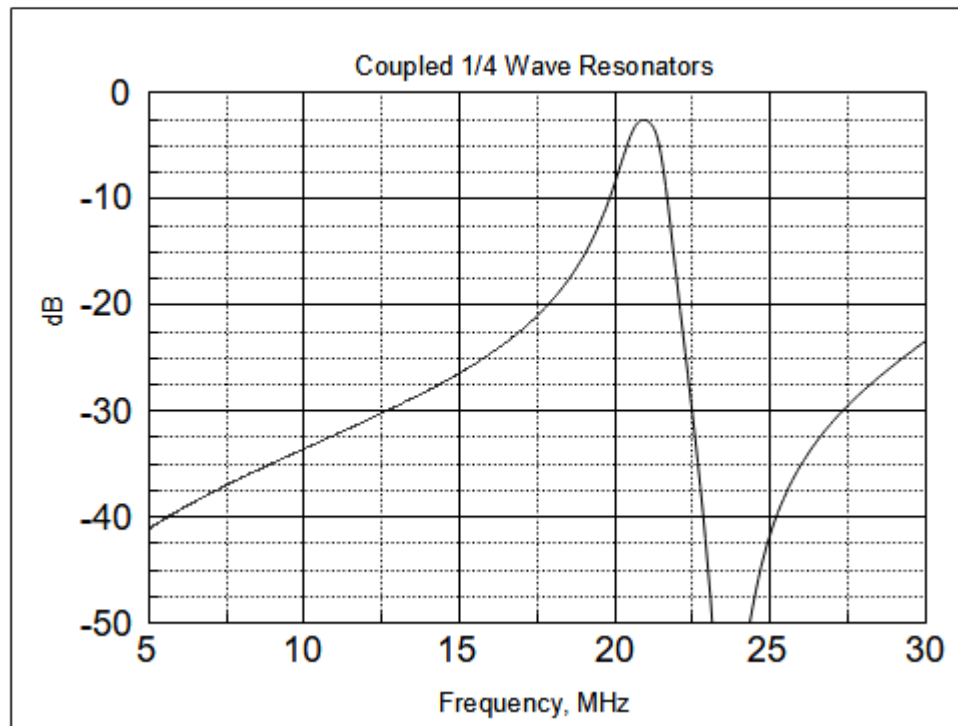


Figure 32. A coupled pair of $\frac{1}{4}$ wave resonators.

There is a simpler way to couple two resonators together which will increase the selectivity for a given tap setting. Two of the 90 degree resonators can be connected together at the top, or open end. This makes the total length of the line 180 degrees, or $\frac{1}{2}$ wavelength and we can then put the input on one side of center and the output on the other. See **Figure 33**.

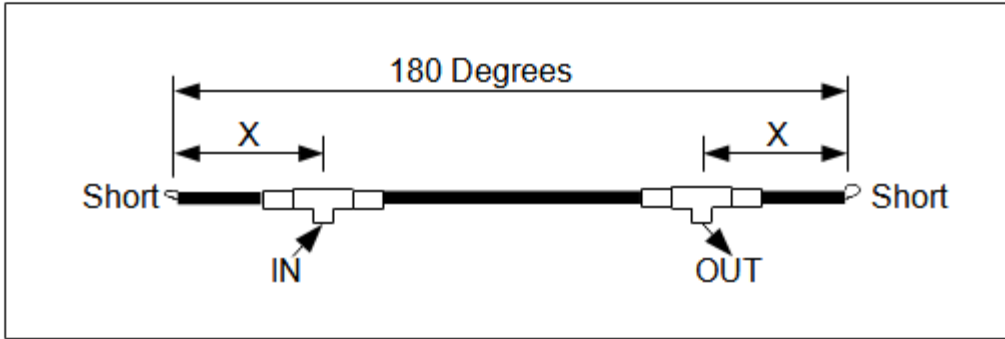


Figure 33. Half wave resonator

The two tap points must be symmetrically positioned in order to minimize the VSWR. This gives the selectivity of two resonators coupled together. See the 21 MHz version in 34. Note the greatly increased selectivity. The attenuation at 14 MHz is now about 30 dB and about 40dB at 7 MHz. At 28 MHz the loss has increased slightly to about 18 dB. The deep null has disappeared.

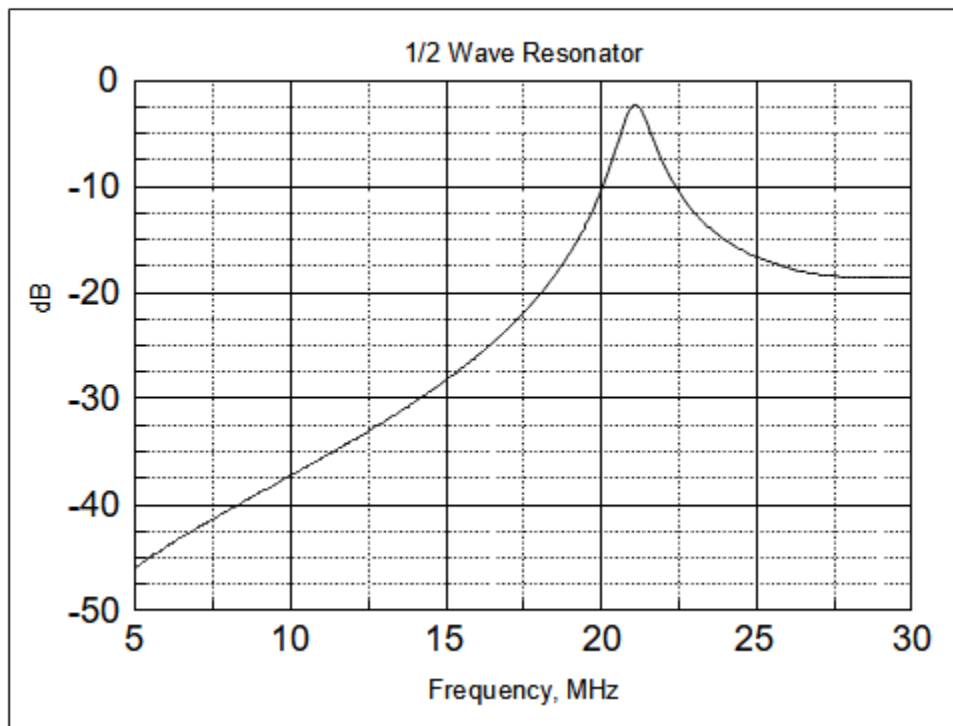


Figure 34. Frequency response of the half wave resonator.

This is now looking useful. It has substantial rejection of lower frequencies. At 14 MHz it's about equal to a single $\frac{1}{2}$ wave open stub. However, it is superior for rejecting the lower frequencies. Note that there will be high rejection in the AM broadcast band.

One of these resonators was built out of RG-213 cable. The lengths were: L2=10" and 2L1=13' 9". Overall length was 15' 5". The center frequency was 21.1 MHz and the loss was 2.17 dB. The 2 dB bandwidth was 21.0 to 21.2 MHz. The rejection at the other ham bands was within 1 dB of the simulation curve values in **Figure 34**. This type of resonator gives good rejection to the lower frequencies, but has too much loss to use in the transmitting line. It can be used in the receive line to help with low frequency overload problems in a multi transmitter environment.

3.4.4 Stub Shorthand

We have developed a shorthand system for describing stubs at Top Ten Devices. When taking data we use these designations on the graphs made for customer cables. The following parts of this book make use of some of that data to illustrate the text. The shorthand is also used in some of the tables and schematics which follow.

Designation	Bands Passed	Bands Nulled	Type	Cut Band
CS-1	80	40,20,15,10	1	$\frac{1}{4}$ WL-S 80
CS-2	40	80	2	$\frac{1}{2}$ WL-O 40
CS-3	40/15	20,10	1	$\frac{1}{4}$ WL-S 40
CS-4	20/10	40,15	2	$\frac{1}{2}$ WL-O 20
CS-5	20	10	1	$\frac{1}{4}$ WL-S 20
CS-6	10	20	2	$\frac{1}{2}$ WL-O 10
CS-7	40	15	3*	$\frac{1}{2}$ WL-S 15
CS-8	40	15	3*	$\frac{1}{4}$ WL-O 15
CS-9	160	80,40,20,15,10	1	$\frac{1}{4}$ WL-S 160
CS-10	80	160	2	$\frac{1}{2}$ WL-O 80

* CS-7 and CS-8 Connected in parallel make up a type 3.

The right hand column describes the stub construction as, for example: $\frac{1}{4}$ Wave Length - Shorted 80 meters.

3.5 Effects of Various Cables

LMR-400 would seem to be a better choice than RG-213, but this is not the case. In 2003 when I made the initial measurement on a $\frac{1}{4}$ wave shorted 20 meter stub made from new LMR400, it had a null depth of 32 dB. This was 2 dB less than the RG-213 stub. I resurrected this LMR stub from the garage, where it's been stored for 5+ years and measured it again. Strangely, it only measured 26 dB at the null this time. I flexed the cable over it's entire length and the null moved somewhat and settled at -28 dB or 4 dB worse than the RG-213. Possibly the two separate shields inside the cable move around with flexing and this seems to change the attenuation. I suggest that LMR400 should not be used for stubs.

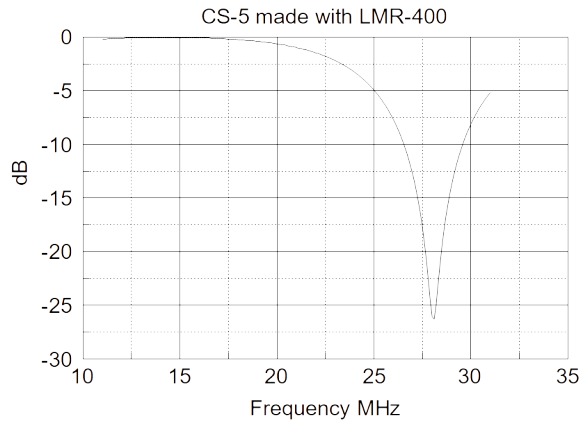


Figure 35. LMR-400 performance.

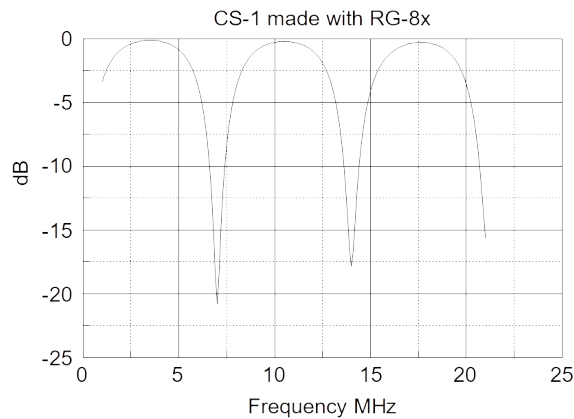


Figure 36. RG-8X performance.

Compare the RG-8X stub shown in **Figure 36** with the 7MHz performance of RG-213 shown in **Figure 21**. There is 4 to 6 dB less attenuation when stubs are made from RG-8X as compared to RG-213. When space or weight is a primary consideration, RG8x does a decent job of attenuating harmonics. At the lower end of the HF bands RG-8X will withstand full legal power. It will warm up when used on the higher bands. To prevent overheating, the stubs should not be coiled up too tightly or stuffed into a small space. A set of double stubs was made up for use at HC8N with RG-8X. Since I was carrying them to the Galapagos Islands, they were packaged into several small aluminum chasses with coax connectors. Each band had a dedicated stub assembly. They worked fine for the first few contests, but after a year or more of use, some shorts developed in the coax. They got a bit warm, even with the 1000 to 1200 watt power level in use at the station, and the center conductor drifted over to the shield in the soft foam insulation.

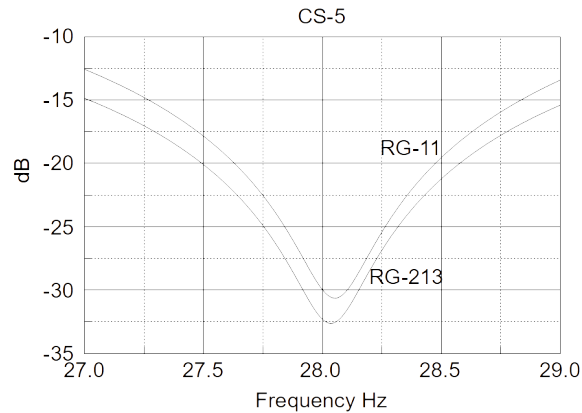


Figure 37. RG-11 compared with RG-213.

Note the slightly reduced attenuation provided by RG-11 when compared with RG-213 shown in **Figure 37**. This may be due to the smaller center conductor in RG-11. There is no mismatch caused by using 75 ohm cable for stubs.

3.6 Using Multiple Stubs

There are a number of different ways to increase the attenuation provided by stubs. The simplest way is to put two stubs in parallel on a tee connector, but this is not very effective, as it simply adds 6 dB to the null by halving the effective shunt resistance. There are some other ways to use two stubs that will double the attenuation of one stub in dB and also add another 6 dB. Thus if a stub has 27 dB of attenuation it is possible to get 60 dB by using two of them coupled with a line or a lumped element.

3.6.1 Type 1 Stubs Coupled with a Coaxial Line

Figure 38 shows the general method of coupling two Type 1, $\frac{1}{4}$ wave shorted stubs. As CC, the coupling cable, is increased in length from zero, the additional attenuation increases rapidly from 6 dB upwards. Interesting things occur with the null depths at the various harmonics as CC is varied.

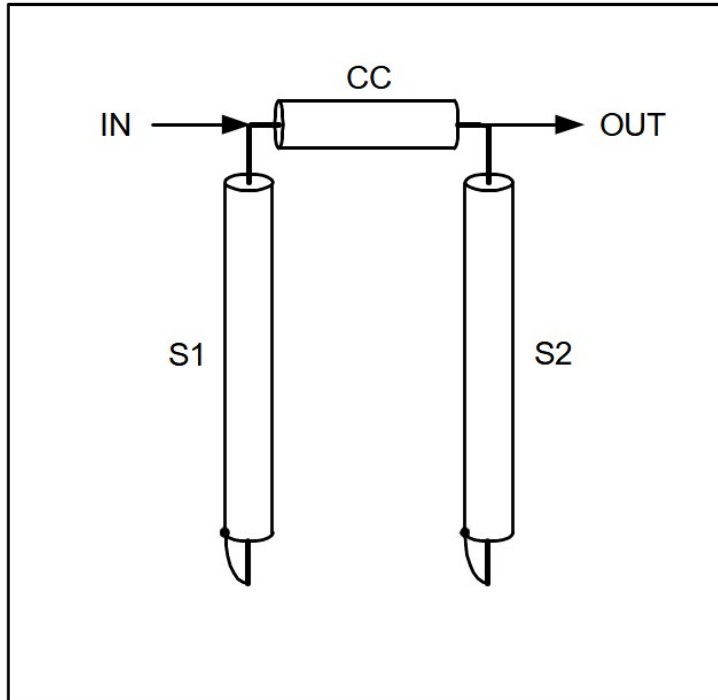


Figure 38 Double stub arrangement.

Table 14, below, shows what happens. CC is measured in wavelengths at the fundamental frequency. The values in the table were obtained by simulating a pair of 80 meter, $\frac{1}{4}$ wave shorted stubs made from RG-213U. So the 2F is 7MHz, 4F=14MHz, 6F=21MHz and 8F=28MHz.

Table 14 Effect of varying the coupling cable length.

CC Length	2F	4F	6F	8F
1/24wl	-70dB	-74	-80	-74
1/16	-73	-76	-80	-42
1/6	-74	-75	-78	-75
1/8	-75	-43	-80	-43

Conventional wisdom has always said the coupling cable (CC) should be $\frac{1}{4}$ wavelength at the second harmonic to minimize that component. That would be $\frac{1}{8}$ wavelength at the fundamental, of course. We can see that the second harmonic is minimized with that length, but let's look at the other null depths. At 4F and 8F the null is just -43dB, whereas with a $\frac{1}{24}$ or $\frac{1}{6}$ wavelength coupler they are down 74 or 75 dB. Therefore, we can save some coax and improve the performance of coupled stubs by a large margin if we use $\frac{1}{24}$ or $\frac{1}{6}$ wavelength couplers. The actual null depths obtained in practice may be a bit less than those shown in the table, but the relative depths will be similar.

Figure 39 and 40 show actual measurements on a pair of $\frac{1}{4}$ wave shorted 40-meter stubs with several coupling cable lengths.

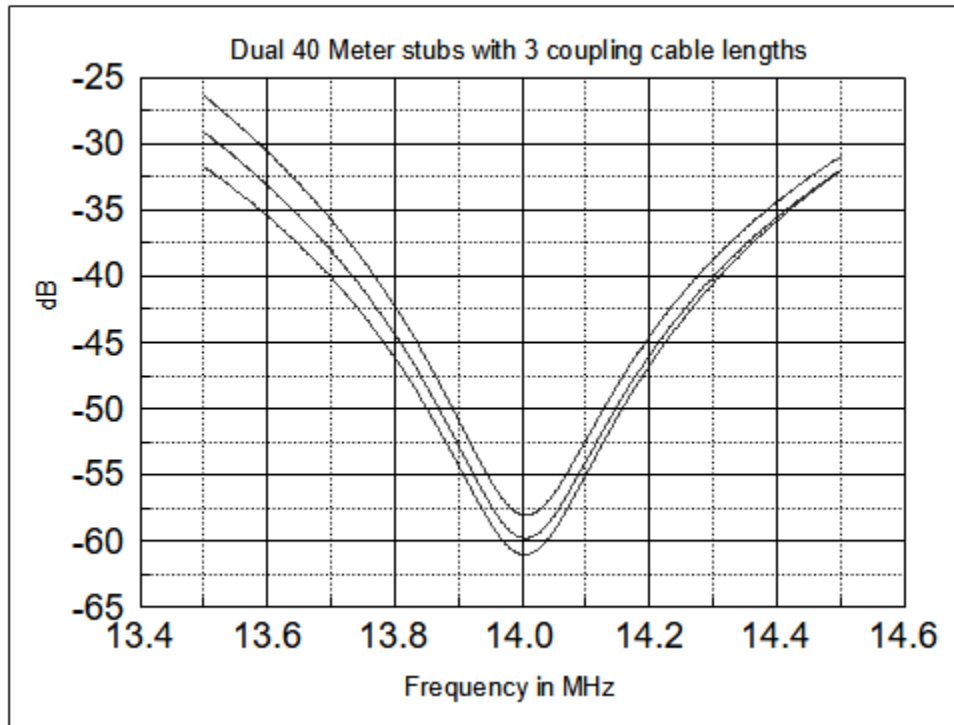


Figure 39. *Three coupling cable lengths and the effect on the 2nd harmonic.*

The lengths in wavelengths at 7 MHz used for the three curves are as follows:
Top=1/16, mid=1/12, bottom=1/8.

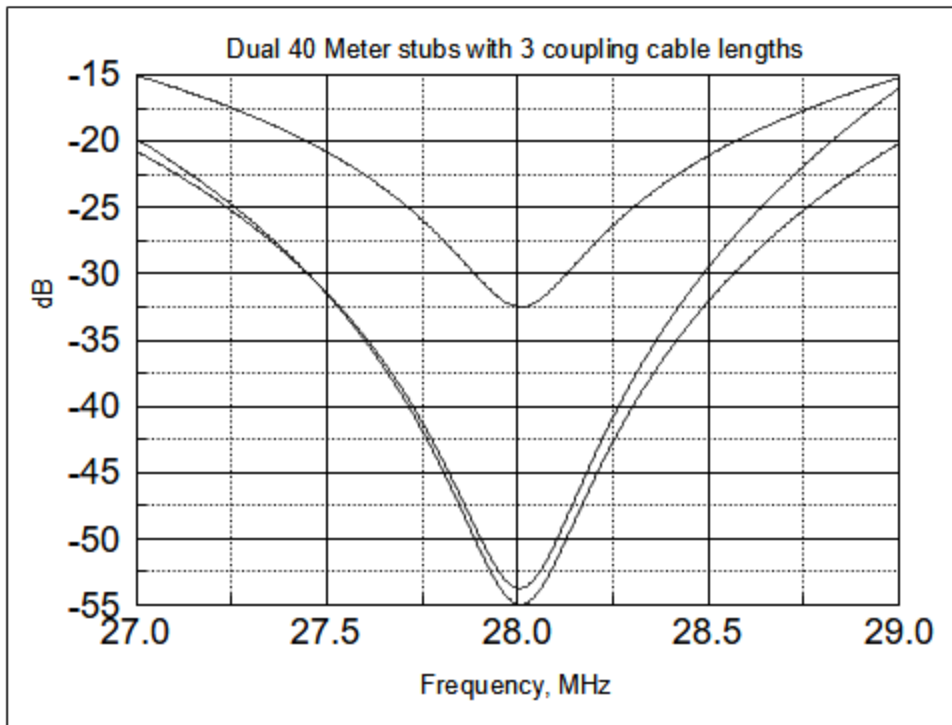


Figure 40. Three coupling cable lengths and the effect on the 4th harmonic.

The lengths in wavelengths at 7 MHz used for the three curves are as follows: Top=1/8, mid=1/16, bottom=1/12.

Table 15 Effects of varying coupling cable length.

CC Length	14 MHz	28 MHz
1/8 wl on 7MHz	-62	-32.5dB
1/12 wl	-60	-55
1/16 wl	-56	-54

This clearly shows that we want to use 1/12 wave couplers to minimize the total harmonic radiation for 2F and 4F. In this case the higher harmonics above 4F are not important.

For 80 meter assemblies, we might use 1/32 wavelength to minimize 2F through 8F total harmonic energy.

3.6.2 Type 2 Stubs Coupled with a Coaxial Line

Half wave open stubs for sub harmonic nulling can also be coupled for greater attenuation. In this case, conventional wisdom says the coupler should be 1/4 wave on the sub harmonic. For a pair of 20 meter half wave open stubs (type 2), the coupling cable should be the same as the stub length, 23.4 feet or 1/4 wave at 40 meters. While this will give the best performance on both 40 and 15 meter

nulls, (see **Figure 41**) a shorter length of 7.5 feet also works well, giving up only 4 dB of null on 40. The null depth on 15 is more critical of coupler length and giving very poor performance when equal to a half wave or a full wave on 15.

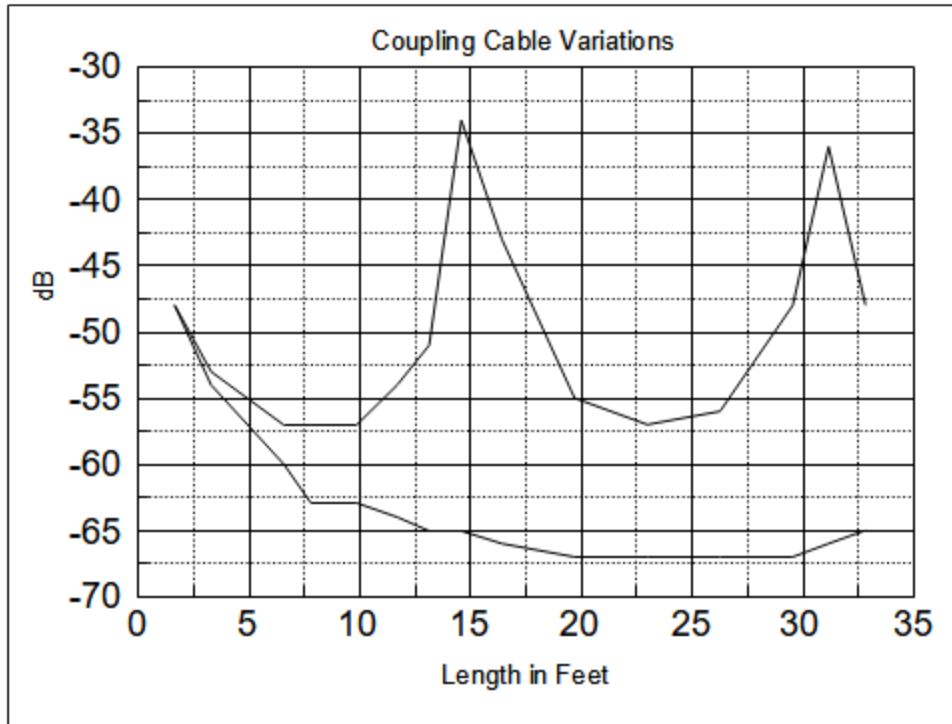


Figure 41. Attenuation vs. coupling cable length for a pair of 20 meter open $\frac{1}{4}$ wave stubs. The bottom curve is for 40 meters and the upper one is for 15 meters.

Figure 42 shows a pair of 20 meter open $\frac{1}{4}$ wave RG 213 stubs coupled with a $\frac{1}{4}$ wavelength 20 meter line for use on 10 meters. This coupling line could be made shorter with little loss of null depth.

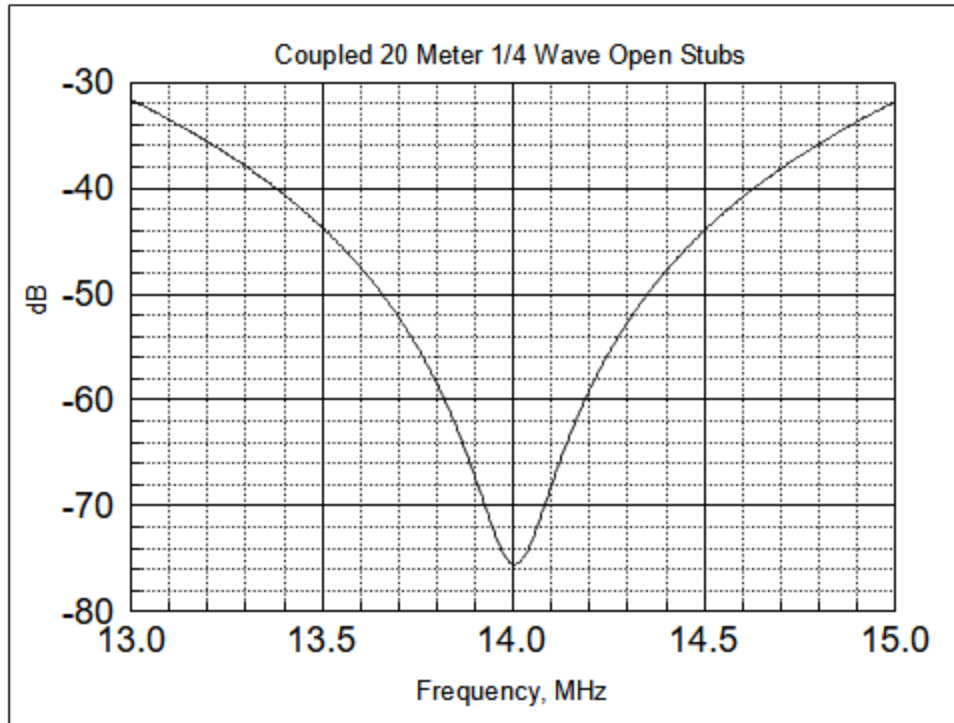


Figure 42. A pair of 20 meter 1/4 wave open stubs coupled with a 1/4 wave line.

3.6.3 Combinations for SSB and CW

Coupled stubs can be cut for different frequencies to widen the null. The obvious use for this is to cover both the SSB and CW bands. This is particularly true for 40, 20 and 15 meters where the sub bands are not that far apart. On 80 and 10 meters the result will be a double null with some reduced attenuation in between. When cutting these stubs it is best to cut one first, then assemble the coupler and second stub for trimming. That is, don't cut the stubs one at a time and then assemble, as there is some interaction when they are coupled. **Figure 43** shows a set of RG-213 stubs cut for 40 meters. Curve 1 shows a pair of stubs cut to the same length for 7.075 to null at 14.150. Curve 2 shows one stub cut to null at 14.000 and the second cut to null at 14.300. While the response over 14.000 to 14.300 is almost flat for curve 2, there is no advantage attenuation wise. As a matter of fact, for 20 meter rejection of both CW and SSB, the best method is to cut both for the band center.

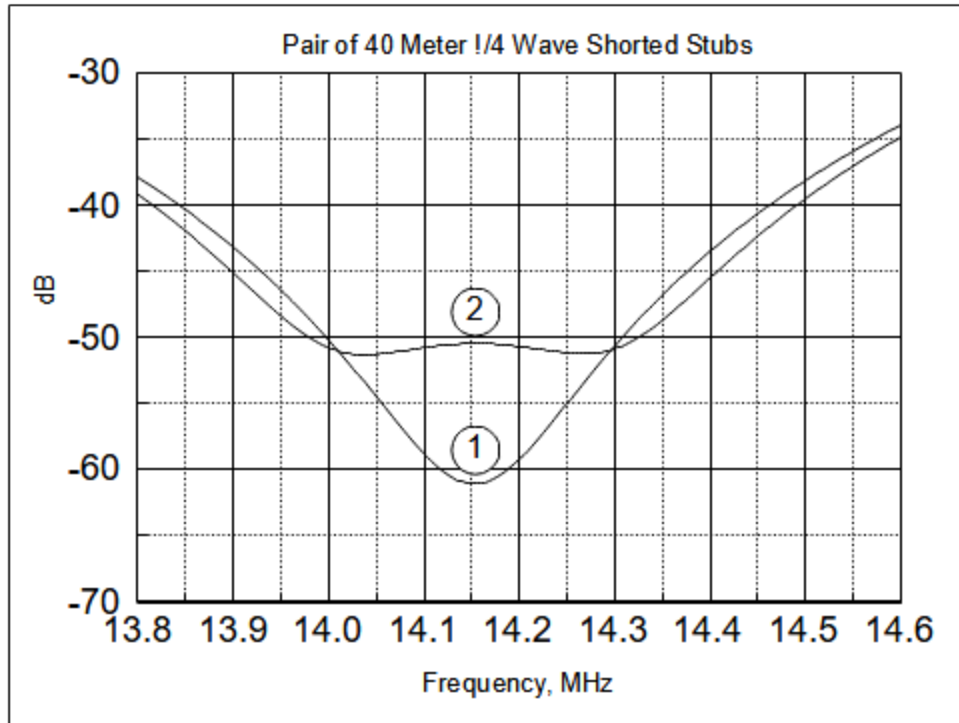


Figure 43 Double 1/4 wave shorted stubs for 7 MHz. See text.

In **Figure 44** we see 2 similar curves, but made with 20 meter 1/4 wave shorted stubs for 10 meter nulling. Curve 1 has two stubs cut for 28.25 MHz. Curve 2 has one cut for 28.05 and the other cut for 28.45 MHz. In this example the stagger tuned stubs actually have a bit more attenuation at the edges of the band of interest. If one is interested in either CW or SSB, the best set up is with both tuned to the same center frequency. Coupling cables between 5 and 6.5 feet long will give results as shown.

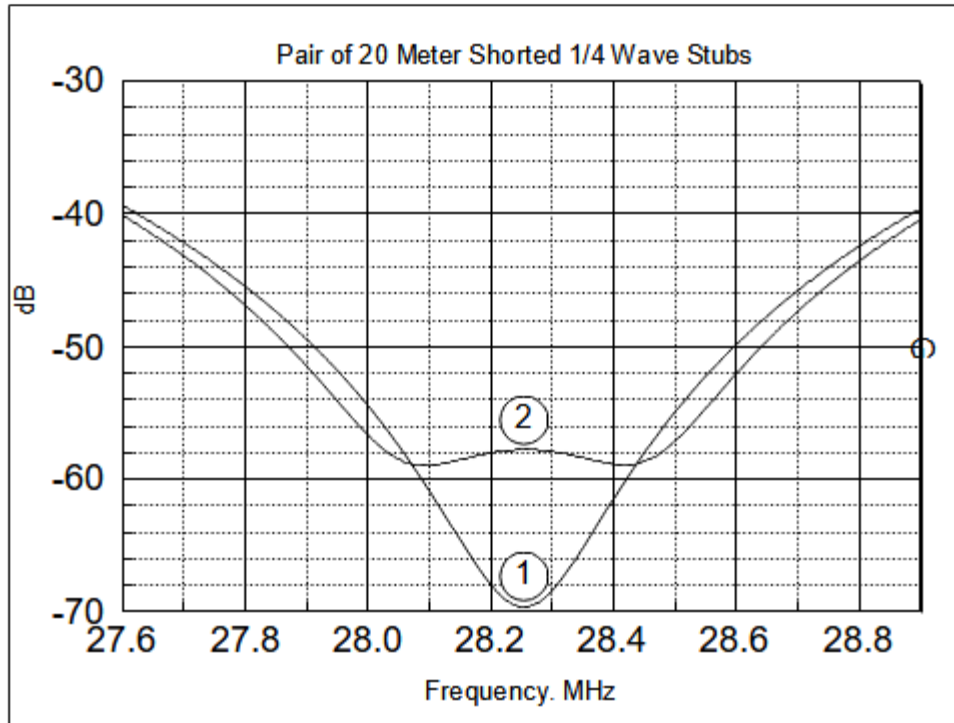


Figure 44 A pair of $\frac{1}{4}$ wave 14 MHz stubs. 1. Both cut for 28.250. 2. One cut for 28.025 and the other cut for 28.450.

3.6.4 Connecting Multiple Stubs

Stubs for various harmonics and sub harmonics can be connected together with tee connectors. The way they are connected can have some effect on how they perform. The best connecting method is one that has no common portion of the transmission line shared between stubs. Put another way, stubs should be connected directly to the transmission line between line and load, and not to another stub. This is best illustrated as shown in **Figure 45**. **Figure 45a** shows two stubs attached to a transmission line going from an amplifier to an antenna. Each stub connects directly to the transmission path. In **Figure 45b**, the same number of adapters is used, but there is a common path between the stubs to the transmission path. This common path can cause interaction between the stubs. For example: stub A is cut to null 28,025 kHz and stub B is cut to null 28,550 kHz. When the two are connected as in **Figure 45a**, they function as designed. When connected as in **Figure 45b**, the nulls are displaced.

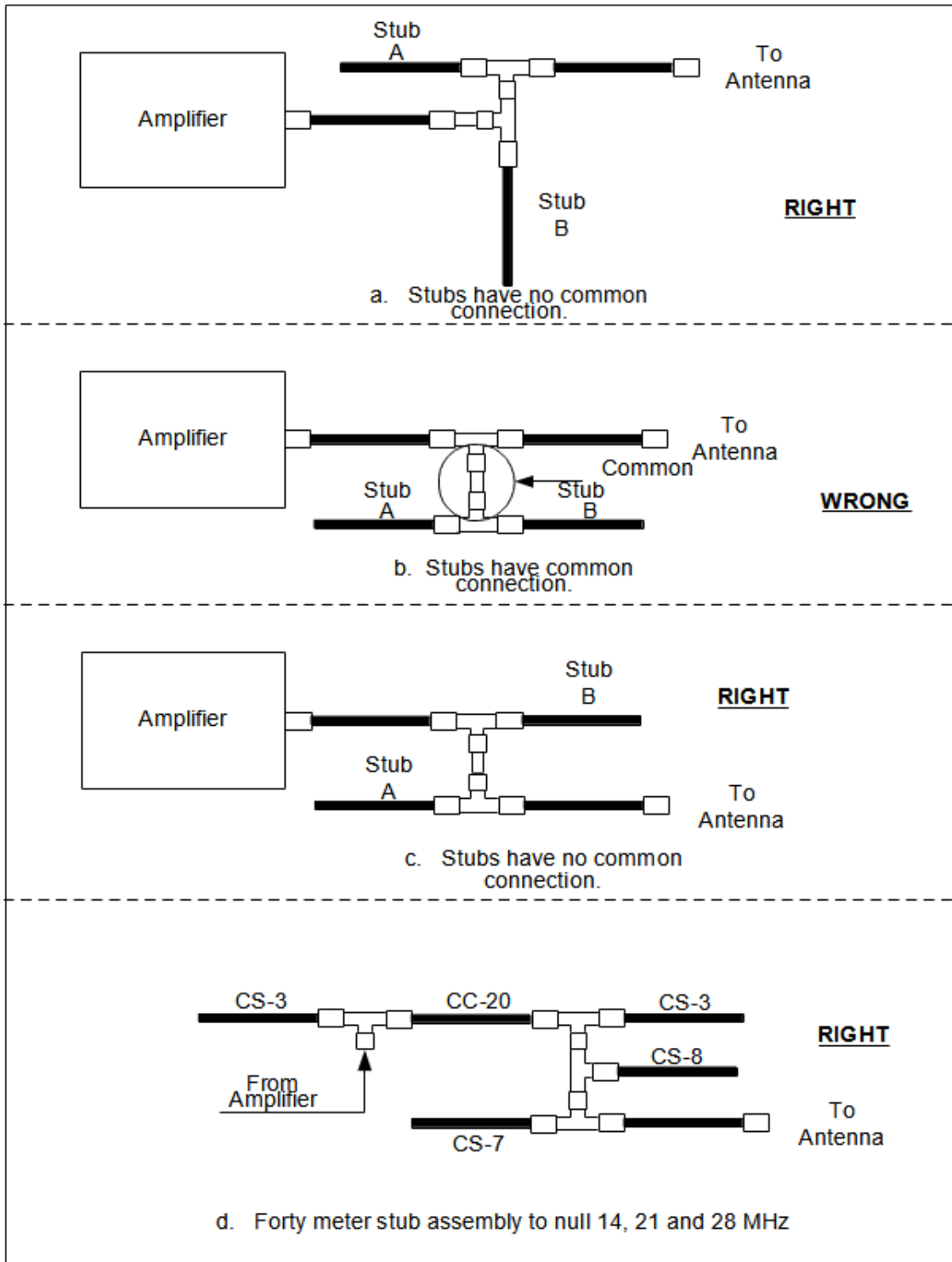


Figure 45 Various ways to connect a pair of stubs.

Figure 46 shows a spectrum analyzer sweep of two CS-5 stubs cut as described above. The actual frequencies came out closer to 28,000 and 28,450 kHz. When connected as shown in **Figure 45a**, there is some interaction, but the nulls stay on the frequencies for which they were cut. The lower null is reduced a few

dB and the upper null is increased a similar amount. See Curve b in **Figure 46**. If the length of line connecting the two together is increased from an inch or two to a few feet, the nulls will merge and become some 50 dB deep while staying at the original frequencies, as shown in the section on combinations for SSB and CW.

The connection shown in **Figure 45b** results in the shifting of both nulls in a lower frequency direction. Any increase in the length of common coupling between the stubs will lower both null frequencies. See **Figure 46** curve b. Note that the nulls are shifted almost 250 kHz. Simply swapping positions of the antenna line and one of the stubs will remove the common connection, as shown in **Figures 45a** and c. This connection has the same frequency response as **Figure 46** curve a.

This basic arrangement of two stubs is used as an illustration of interaction between stubs. Two stubs with nulls very close together were chosen for this example. It becomes more critical when connecting multiple stubs. The more tees and barrels required for connections, the more confusing it becomes to avoid the common connection between stubs. **Figure 45d** shows a 40 meter, single band setup which might be used for a multiple transmitter contest station. The pair of CS-3s produces nulls at 14 and 28 MHz for the 2nd and 4th harmonics, while the CS7/8 null the 3rd harmonic at 21 MHz. Note the connection methods for eliminating common sections between stubs.

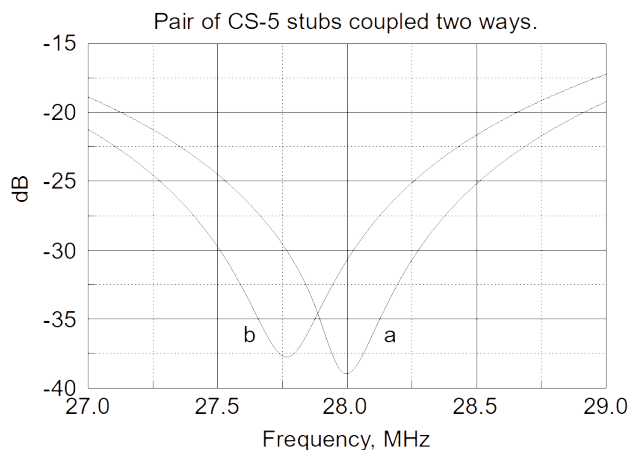


Figure 46 Two stubs connected as in **Figure 45a** and **b**.

Each stub must be cut individually with a tee in line.

3.7 Losses and VSWR

If stubs are cut accurately for the operating frequency range, the loss will be very low. A pair of CS-5s cut for 20 meter SSB and CW service show a loss of 0.08 dB over the 14 to 14.35 MHz band. The range for less than ½ dB loss extends from 11 to 15 MHz.

The VSWR at the operating frequency is 1.00. The frequency range for the VSWR to be less than 1.10 is 13.1 to 14.9 MHz.

Properly cut stubs have essentially zero effect at the operating frequency. See **Figure 47** for another example.

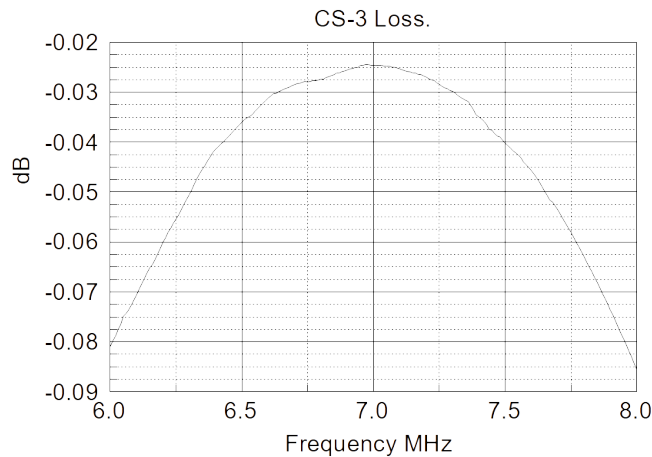


Figure 47. Forty meter loss for a single CS-3.

3.8 Band switching stubs

When one transmitter is to be used on more than one band, it becomes necessary to switch the stubs. There are many ways to do this. In **Figure 48** the stubs are selected separately from the antennas. One advantage here is that multiband antennas are easy to accommodate. The bandswitching stub assembly could be located in the shack while the antenna switch is at the tower.

When a relay or manual switch is used to select stubs as shown, the stub length must be adjusted for the equivalent length in the switch. This is best done with the stub attached to the switch while making the final cut of cable.

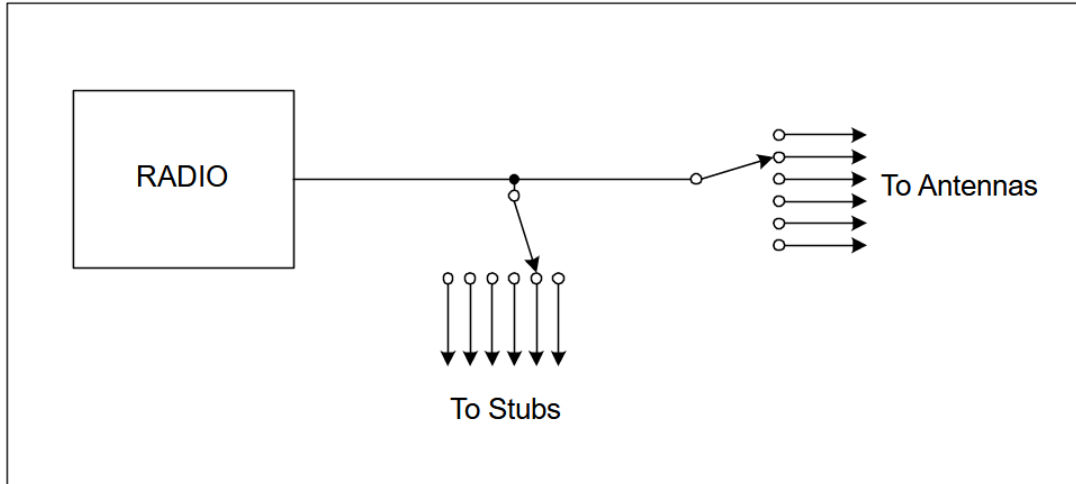


Figure 48 Separate switches for stubs and antennas.

In **Figure 49**, the stubs are placed on the coax lines going out to monoband antennas after the antenna selector switch. The switch could be in the shack or out at the tower and remotely controlled.

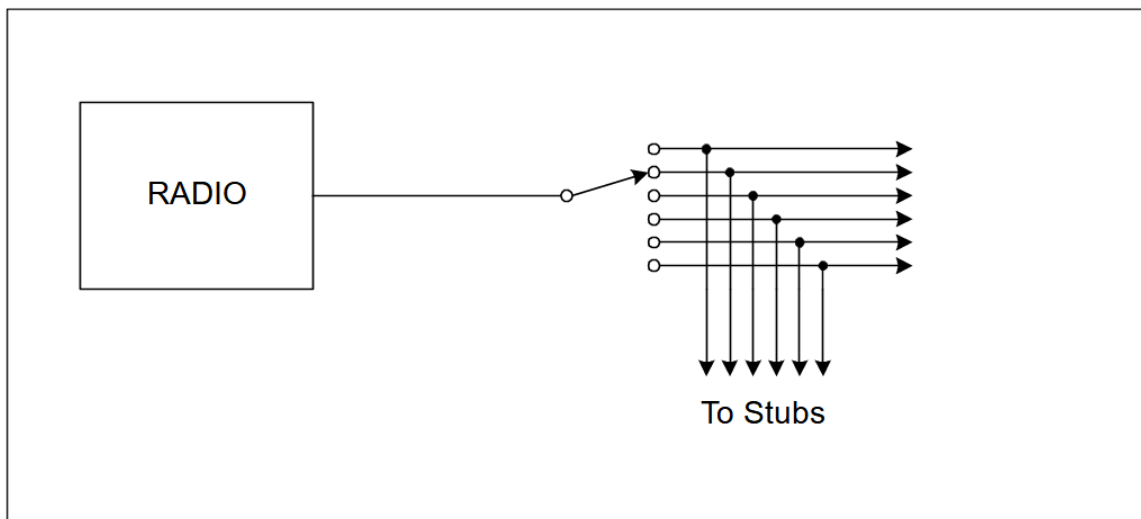


Figure 49. Common switching of stubs and antennas.

Band switching stub assemblies can be made up in several differing combinations. Either **Figure 48** or **49** can be used for the actual switching. One possible assembly type is shown in Table 16. It can be made with a 6 way relay switch and it provides some attenuation for all six contesting bands as shown in Table 17.

Relay	160	80	40	20	15	10
1	CS1	CS1				
2	CS2		CS2	CS2		
3			CS5	CS5	CS5	
4			CS6		CS6	CS6
5				CS4		CS4
6			CS7/8			

Table 17 Bands nulled with method 1

Band passed	Bands nulled
160	80
80	40,20,15,10
40	80,20,15,10
20	80,40,15,10
15	20,10
10	40,20,15

Method 2, Table 18, trades harmonic rejection on 15 meters for better rejection on 160. Other combinations can be arranged as desired. If a switch with more than six positions is used, additional stubs can be accommodated.

Table 18 Method 2.

Relay	160	80	40	20	15	10
1	CS9					
2		CS1				
3			CS3		CS3	
4				CS4		
5				CS5		CS5
6						CS6

The diode logic required to do the switching for these methods can be determined directly from the tables. Consider that a band decoder is driving the relay box and we are using method 1 bandswitching. Diodes would be connected from the band outputs to the relays as follows:

- 160 to 1 and 2
- 80 to 1
- 40 to 2,3,4,and 6
- 20 to 2,3, and 5
- 15 to 3 and 4
- 10 to 4 and 5

This is shown schematically in **Figure 50**. The relays ground all of the unused stubs. Diode protection is needed on each relay to keep the fly back surge from damaging the band decoder transistors. A 10 ohm resistor in the power line protects the pc traces from shorts on the power bus. A bypass capacitor keeps stray RF from getting back onto the power supply input. This schematic is for the Top Ten Devices Six Way Relay Box, but may be easily home built. The relays have 10 Ampere contacts and are very reliable. There are thousands of them in use worldwide. The band decoder requires a “sink” output for this schematic. That is, it must supply a ground to activate the desired relay. To accommodate a decoder with “source” outputs, the diodes all must be reversed and the +12V input should be grounded.

It is always a good idea to bypass diodes with a 0.01 uF capacitor to avoid acting as a mixer or harmonic generator from the strong signals present in transmission line systems.

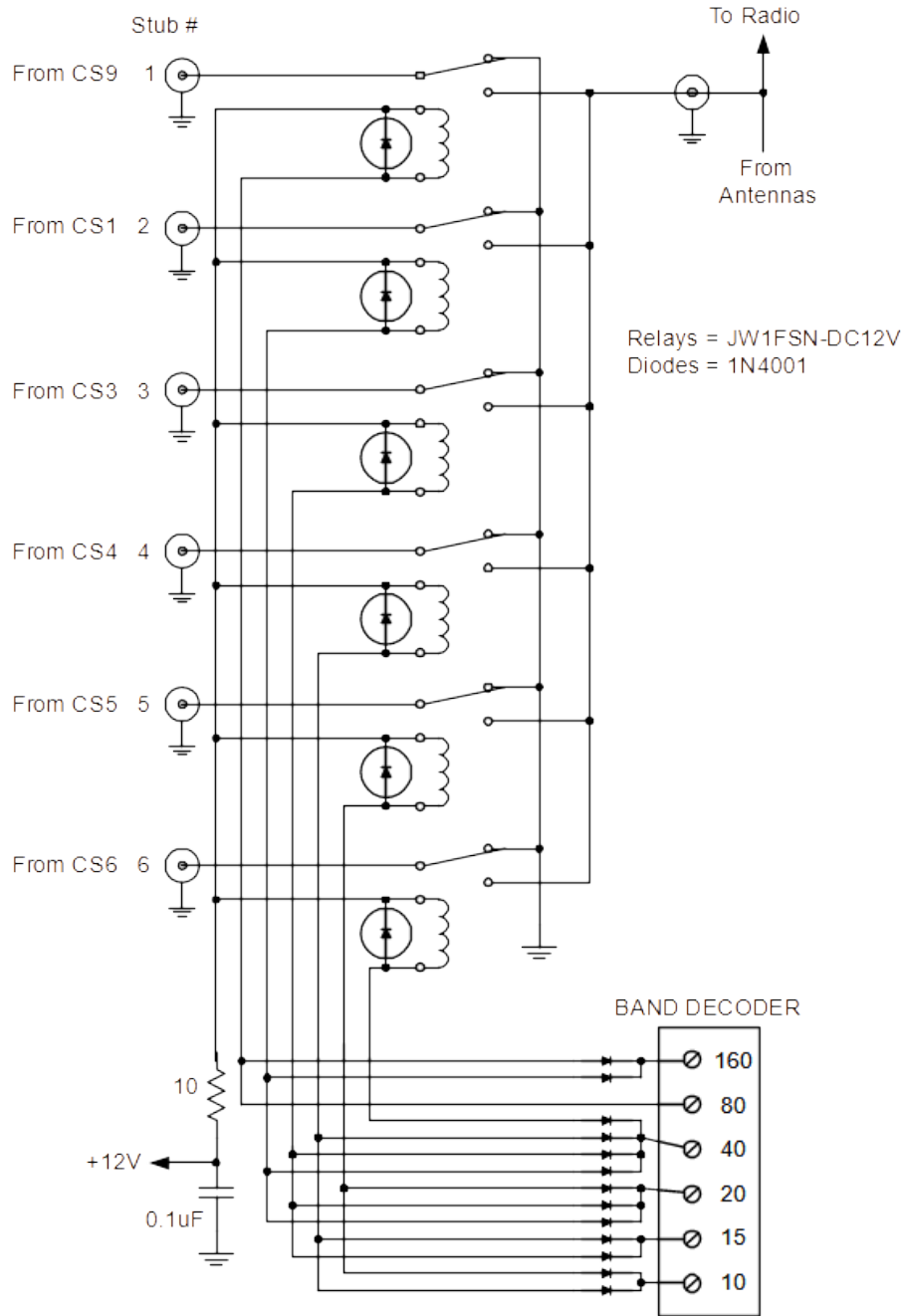


Figure 50. Bandswitching stubs by method two

Another method using a switched open stub is shown in Table 19. All normally closed relay contacts are grounded except for relay 2, which is open circuited.

Table 19 Method 3

Relay	160	80	40	20	15	10
1	CS2S		CS2S	CS2S		CS2S
2		CS2F		CS2F		CS2F
3			CS3	CS3	CS3	
4			CS4			CS4
5						
6	CS2	CS2				

Note. CS2S and CS2F refer to a two ended CS2 with a start and a finish.

3.9 Tuning Stubs with Lumped Elements

Adding a capacitor or inductor to a stub in various ways will shift the stub null frequency. Reasons for doing this include the following:

1. Very small frequency increments can be made which might be difficult to cut.
2. If a stub is inadvertently cut too short, it can be lowered to the desired frequency with a fixed inductor or capacitor.
3. Large changes in frequency can be made without changing the stub length.
4. One stub can be used on two frequencies under relay control; i.e. SSB and CW.
5. One stub can function as a shorted Type 1 or an open Type 2 and may be tuned to any frequency between the two nulls.

Several simulations were done to test the amount of variation obtainable. In all of the examples a 23' 4" stub was used. When shorted it has nulls at 14,000 and 28,000 kHz. When open it has nulls at 7,000 and 21,000 kHz. We examine the primary, or lowest, null in each case. **Figures 51a** and **51b** show two ways to inductively load a Type 1 stub. In both cases the frequency is shifted lower. The resulting shift in frequency is identical for the two circuits shown. The inductors used are very small. An inductor as small as 0.02 uH will shift the null in this example by 155 kHz at 14 MHz. An inductance of this amount can be produced with one inch of wire.

Figure 51c shows a series capacitor used to increase the frequency of a Type 1 stub. Large values of capacitance are needed to shift the frequency in small amounts. The capacitor can be placed in series with the stub input or it can replace the short circuit on its far end as shown in 51d. The same shift occurs as capacitance is varied in either case.

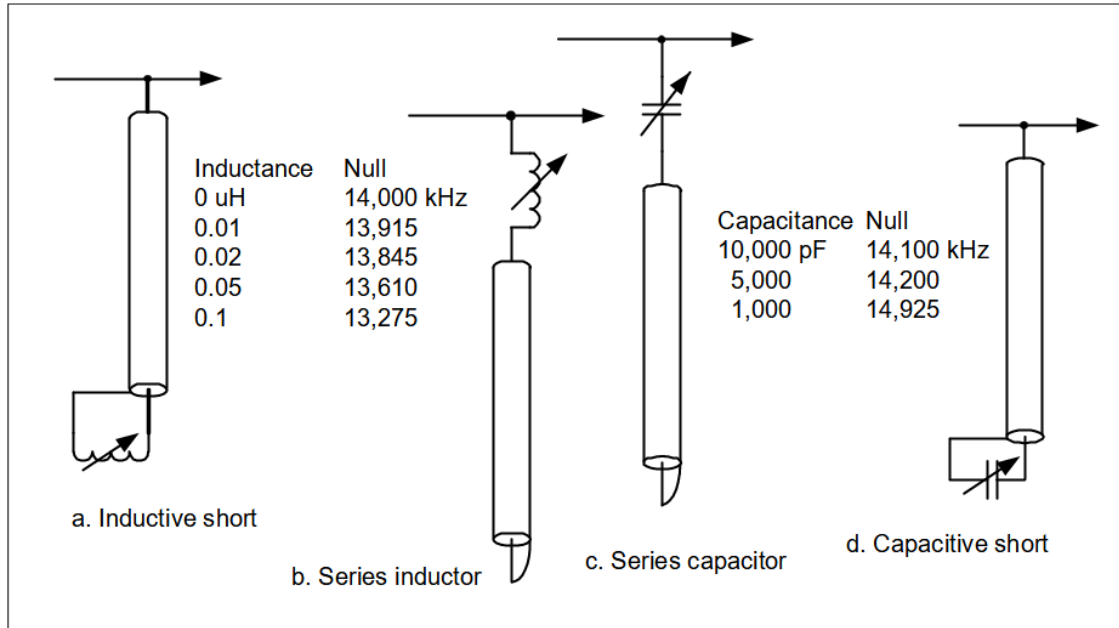


Figure 51 Tuning a shorted stub.

Figure 52 shows the tuning methods applied to a Type 2 stub. In all of the examples a 23' 4" stub was used. Again, when placed in series with the input to the stub, an inductor is used to lower the frequency and a capacitor is used to raise it. Results are shown in Fig 52 a and b.

In 52c a capacitor at the far end of the stub is shown. In this case the frequency is shifted lower by some very small values of capacitance. Note that this appears to be the same schematic as **Figure 51d**, but because the capacitor is located at a high-impedance point, the effect of the capacitor is different. If we start with a shorted, or Type 1 stub and replace the short with a very large capacitor, the null at 14 MHz moves up very slightly. However, with a short there is also a null at 0 frequency (dc). Replacing the short with a capacitor moves this null up in frequency to some low value. As the capacitance is reduced, the nulls move up in frequency. When the capacitance is reduced to the low pF range, the nulls are as seen in **Figure 51c**. When the capacitance goes to zero, the stub is a Type 2.

If high quality components are used for tuning the stubs, there will be little or no reduction in the null depths obtained without the lumped elements. Capacitor voltage ratings should be at least twice the expected voltage on the transmission line.

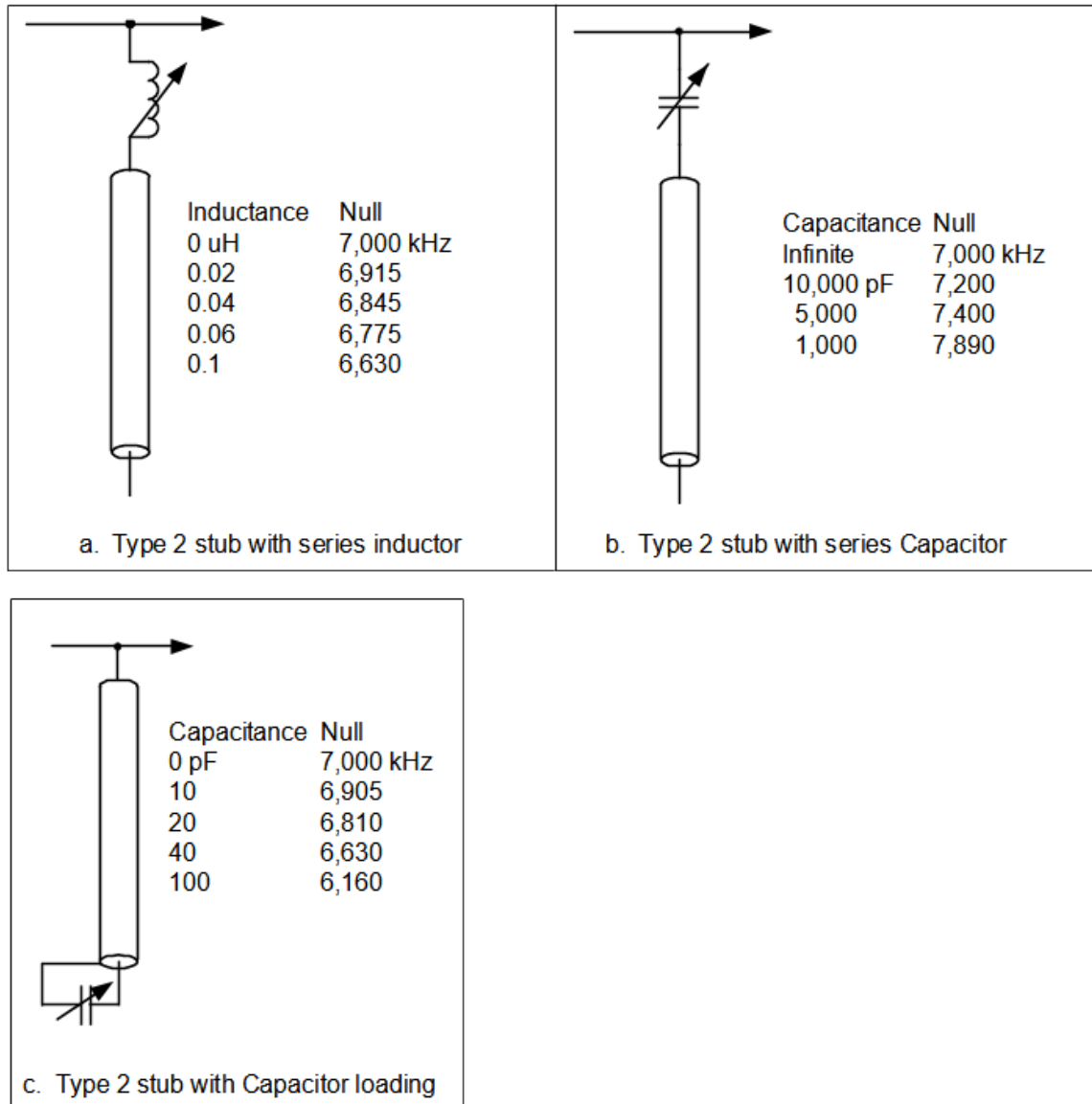


Figure 52. Tuning an open stub.

Figure 53 shows the response of a pair of Type 1, CS-3 stubs coupled with a 1/8 wave line. If both stubs are cut for nulls at 14,050 kHz, a -63 dB null occurs. When both are cut for nulls at 14,250 kHz, the same depth null occurs in the SSB band. If one of the stubs is cut for 14,050 and the other is cut for 14,250, the broad null depth is about -56 dB. With both stubs cut to null at 14,050 kHz and one has a 5,000 pF capacitor inserted as shown in **Figure 52c**, the double null produced is almost identical to that using two different cuts. When both stubs are cut for 14,050 and modified with 5,000 pF capacitors, the SSB null created is similar to cutting them for SSB. With relay selection of a short or a 5,000 pF capacitor at each end, any of the three characteristics could be obtained remotely.

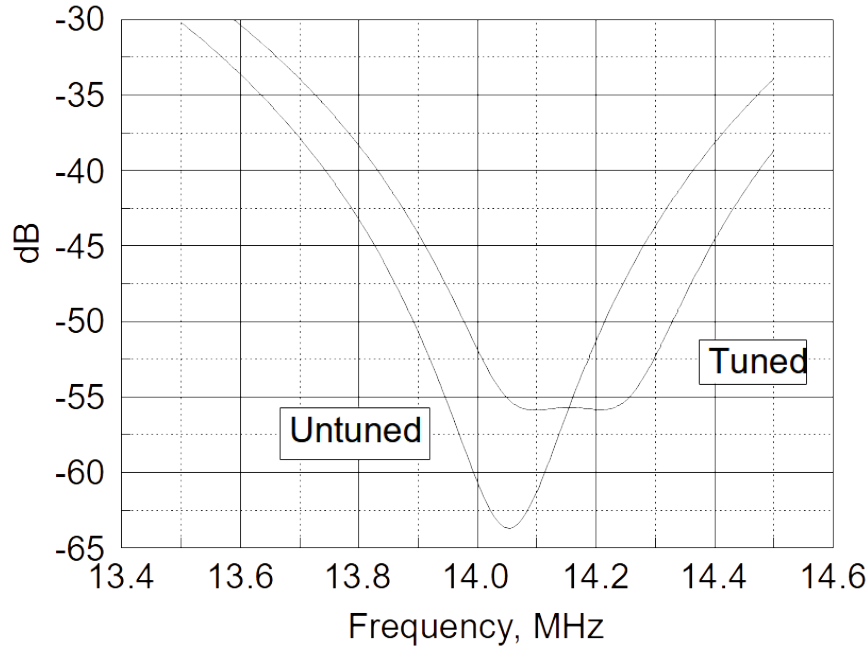


Figure 53. Tuning a pair of stubs.

4.0 FILTERS

4.1 Band pass Filters

Bandpass filters would seem to be the ideal way to improve isolation in the multi transmitter environment. Radio 1 is transmitting on band 40 and radio 2 is receiving on 20. The 40 meter signal needs to be reduced at the input of radio 2. A band pass filter on radio 2 which passes 20 and removes 40 should do the job. Radio 1 is also transmitting some harmonic energy on 20 which will interfere with radio 2, even with the band pass filter in place. Thus a second band pass which passes 40 and removes 20 meter energy before it is transmitted filter is needed on radio 1.

So there are two uses for band pass filters: transmitter harmonic reduction and out of band energy reduction for receivers. Protecting the receiver from out of band energy is not a difficult task. The energy levels involved are low, usually less than 10 watts. Filters can be designed using small, inexpensive components which will do a good job when placed in the receiver input. Transmitter harmonic reduction presents a more difficult requirement on the components needed. Most of the commercial filters available in the amateur market are designed for 100 or 200 watts maximum. While it is not impossible to design a band pass filter which will take 1500 watts, the components required for such a design are quite expensive.

The simplest band pass filter that is of any use is the two pole type. Several of the commercially available units for amateur use are two pole designs. Three

pole filters provide a bit more attenuation and are not much more difficult to design and build. While there are several circuit configurations that can be used, the two resonator, top coupled arrangement is probably the most common design. It uses a minimum of parts and the component value spread is not excessive.

Figure 54 shows the response of a home designed 160 meter band pass filter. This filter is a 2 element Butterworth type, with a series input circuit. It shows how easy receiving filters are. The design is can be done in Elsie in a few minutes. The pass band loss is not a problem for 160 meter receiving use. Construction is simple and the parts only cost a few cents. The schematic is shown in **Figure 55**.

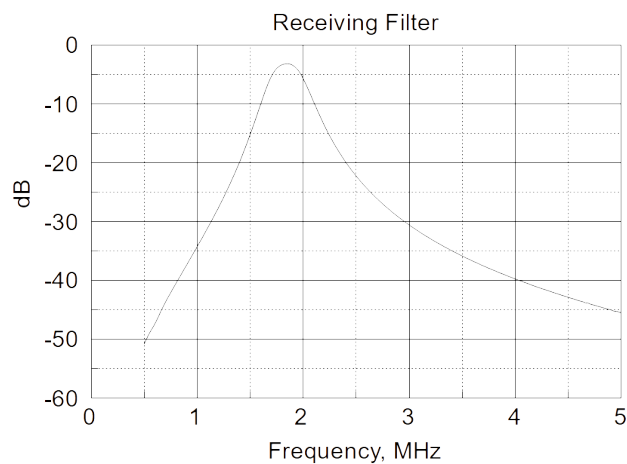


Figure 54. The 160 meter receiver filter.

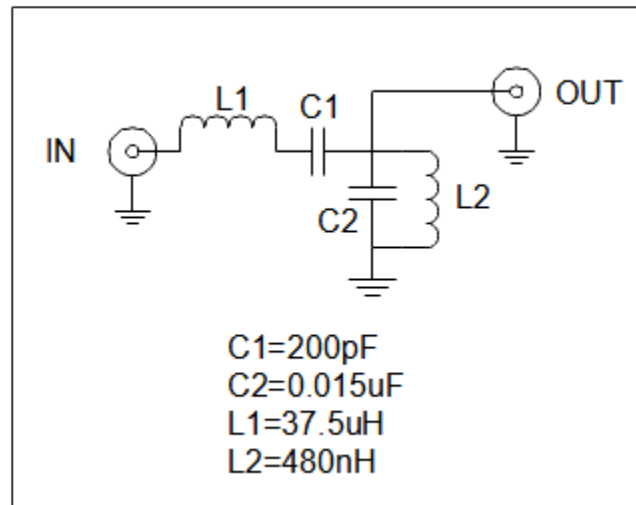


Figure 55. Receiving filter schematic.

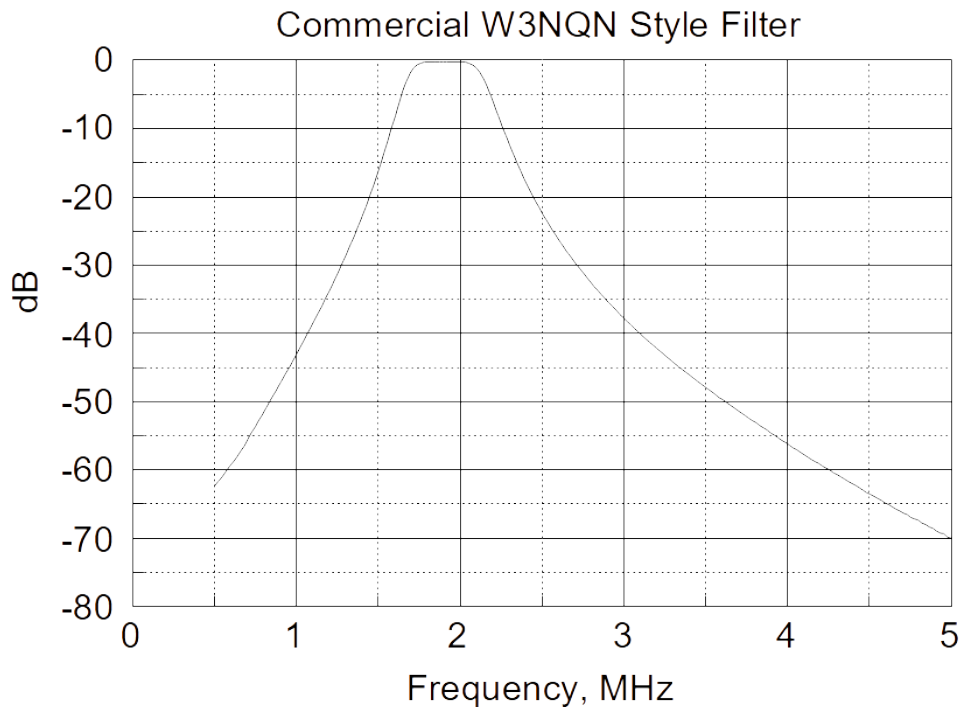


Figure 56. A three pole NQN style filter.

Figure 56 shows the response of a commercial single band 160 meter band pass filter. This design appeared in QST for May 1998. Powdered iron toroids are used to provide good Q and low passband loss.

Figures 57a through **57d** show the response of a type “I” commercially made band switching filter. **Figure 57a** shows the responses for 160, 80 and 40 meters with overlapping plots. The responses of these three filters continue into **Figure 57b**, which shows the 10 to 30 MHz range. **Figure 57d** shows the responses for 20, 15 and 10 meters. The response for these three filters in the 0 to 10 MHz range is shown in **Figure 57c**. The effects of the additional wiring and relays on the off frequency rejection in the stop band are clear. The ultimate rejection on this filter is much inferior to single band units.

Table 20 shows the band to band attenuation. The top line is the transmitting band and the left column is the receiving band. For example: The transmitting band is 40 meters, so the filter in use is 40. When receiving through the 40 meter filter, 80 meter energy is attenuated by 18 dB, etc. To meet our “no damage” goal of -45 dB, an additional 27 dB of isolation is required. Hopefully, the antenna to antenna isolation will exceed that.

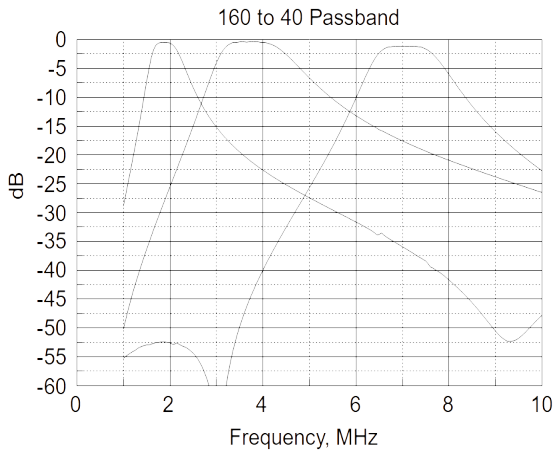


Figure 57a.

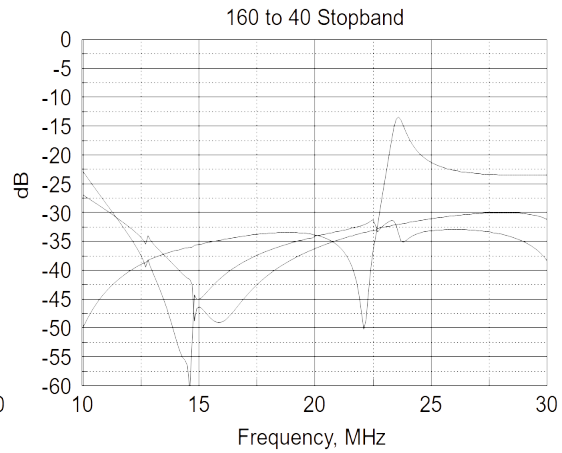


Figure 57b

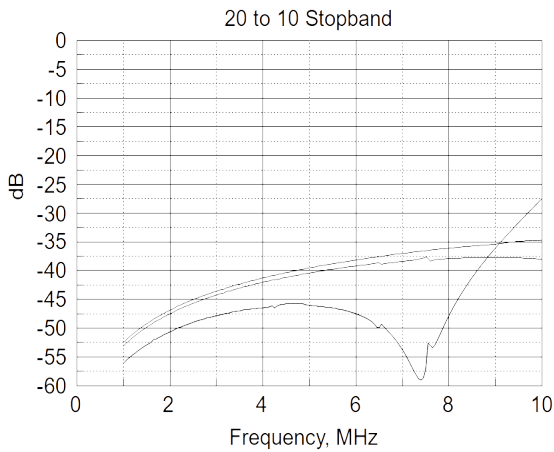


Figure 57c

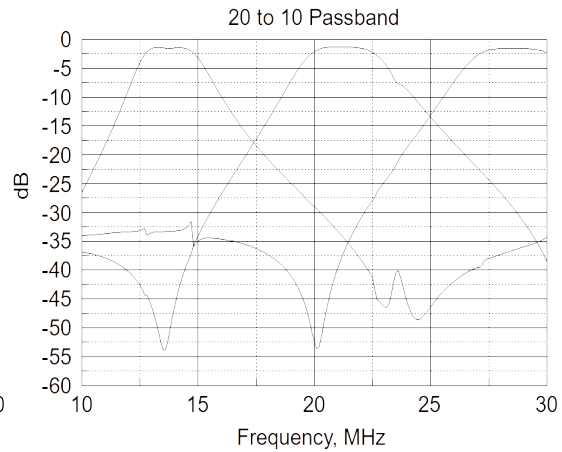


Figure 57d

The tables below summarize the isolation for each band. The top line is the transmitting band. The Rx band is shown in the left hand column. For example: in table 20 a receiver on 80 meters would have 18 dB isolation from the energy picked up on it's antenna from a 40 meter transmitter.

These filters are 2 element designs and as such do not have very much adjacent band rejection in many cases. There is also considerable "fly back", or reduced isolation due to parasitic effects such as stray inductance, capacitance and coupling between coils.

Table 20 Type "I" Commercial band switching filter response

Tx	160	80	40	20	15	10
Rx						
160	_	-20	-38	-36	-35	-23
80	-27	_	-18	-42	-36	-35
40	-50	-50	_	-50	-36	-35
20	-48	-44	-52	_	-34	-41
15	-47	-41	-37	-43	_	-29
10	-46	-40	-37	-36	-36	_

A type “D” commercial band switching filter was also tested and the result is shown in Table 21. It should be noted that only one sample of each filter was tested. Since they are hand wired, there will be some variations from unit to unit. This filter uses air wound inductors.

Table 21 Type “D” Commercial band switching filter response

Tx	160	80	40	20	15	10
Rx						
160	_	-34	-46	-30	-22	-18
80	-55	_	-34	-42	-29	-22
40	-50	-52	_	-43	-46	-43
20	-50	-45	-42	_	-57	-50
15	-57	-50	-56	-59	_	-38
10	-53	-51	-43	-50	-33	_

4.2 Hybrid Low Pass Filters

Low pass filters using standard inductors and capacitors typically have a rather slow roll off above the cutoff frequency unless they are constructed with a lot of components. Since we are interested in harmonic reduction from our transmitters and not concerned with frequencies in between harmonics, stubs usually give better performance. It is possible to combine the standard low pass filter using lumped components with stubs to improve the performance at the harmonics. Low pass filters are composed of series inductors and shunt capacitors. Open circuited transmission lines having an equivalent capacitance can replace the capacitors.

Two stubs and an inductor are required for a 3 element filter. The design technique is to measure or calculate the capacitance of the stubs and then to find a filter design which uses that capacitance. Using the Elsie software is the best way to do this.

Let’s start with an example. I have two open stubs cut for 28.8MHz and I want to make a filter with them that passes 14MHz and nulls 28MHz. To calculate the capacitance we would first figure how long the stubs need to be.

$$\text{Length} = 163.525 / 28.8 = 5.678 \text{ feet.}$$

We are using RG-213 which has 29pF/ft capacitance.

Stub capacitance= $5.678 \times 29 = 164.7\text{pF}$.

But we will be connecting the stubs with a UHF tee connector that has a measured capacitance of 9.3pF by itself. So the total capacitance will be 174pF.

Next, we go into Elsie and design a 3 element low pass filter with shunt capacitor input and output. A Butterworth filter will have too much loss at 14MHz, so we start with a 0.1 dB Tchebychev design. We adjust the cut off frequency until the required C is 174pF. At a cutoff of 18.8MHz, the required C is 174pF. The schematic tells us we need an inductor of 0.486 uH. This is easy to make with a few turns of #14 wire and will self support.

A breadboard of this filter was constructed and tested with the N2PK network analyzer. The sweep is shown in **Figure 58A** and schematic in **Figure 58B**.

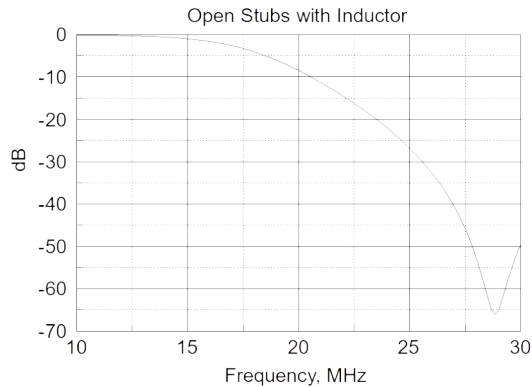
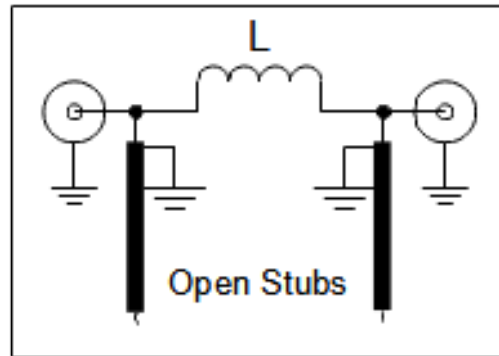


Figure 58A. Response.



58B Schematic.

The null depth is 66 dB, which is the same as we would obtain by coupling the two stubs with a coax line at 28MHz. The loss at 14 MHz measured 0.11 dB. The VSWR is very close to 1.0.

Table 22 shows the cutoff frequencies, inductance and RG-213 coax lengths required for each amateur band to duplicate this kind of performance. The numbers are for the low end of each band and can be scaled for frequencies up higher.

Table 22. 3 Pole filter design parameters.

Pass Band	Null Freq.	Coax Length	L (uH)	Cut Off Freq.
Up to 3.5 MHz	7 MHz	24.3'	1.99	4.59 MHz
Up to 7.0 MHz	14 MHz	12'	0.996	9.17 MHz
Up to 14.0 MHz	28 MHz	5.83'	0.498	18.35 MHz

Note that the open stubs will also null at their third harmonic, so the 3.5 MHz filter will null both 7 and 21 MHz.

It is also possible to null more than one frequency with this design. For example: one of the stubs in the 14 MHz design can be split into two. One cut for 21 MHz and the other for about 42 MHz to make up the difference in capacitance needed for low VSWR. Note that 10 pF must be added to each stub in the design for the Tee connectors required. The nulls will not be as deep as when two stubs on the same frequency are used.

This method of coupling stubs has one disadvantage as the coil must be mounted in a shielded box. Not only will the coil radiate, but it will have dangerous voltages on it when high power is in use.

4.3 Bandstop filters

What is a band stop filter? It operates in the reverse manner that a band pass filter does. When designed for a particular center frequency, it attenuates a band of frequencies around the design center. The function is similar to coaxial stubs, but the bandwidth and frequency parameters are readily changed through the design process.

Notch filters have been used for transmitter harmonic reduction in the amateur literature before. Notch filters generally use single pole LC filters or combinations of them for different bands. They typically have sharp nulls at the design frequency. There usually is no attempt to match the line impedance in simple designs. Filters which are matched to 50 or 75 ohms are easy to design and they will produce the expected results.

4.3.1 A 10 meter second harmonic filter

Figure 59 shows a band stop filter designed to reduce TVI from the second harmonic of a 10 meter transmitter. It consists of three tunable circuits as shown in the schematic. The frequency response is shown in **Figure 60**. The design was done with Elsie and started out with a Butterworth filter having a center frequency of 57MHz and a pass band of 20MHz. The design values were rounded to even numbers. This unit was assembled from parts at hand with no attempt to optimize by design. Small variable capacitors were used along with small coils of #16 wires. Loss at 28 MHz is 0.01 dB and it is a bit less at lower frequencies, so it can be left in line for transmitting full amateur power on all 9 HF bands. VSWR at 28 MHz is 1.03

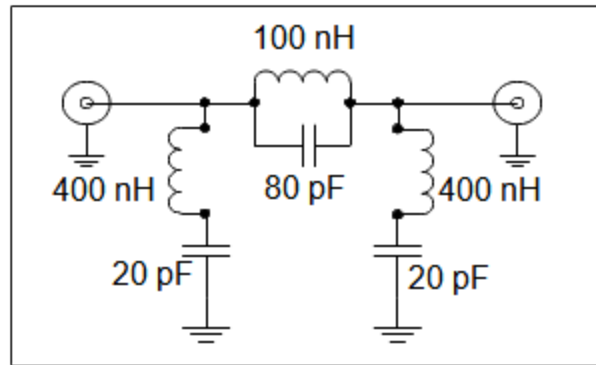


Figure 59. Channel two band stop filter.

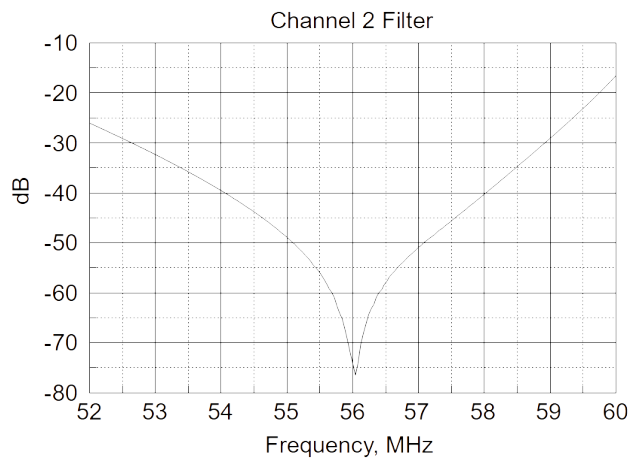


Figure 60. Measured frequency response of the channel two filter.

4.3.2 Filters for operating on 80 and 75 simultaneously.

I was asked to build some filters or stubs to enable operation on 80 and 75 by separate stations on the 2007 Ducie Island expedition. in early December 2007. There was concern by the operators that the close proximity of the Beverage antennas to the transmit antennas would cause receiving problems. As we have seen earlier, while some separation can be obtained with stubs, it is minimal and the losses are high. A quick preliminary design of a simple band stop filter with the Elsie program showed a possible solution. A three element Butterworth design could provide 40 dB attenuation at the null frequency and would not attenuate the desired listening frequency by more than a dB or so. The inductor requirements were modest, as a Q of 75 or more at the operating frequency would be adequate. Two of the required inductors were in the 50 to 60 uH range, while one was in the less than 500 nH range. A few commercially available inductors having values in the desired uH range were obtained and a selection was made based on Q measurements. The design was then tweaked to use the selected inductors. The nH chokes that were tested had miserably low Qs and were deemed not suitable. A few turns around a type 2 powdered iron toroid core showed more promise and was used in the final design.

To optimize the null depth each element (tuned circuit) must be adjustable. The shunt elements L1 and C1 require only a small variation in capacitance for tuning, and trimmer capacitors can do the job. The trimmers were padded with a series capacitor and a shunt capacitor to reduce the sensitivity of tuning as the null is very sharp.

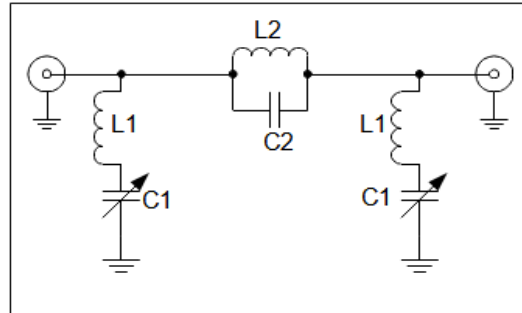


FIGURE 61.

Filter schematic.

Component	3501 Null	3784 Null
L1	55uH	55uH
L2	0.39uH	0.33uH
C1	36.7pF	31.6pF
C2	5280pF	5360pF

Note 1: C1 and C2 are made of several capacitors in series parallel.

FIGURE 62. Component values.

Construction was done on readily available prototype PC boards as there was not enough time to do custom work. Simple point to point wiring is adequate. Four units were built to meet the requirements, two for 80 and two for 75 meters.

Figure 63 shows the construction. They needed to be rugged enough to withstand the long journey via air, sea and ground to their operating positions on Ducie Island.

Small aluminum boxes were used. The boards were mounted on four threaded standoffs. Type F connectors were used according to the 75 ohm system requirements.

A three element filter is simple enough so that formal methods are not required, and alignment can be done by alternately adjusting each tuned circuit. Watching the loss at the desired pass frequency and the null depth on a network analyzer

while adjusting allows a good compromise to be found. **Figure 64** shows the sweep response of one of the units.

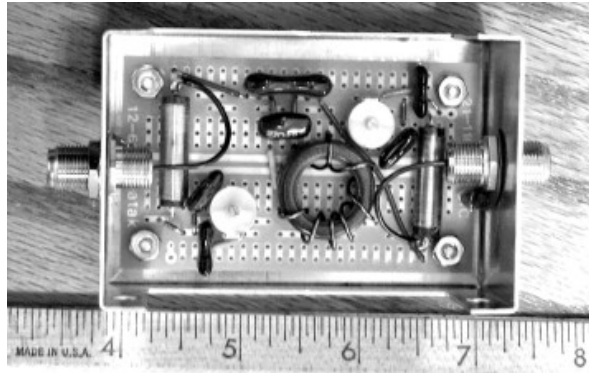


Figure 63 Construction.

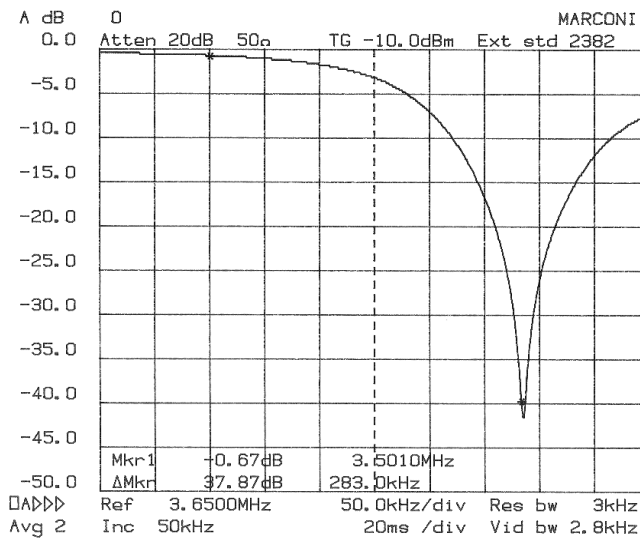


Figure 64. Typical sweep response

4.3.3 An 80 meter Band Stop Filter

Band stop filters can be designed to do the same type of jobs that are usually reserved for coaxial cable. The schematic of an 80 meter band stop filter is shown in **Figure 65**.

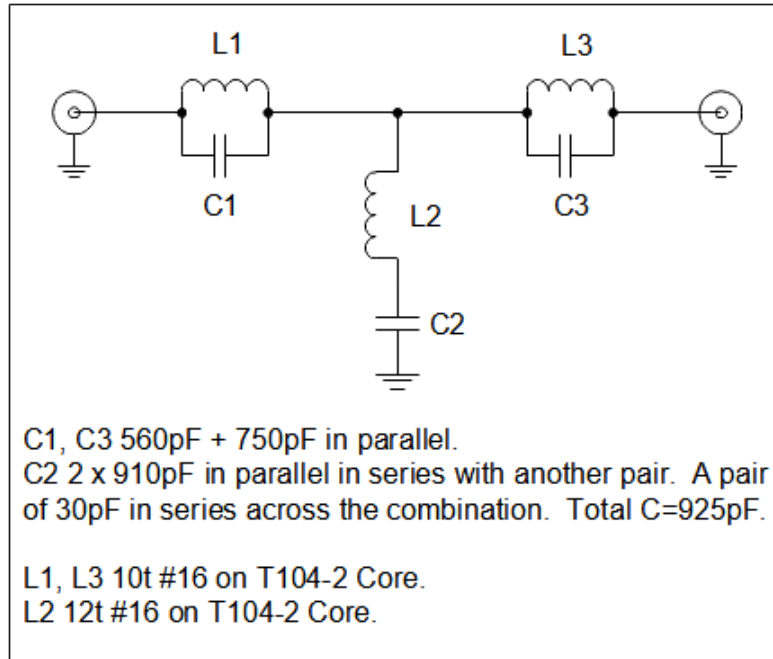


Figure 65. 80 Meter Band Stop Filter designed for the CW end.

The large toroids have high Q and with #16 wire, the losses are very low. At 1.8 MHz, the loss measures 0.07 dB and at 7 MHz it's 0.1 dB. Substantial power can be passed with such low losses. The capacitors are wired in series and in parallel to take more current and voltage. C2 gets the highest voltage at a given power. This design will take 400 watts steady state at low VSWR and higher power in CW or SSB service. The power lost in the filter when running 400 watts on 160 meters would only be a little over 6 watts and 9 watts on 40 meters.

The frequency response is shown in **Figure 66**. Attenuation is over 70 dB from the band edge to about 3670 kHz. This is superior to a pair of coupled stubs and it is much smaller.

Construction is shown in **Figure 67**. The #16 wire is stiff enough to support the toroids. A piece of PC board supported on L brackets is mounted in the center of a 2.25 x 2.25 x 5 inch Bud mini-box. A pad cutter was used to allow the wires to pass through the PC board with out shorting to the ground plane. Additional ground connections on the bottom end of the board improve the isolation through the filter.

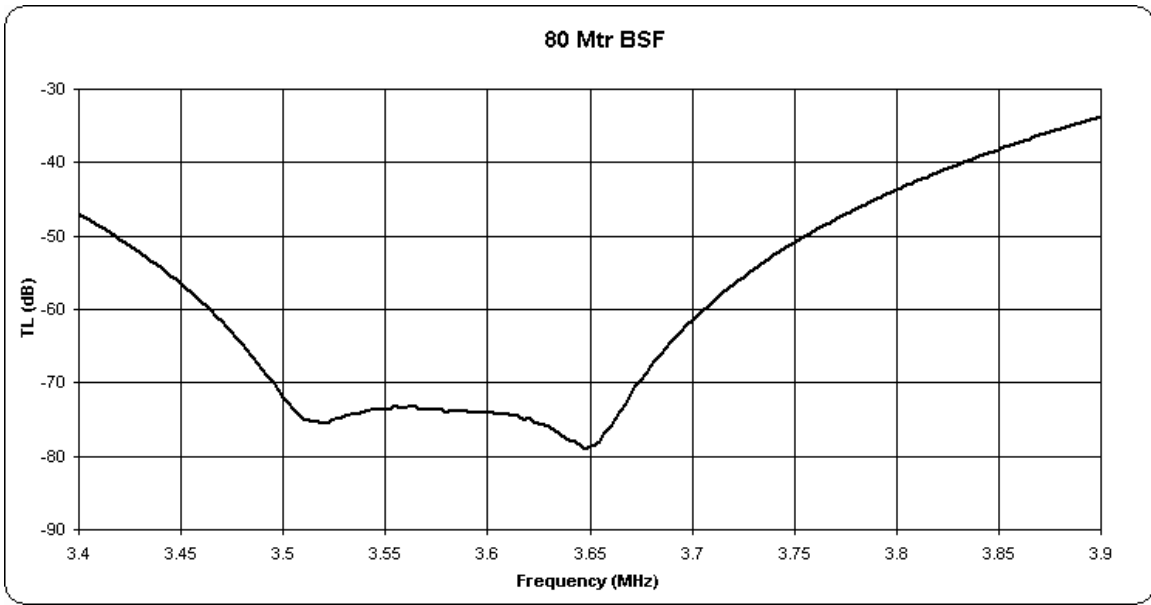


Figure 66. Frequency response of the 80 meter band stop filter.

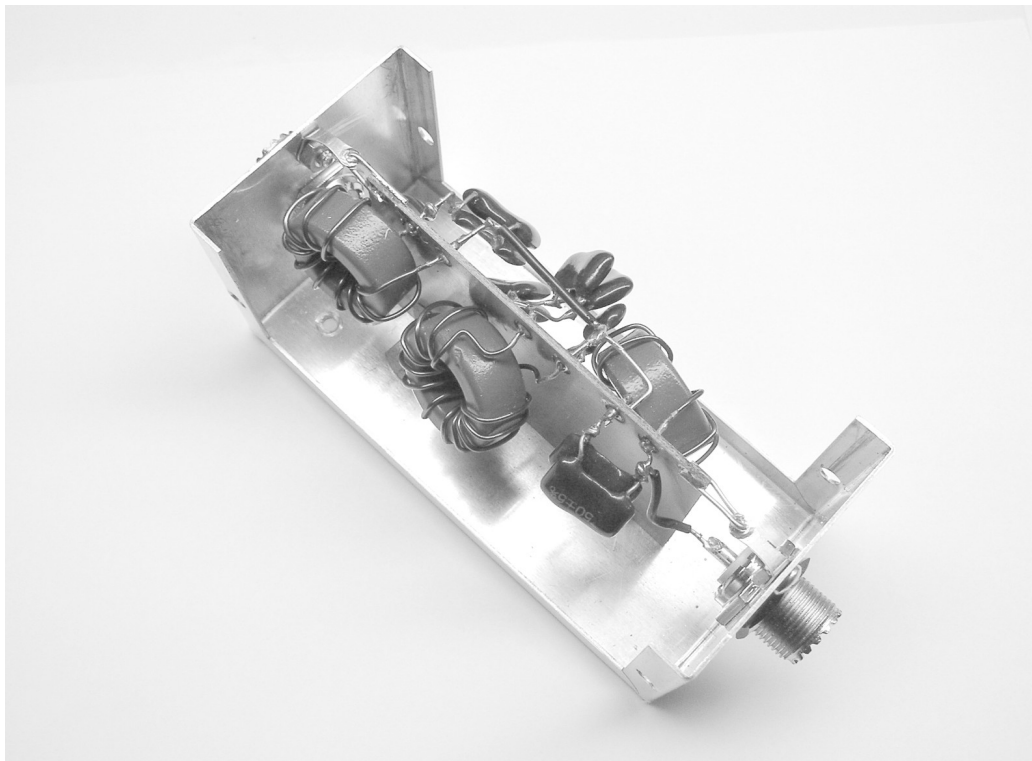


Figure 67. Construction details of the 80 meter band stop filter.

5.0 THE TWO RADIO, ALL BAND STATION

Many operators choose to use two radios while operating all band contests, the only restriction being that just one radio can transmit at a time. Usually, one radio is used for running stations while the second is used for finding multipliers or other stations that have not been previously worked. **Figure 68** shows a typical block diagram for such a station.

Each chain consists of a radio, a band pass filter, an amplifier and a 6 way antenna selector switch. The outputs of each 6-way switch are joined in a set of six A/B switches, or two-way switches. Following each A/B is a set of stubs for the particular band involved. The antenna for each band then is connected via transmission line to the stub set. Usually an automatic band decoder is used to select the bands in use by the BPF and the switches.

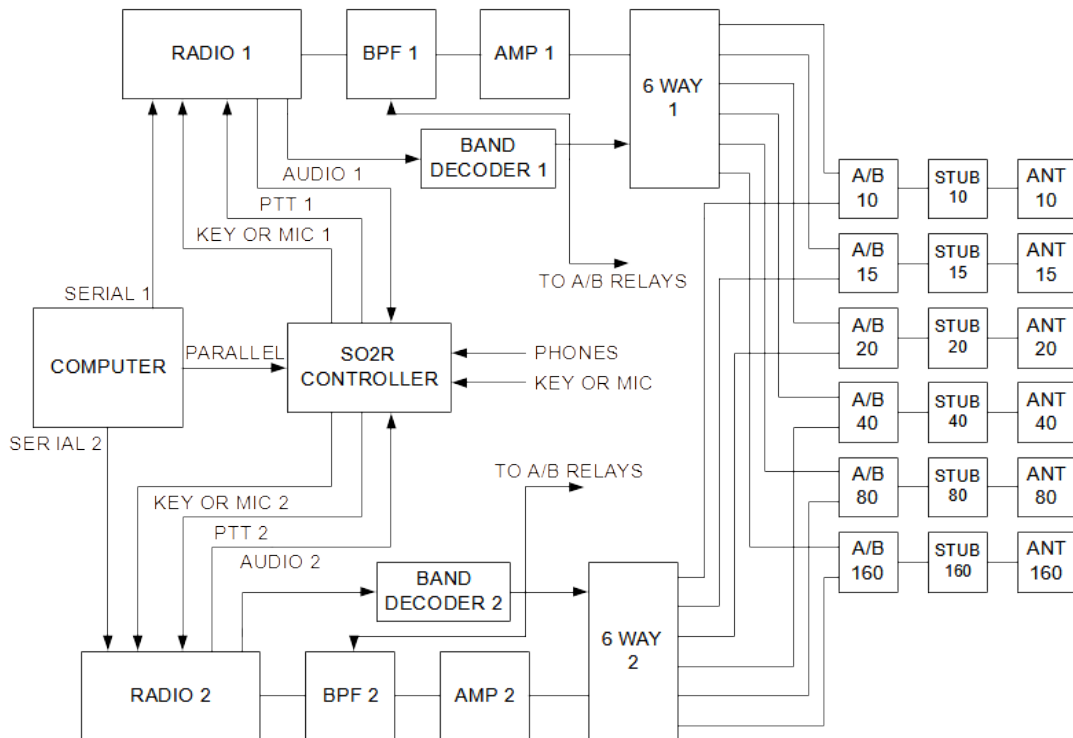


Figure 68 An SO2R equipment setup.

The characteristics of the radio, BPF, stubs and amplifier have been covered in previous sections. The 6 way and A/B switches used in this example are those manufactured by Top Ten Devices, Inc. Typical isolation between ports on the 6

way switches is 40 dB. The isolation of the A/B switches is 85 dB on 10 meters, 95 dB on 20 meters, and even greater on the lower bands.

Note that there are two paths which energy may take to get from one radio to the other. The obvious path is through the 6 way, through the A/B, through the stub assembly, out the antenna then back into the second antenna, etc.

From **Figure 1**, the dBm ladder, we have learned that we need about 147 dB or more isolation to run a second radio with little interference. Let's see what a typical set up like our SO2R station would look like.

Lets assume that Radio 2 is transmitting on 20 meters and Radio 1 is receiving on 10. Table 23 lists the attenuations for the fundamental and harmonic components being transmitted.

The 20 meter radio 2 transmitting energy is 140 dB down and so is the 10 meter second harmonic at the radio 1 receiver.

TABLE 23 Radio 1 isolation from radio 2 through antenna coupling.

	20 meter fundamental	10 meter 2nd harmonic
Power out reference	0 dB	-40 dB
Antenna isolation	-35 dB	-35 dB
BPF	-40 dB	0 dB
Single stub	-30 dB	-30 dB
Double stub	-65 dB	-65 dB
Isolation (1 stub)	-105 dB	-105 dB
Isolation (2 stubs)	-140 dB	-140 dB

The not so obvious path is from the 6 way port which is transmitting to the port which is in use by the second radio's A/B switch and then through the A/B switch isolation. For example: Radio 2 is transmitting on 20 meters and Radio 1 is receiving on 10. Some fundamental and 2nd harmonic energy pass from the 6 way 20 meter port to its 10 meter port. From there it goes to the 10 meter A/B and passes through to the 10 meter receiver of Radio 1. Table 24 lists the attenuations obtained through the switch leakage paths.

Table 24 Radio 1 isolation from radio 2 through switch leakage.

	20 meter fundamental	10 meter 2nd harmonic
Power out reference	0 dB	-40 dB
6 way isolation	-45 dB	-40 dB
A/B isolation	-95 dB	-85 dB
BPF	-40 dB	0 dB
Isolation	-180 dB	-165 dB

Since the antenna-to-antenna path has less isolation, it will be the controlling amount. However, if other types of switches are used which have less isolation this may not be true. If the switches have less isolation some of the benefit of the stub assemblies will be nullified. Note that the combined isolation of the 6 way and A/B switch is a minimum of 125 dB on 10 meters, which far exceeds any other switches useful for SO2R that are available for amateur use.

In order to enjoy the kind of isolation that is possible with these switches, great care must be taken in the station grounding and cabling. An easy performance test is to use shielded dummy loads in place of the antennas and measure the amount of isolation by listening on the non-transmitting radio. The interference level should be lower than when the antennas are in place. It's best to do this with only two bands at a time to eliminate sneak paths through unused antennas.

5.1.1 A 2006 WRTC competitor station design

The station is designed around a pair of FT 1000MP radios. The left side station consists of Radio A and the right side is Radio B. Each side has a band decoder for automatic antenna and band pass filter selection. Each side has a Dunestar band pass filter. The A side is the transmitting station and has the amplifier. Each side has a computer running contest logging software suitable for a 2 station operation. Each computer has a W1WEF keying interface connected to a parallel port. Each interface has a paddle connected. Radio B has a foot switch to initiate control.

A custom designed control box is placed to the left of each radio. These boxes control the keying and receive audio. Either computer may key radio A, but computer A has precedence. A red indicator on each box is on when radio A is transmitting. A green indicator is on when computer B is in control. If radio A starts to transmit, it will take over. Each operator may listen to audio as follows: A in both ears, B in both ears, A in one and B in the other and A plus B mixed or the reverse.

A two way splitter connected to radio A can be used if both radios select Rx mode. If radio A is running a pile up, radio B can also hear the band through radio A's antenna. In this manner, both operators can run the same pile up or radio B can hunt multipliers on the band that radio A is transmitting on. When the

Rx mode is turned off, radio B can listen to either of the two antennas not in use by radio A.

Three A/B relays are controlled by the band decoders to select the proper antenna. The output lines to the antennas have some stubs to minimize harmonic interference to radio B when A is transmitting. An antenna tuner in the 80 meter line allows operation on both ends of the band. It can also be used to tune the 80 meter dipole to a different band for receiving with radio B.

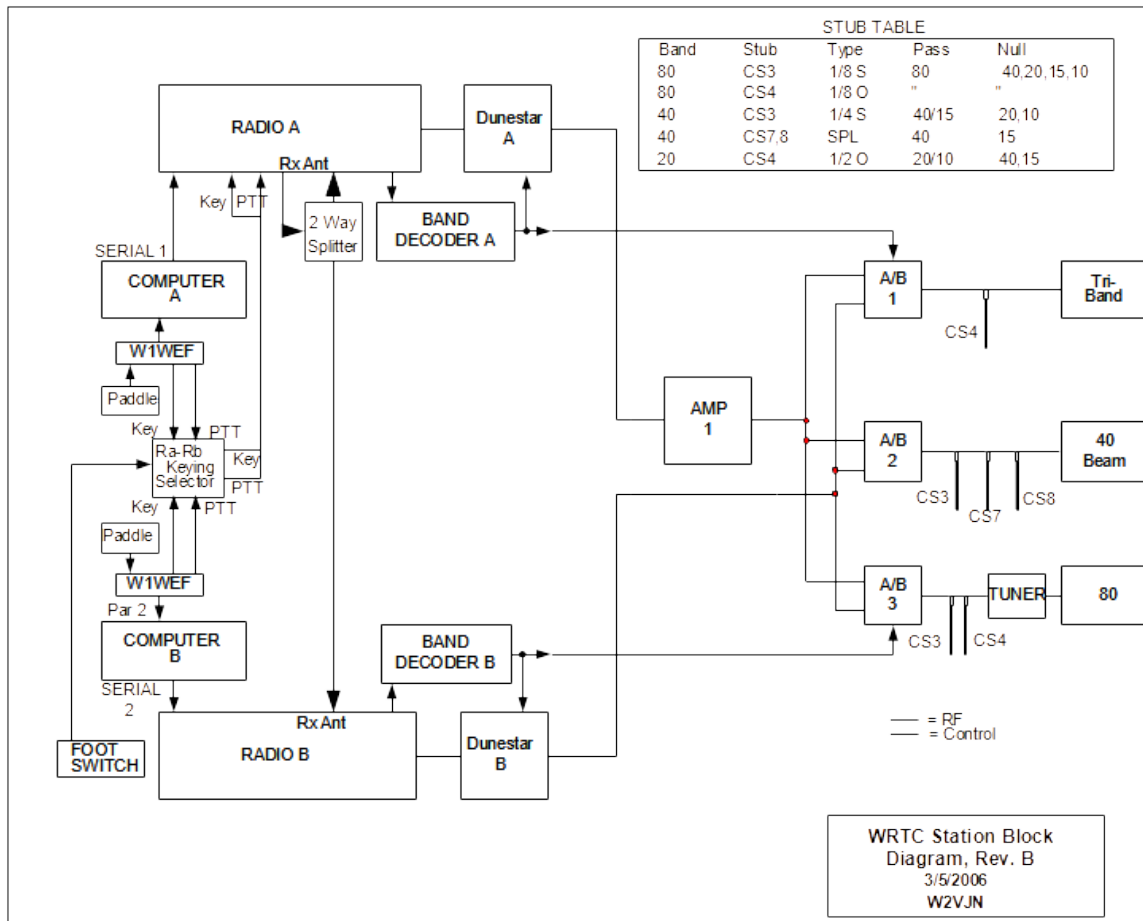


Figure 69 WRTC competitor station design

Each operator had a paddle. The A paddle keyed Radio A, but the B paddle could key Radio A if he pressed a foot switch. In all cases, Radio A took precedence. When A was transmitting, B was locked out. When B was transmitting, A would override B if A pressed his key. All SSB transmitting was to be done by Radio A.



Figure 70 WRTC station test set up.

Two FT 1000MPs and the two control boxes were set up in the US for a functional test before delivery. **Figure 70** shows the control boxes on top of the MPs. In actual operation each was placed to the left of the radio.

The interconnections were all made with RCA phono plug audio cables for reliability.

5.1.2 A multi single station design for WRTC 2006

Operators of the multi-national, multi single station included W6NV, CT3IA, PY3DX, PP5AMP and W2VJN. The operation took place at a station owned by PP5JR. It was located inland some 30 miles from Floianopolos on top of a high plateau in a very isolated spot. Some very big antennas and full KW amplifiers made for lots of fun.

Figure 71 shows the station block diagram. Figure 72 show the station in operation. The call used, ZX5J, is PP5JR's contest call.

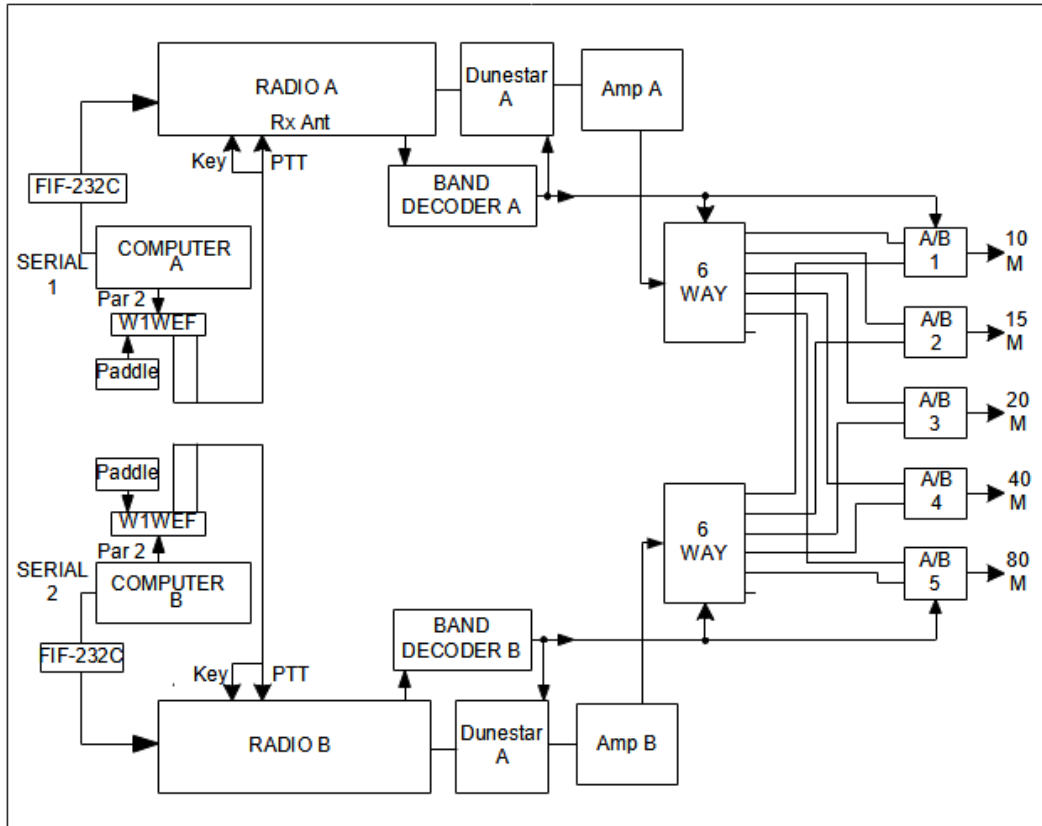


Figure 71 WRTC multi-single station design.



Figure 72 W6NV (foreground) and PY3DX operating ZX5J in the WRTC 2006.

5.3.1 Using the Station Radios to Measure Isolation

It isn't necessary to have a lot of expensive lab test equipment to make some of these isolation measurements. In a multi-transmitter station the required equipment is already there. Before testing with the antennas, it's a good idea to do a test with dummy loads to verify that most of the energy transmitted is staying inside the coax and radio cabinets. Put a 100 watt dummy load on radio 1 antenna connection, and it will act as the transmitter. A second dummy load goes to radio 2 antenna connection, and it can be a low power unit. Radio 2 will be the receiver. If there are filters, stubs or switch boxes in line, they should be included in the test. If any coax is not properly connected or if there is poor shielding on some piece of gear, it will show up in this test.

Before we can work on the isolation problem with actual antennas connected, we must reduce the coupling without the antennas. It's best to test all combinations of bands and make up a table like Table 25.

Table 25 Station to station isolation.

	Rx	160	80	40	20	15	10
Tx							
160		Nil	Nil	Nil	Nil	Nil	Nil
80		-	Nil	Nil	Nil	Trace	Nil
40		-	-	Trace	Nil	S1	Trace
20		-	-	-	Weak	-	Trace
15		-	-	-	-	S2	-
10		-	-	-	-	-	S1

The transmitting band is in column 1 and the receiver frequency is listed across the top row. This test included a pair of FT-1000MPs with 3-foot coax cables to dummy loads. The transmitting radio was set to 100 watts. This table will be the benchmark as other components and antennas are added to the stations.

Two types of important measurements can be made with just a transmitter and a receiver. Let's say that the isolation between the 14 MHz and the 28 MHz station antennas is of interest for SO2R:

Set up the 20 meter radio as the transmitting station and the 10 meter radio as the receiver.

Put the corresponding antennas on these radios.

Reduce the power of the 20 meter station to 1 watt. This should be measured on a Wattmeter for most accurate results.

Set the 10 meter radio to the second harmonic of the 20 meter radio.

Press the key on the 20 meter radio and observe the S meter on the 10 meter receiver. Rotate both antennas so they are facing each other and

the S meter is reading the peak value. This is the worst case direction. If the S meter is reading more than S9, insert some attenuation to reduce the reading to S9 or slightly above.

Now we are going to assume that the S meter reads S9 when the signal level is about 50 uV, or -73 dBm. We can calculate the isolation between the 20 meter station and its second harmonic on 10 meters.

$$\text{Harmonic isolation} = P_{tx} - \text{attn} - (-73 \text{ dBm}) - \text{db above S9}$$

Where P_{tx} = transmitter power in dBm. (1 Watt = +30)
 Attn = attenuator setting in dB

Thus, with 1 watt, the P_{tx} is +30 dBm. I made this measurement with a pair of FT-1000MPs. The S meter read S9 with 18 dB of attenuation in line. Using the above equation:

$$\text{Isolation} = +30 - 18 + 73 - 0 = 85 \text{ dB}$$

With this much isolation it is safe to raise the transmitter power and repeat the measurement. Transmitter harmonics will be increased as power is increased, so a very low power measurement will produce an overly safe isolation measurement. The idea here is to find out if the harmonic will have a significantly detrimental effect.

It is also possible to determine if the fundamental power to be received on the transmitting frequency will cause damage. To do this, the setup is the same; however, radio 2 is set to the transmitter frequency while leaving the original antenna switched on it. Again set the radio 1 to 1 watt output. Tune it in on the radio 2 and add attenuation until the S-meter is not at full scale. As an example, the measurement made here was S9 + 55 dB with 18 dB attenuation in line. Using the above equation, the isolation becomes:

$$\text{Isolation} = +30 - 18 + 73 - 55 = 30 \text{ dB}$$

This tells us that the 10 meter station will be receiving 20 meter energy which is only 30 dB down. If the 20 meter station is running 1500 watts, the 10 meter receiver will absorb $+62 - 30 = 32$ dBm or 1.58 watts. Clearly, this would fry some of the receiver's input components. The two antennas in use for this test are tri-banders, spaced 350 feet. Tri-banders represent the worst case. If monobanders were in use, the 10 meter antenna on radio 2 would not pick up as much energy when set to 20 meters.

Several measurements were made in this manner. Some of the measurements were repeated using lab equipment to see if there was correlation, and the results are shown in Table 26.

Table 26 Testing isolation with the radios

Tx Band	Rx Band	Rx Ant	S Mtr	Attenuator	Isolation	Measured
80, 1W	80	80	S9+25	18	60dB	
80, 100W	40	40	S9	0	123dB	
40, 1W	40	Tri-bander	S9+10	18	75dB	
40, 100W	15	Tri-bander	S9	6	117dB	
20, 1W	20	Tri-bander	S9+55	18	30dB	25 dB
20, 1W	20	40	S9+40	18	45dB	46dB
20, 1W	10	Tri-bander	S9	18	85dB	
15, 1W	15	Tri-bander	S9+50	18	35dB	29dB
15, 1W	15	40	S9+35	18	50dB	44dB
10, 1W	10	Tri-bander	S9+60	18	25dB	30dB

Some of the combinations shown will definitely be safe with no additional isolation above the antenna separation. Eighty and forty meters can transmit full power while the second station is using the tri-bander. When the transmitting station is on 20 and the receiving station is on 10 there will be 85 dB isolation for the second harmonic, but only 25 dB for the fundamental. This would not be safe. A band pass filter or stub that adds something over 25 dB of 20 meter rejection is required to prevent damage.

The lab equipment measurements show that there is an error of +6 to -5dB in the readings obtained with the radios. We learned earlier that isolation of 47 dB is needed to prevent receiver damage when 1500 watts is being used. Since this is a critical value, we will want to increase it to about 57 dB when measuring with radios (and not lab equipment). Thus, the table tells us that some of the combinations shown are capable of creating damage.

6.0 Characteristics of real radios

6.1 Receiver Noise Floor

Typical values for current equipment range from -130 dBm to -145 dBm in a 500 Hz bandwidth. The bandwidth of 500 Hz is most commonly used when receivers are evaluated. We can also express these numbers in terms of noise figure. While this is more commonly used for VHF equipment, it is also a useful indicator. In a 500Hz bandwidth a perfect, noiseless receiver would have a noise floor of -147 dBm and a noise figure of 0dB.

At VHF, the receiver noise figure is normally desired to be as low as possible. Good quality receivers or converters have values that are less than 1 dB. For VHF frequencies external noise levels are very low, so the usual limit on signal detection is the receiver internal noise. The equivalent noise floor is -146 dBm or better.

How low does the receiver noise floor need to be for the HF contesting bands? It is now generally accepted that external noise is the limiting factor for signal detection below 30 MHz. Before there were any academic studies made of noise sources that limit receiving signal levels, radios were made to be as sensitive as possible. Sensitivity was one of the most important characteristics of a good receiver. Radios had at least one low noise RF stage and noise floors were typically in the mid -140 dBm range. With tube designs it was not difficult to obtain a noise figure of 3 or 4 dB in an HF radio. There was not so much concern for dynamic range as there is today. If strong signals were limiting signal reception due to intermodulation, some front end attenuation was applied. This was very effective for DX-ing and contesting on 80 and 40 meters. At higher frequencies signal levels tended to be lower, so the good sensitivity was very effective on 10 meters.

When solid state receivers came along, designers had started paying attention to noise studies made by the ITU* indicating that noise figures of 10 or 15 dB provided adequate sensitivity at 30 MHz. One of the first solid state radios which was accepted by contesters was the TR-7 by Drake. It had a noise figure of about 18 dB. To give it a good dynamic range, the usual RF stage was eliminated. It didn't take long for comparisons to be made between the TR-7 and previous generation radios. At W2VJN a Collins S line was side by side with a new TR-7. The 75S-3 receiver had a 4 dB noise figure, giving it a 14 dB sensitivity advantage. Many A/B tests showed 10 meter signals that were 100% readable on the 75S-3 were not even audible on the TR-7.

Recently, similar tests were run with a popular radio having a -137 dBm noise floor and another having a -144 dBm noise floor. Similar results were obtained. While signals were audible on both radios, some that were readable on the -144 dBm radio were not on the -137 dBm radio. This was at the bottom of the sunspot cycle when 10 meter noise levels were very low and signals were quite weak. Receiver designers must still be using data which does not relate to amateur use. Most available information was obtained using omni directional antennas. Antennas with good directivity can reduce the received noise by 12 to 15 dB.** A competitive station with good antennas and a quiet location should have a receiver noise floor that makes full use of its capabilities. This means a noise floor in the region of -145 dBm in a 500 Hz bandwidth on 10 meters. If the radio of choice has an inadequate noise floor, an external preamp can be added.

* International Telecommunications Union

** Low Band DXing, 1.2.4

6.2 Intermodulation Dynamic Range (IMDDR) and Roofing Filters

“Roofing filter” is the current buzzword in high-end radios. Just what does it mean? Basically, a roofing filter is simply the first IF filter in a radio. It is usually placed as close to the first mixer as possible to be effective. The term “roofing” means that it protects the rest of the radio following it from out of the passband signals.

Modern radios have two basic designs: radios that receive only ham bands that use a first IF in the HF region, typically between 4 and 10 MHz and radios that have their first IF in the VHF region, well above 30 MHz. The latter are usually called "Up Conversion" radios. Let's examine some of the advantages of each.

The Orion, K2, K3 and Omni are the first type. The Yaesu, Kenwood, and Icom radios are the second. Ham band only radios can actually cover most of the HF region with exception to the IF frequency in use. The up conversion radios can cover the entire HF spectrum down to almost zero frequency.

Roofing filters for the ham bands only radios are easy to make and have been available for many years. In the up conversion radios, the first IF is at VHF, somewhere in the 40 to 75 MHz region. All are crystal filters, either discrete or monolithic types. Narrow bandwidth VHF filters have not been available until fairly recently, so the radios with VHF IFs built before about 2005 typically use 10 to 20 kHz wide filters.

The ability of a radio to ignore strong signals near the tuned frequency is greatly enhanced by a narrow roofing filter. Ideally, the final desired selectivity should be in the first IF to protect the following high-gain stages from strong off frequency signals. At the lower IFs it is possible to use filters as narrow as 200 Hz. At VHF it is not yet possible to make practical filters that narrow. Three or four kHz is about as narrow as can be obtained in the VHF region.

Following the antenna connection, most radios have an LC band pass filter. This filter is usually as wide as an amateur band or even wider. So, the first mixer may have tens or hundreds of signals at its input while you are trying to separate out one signal for copy. The ability of the first mixer to handle these signals without excessive intermodulation is a function of its circuit design. It does have a limit, above which there is intermodulation that becomes stronger than the noise floor of the radio. The signal level that produces intermodulation at the level of the noise floor determines the dynamic range of the radio. This characteristic is generally measured with just two signals of the same strength and some particular frequency spacing. When the signal spacing is much greater than the roofing filter bandwidth, the dynamic range of the radio is determined by the first mixer and any other early stages. Dynamic range is measured at several bandwidths such that the signals fall within the roofing filter and also outside of its skirts.

Most high-end radios today have a dynamic range in the area of 95 to 105 dB for signal spacing of 20 kHz or more. As the signal spacing decreases, at some point they fall into the passband of the roofing filter. They then impinge on the following second mixer and IF stages, which will create intermodulation at lower levels. Thus for closely spaced signals, the receiver dynamic range drops dramatically. The transition width from the first mixer dynamic range limit to the second mixer limit is determined by the bandwidth of the roofing filter.

Let's make some guesses as to the signal levels involved here. Assume we have a radio with 100 dB dynamic range and a noise floor of -135 dBm for signals spaced 20 kHz. This means there are two signals – one separated from the receive frequency by 20 kHz and the second by 40 kHz – and they create a false response on the receive frequency when their level is -35 dBm, or 100 dB above the noise floor. How strong is that? Well, S9 in a typical radio is -73 dBm, so these signals are 38 dB above S9. Any signals weaker than that will cause no problem for these or wider signal spacings.

Now, suppose the dynamic range within the roofing filter bandwidth is only 70 dB and the filter is 12 kHz wide (as it is in the FT-1000MP). Two signals spaced at 3 kHz or less will fall inside this filter, and if they are $70 - 135 = -65$ dBm or stronger, they will cause intermodulation signals in the passband. This is only S9 +8 dB per signal. In a radio contest it is possible to have several signals in the +/- 6kHz band around your tuned frequency that are stronger than S9 +8 dB, and this is why we hear false signals under those conditions.

Narrowing the roofing filter has no effect on widely spaced signals, as the intermodulation takes place earlier in the signal chain of the receiver. However, it can improve the receiver performance for close-in signals. In the preceding example, if we reduce the roofing filter bandwidth to 4 kHz, as the Inrad Roofing Filter mod does, the widest separation that causes a problem becomes 1 kHz instead of 3. This can reduce the interference substantially in crowded band conditions. So it's apparent that signals spaced at the roofing filter bandwidth divided by 4 is the minimum spacing at which the dynamic range of the radio is improved.

Shall we go as narrow as possible? Suppose we use a 200 Hz roofing filter. Signal spacings down to 50 Hz is improved. Doesn't this seem a bit close to operate next to your neighbor in a contest? I think the DX station would have some difficulty trying to copy one of you and not the other. What is reasonable? Maybe something that starts attenuating at signal spacings of 100 Hz makes sense. This is a roofing filter with 400 Hz bandwidth. The other advantage of making the filter a bit wider is that the insertion loss is not as great. Insertion loss can reduce the sensitivity of the radio.

Is an 8-pole filter necessary? How does a 4-pole filter compare? One difference between the two filters is insertion loss. For a 500 Hz filter this difference can be a difference of about 5 dB for a 9 MHz filter. The receiver overall gain should be kept fairly constant as filter bandwidths change to preserve the AGC characteristics and to keep the S meter reading constant. Also, the receiver noise floor can suffer if a gain reduction is close to the front end. We need to insert an amplifier or otherwise change the gain to make up for the extra filter loss when a narrow 8-pole filter is selected. This can reduce the dynamic range of the radio.

So 4-pole filters have an advantage, particularly for narrow bandwidths, even though the selectivity is not as good for signals falling down the skirts. There is less advantage in going to a wider filter such as a 2400 Hz bandwidth. For example, Inrad's 10 pole, 2400 Hz filter has an insertion loss of about 2.2 dB,

while the 4-pole filter with the same bandwidth has a loss of 1 dB. The difference of 1.2 dB is small enough that it could be ignored and the 10-pole filter would provide better off-channel rejection. Thus for the SSB bandwidths a good 8 or 10 pole filter will outperform a 4 pole filter, but for the narrow bandwidths the simpler filter is best.

Here are some of the results obtained with the many radios which I have designed roofing filter mods for. These have all been measured in my lab with the same type of setup each time.

TS-850S

TONE SPACING	IMDDR	IMDDR
	OEM	MODIFIED
3 kHz	74 dB	84 dB
5 kHz	75 dB	87 dB
10 kHz	83 dB	97 dB
15 kHz	92 dB	101 dB
20 kHz	95 dB	102 dB

TS 930S

TONE SPACING	IMDDR	IMDDR
	OEM	MODIFIED
5 kHz	80 dB	90 dB
10 kHz	85 dB	93 dB
15 kHz	---	97 dB
20 kHz	90 dB	96 dB

TS 950SDX

TONE SPACING	IMDDR	IMDDR
	OEM	MODIFIED
3 kHz	77 dB	87 dB
5 kHz	77 dB	93 dB
10 kHz	92 dB	99 dB
15 kHz	98 dB	100 dB
20 kHz	98 dB	100 dB

IC 756 PRO

TONE SPACING	IMDDR	IMDDR
	OEM	MODIFIED
1 kHz	69 dB	69 dB
2 kHz	69 dB	82 dB

3 kHz	71 dB	83 dB
5 kHz	79 dB	93 dB
10 kHz	87 dB	96 dB
15 kHz	88 dB	96 dB

IC 756 II

TONE SPACING	IMDDR	IMDDR
	OEM	MODIFIED
2 kHz	78 dB	83 dB
3 kHz	78 dB	86 dB
5 kHz	81 dB	92 dB
10 kHz	85 dB	100 dB
15 kHz	91 dB	102 dB
20 kHz	94 dB	103 dB

IC 756 PRO III

TONE SPACING	IMDDR	IMDDR
	OEM	MODIFIED
2 kHz	79 dB	83 dB
3 kHz	79 dB	88 dB
5 kHz	80 dB	99 dB
10 kHz	91 dB	103 dB
15 kHz	103 dB	105 dB

IC 765

TONE SPACING	IMDDR	IMDDR
	OEM	MODIFIED
5 kHz	78 dB	90 dB
10 kHz	88 dB	93 dB
20 kHz	87 dB	98 dB

IC-781

TONE SPACING	IMDDR	IMDDR
	OEM	MODIFIED
2 kHz	75 dB	79 dB
3 kHz	75 dB	90 dB
5 kHz	76 dB	92 dB
10 kHz	84 dB	94 dB
15 kHz	89 dB	95 dB
20 kHz	93 dB	96 dB

FT-920

TONE SPACING	IMDDR	IMDDR
	OEM	MODIFIED
2 kHz	75 dB	82 dB
5 kHz	76 dB	97 dB
10 kHz	86 dB	102 dB
20 kHz	101 dB	103 dB

FT 1000 MP Field

TONE SPACING	IMDDR	IMDDR
	OEM	MODIFIED
2 kHz	69 dB	84 dB
5 kHz	76 dB	91 dB
10 kHz	-----	99 dB
20 kHz	94 dB	104 dB

OMNI VI+

TONE SPACING	IMDDR	IMDDR
	OEM	MODIFIED
2 kHz	79 dB	94 dB
5 kHz	82 dB	97 dB
10 kHz	86 dB	98 dB
20 kHz	95 dB	98 dB

While results for most of the radios are quite good, the improvement in the Omni VI+ is rather spectacular. Compare the modified Omni VI+ with the Orions listed below. This is because the Omni VI+ is a ham bands only radio and uses a first IF of 9MHz. The roofing filter used in the modification is an Inrad 500 Hz four pole unit. All the other radios listed above are up conversion radios and the filter in the mod kit is in the VHF band. These filters have typical bandwidths of 4 to 5 kHz. This modification has proven to be a major improvement in the reduction of spurious signals during crowded band conditions such as during contesting. These modification kits can be obtained from International Radio in Aptos, CA.

Compare these results to some of the more recent radios which have narrow roofing filters included by the manufacturer. Data has been taken from the ARRL Lab measurements.

ORION

TONE SPACING	IMDDR
2 kHz	88 dB
5 kHz	93 dB
20 kHz	95 dB

--	--

ORION II

TONE SPACING	IMDDR
2 kHz	95 dB
5 kHz	96 dB
20 kHz	92 dB

IC 7800

TONE SPACING	IMDDR
2 kHz	80 dB
5 kHz	89 dB
20 kHz	104 dB

FT 2000

TONE SPACING	IMDDR
2 kHz	63 dB
5 kHz	84 dB
20 kHz	95 dB

FT 9000

TONE SPACING	IMDDR
2 kHz	87 dB
5 kHz	96 dB
20 kHz	98 dB

K3

TONE SPACING	IMDDR
2 kHz	102 dB
5 kHz	102 dB
20 kHz	103 dB

The Elecraft K3 is the only radio with over 100 dB dynamic range as close in as 2 kHz.

6.3 Transmitter Speech Processor Performance

Most radios are equipped with a speech processor. Its purpose is to increase the average power when transmitting SSB. There is a rather simple method to

evaluate the improvement obtained. For a number of years I have been running this test on various radios. Here is how it is done. Please refer to **Figure 73**.

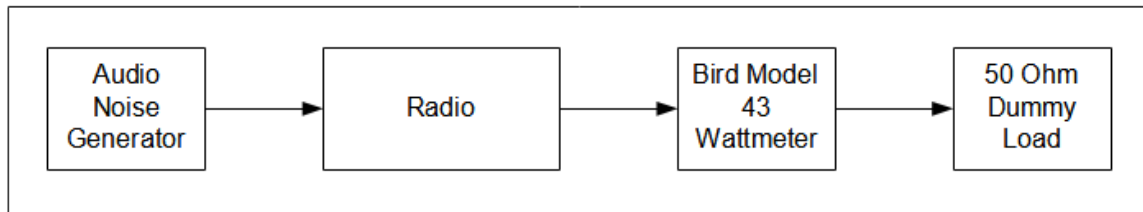


Figure 73 *Speech processor test set up*

An audio noise generator with flat, white noise from 10 Hz to 20 kHz is used as a signal source. It may be argued that white noise is a suitable substitute for speech and it is very easy to measure and to reproduce. There are many such generators that show up on eBay for very low prices. The noise is fed into the microphone jack of the radio being tested. The power output of the radio is monitored on an average power reading wattmeter. A good 50 ohm dummy load is used to absorb the test power.

The radio power output is set to be 100 watts peak so as to standardize the measurements. First the mic audio control is used to set the ALC level of the radio to the level suggested in the operating manual. This is done without the speech processor. The transmit audio bandwidth should be the default or 2400 Hz. If there are any audio tailoring or shaping adjustments, they should be set for a flat response. The power output reading is then recorded.

Next, the speech processor is turned on and the appropriate controls are set to the suggested levels in the manual. If there is no suggested level of performance suggested, I used 10 dB of clipping. The results are shown in table 27.

Table 27. Speech processor performance.

Radio	Po Ave	Po Processor	Ave dB Increase
K3 v1.88	15W	50W	5.2dB
TS-930S	16W	47W	4.7dB
TS-850	25W	65W	4.2dB
TS-940S	16W	35W	3.4dB
IC-781	15W	33W	3.4dB
FT 1000 MP	24W	49W	3.1dB
IC-751A	19W	38W	3.0dB
IC-756 PROII	17W	34W	3.0dB
FT-1000	20W	38W	2.8dB
IC-756 PRO	13W	22W	2.3dB
IC-761	20W	31W	1.9dB
JST-135	35W	50W	1.6dB
Omni VI+	13.5W	19W	1.5dB

IC-756	17W	23W	1.3dB
IC-756 PROIII	17W	23W	1.3 dB
FT-2000	16.5W	23W	1.0dB
IC-7800	15W	18W	0.8dB
FT-920	30W	36W	0.8dB

Note that the average power for a peak power of 100 watts is all over the lot. This indicates that the radios with more than 15+/- watts already have some speech processing before the actual processor is engaged. This is probably due to differences in the ALC rise and fall time constants. These differences in power are not significant as the operator using an amplifier will have to set power for proper drive. The dB increase is very important, however. Using the K3 as an example, an increase of 5.2 dB represents a radiated average power of 3.33 times. This kind of increase in average power is difficult and expensive to obtain with antenna improvements or increased amplifier power output. So the speech processor dB improvement is quite important. It is not something that has been measured or rated in any publication before. If you are a contester or a serious DXer, it's something to be considered when buying a radio.

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 - 1.2 The RF Connection. www.therfc.com
 - 1.3 Lands Precision, Inc. (Connectors) www.landsprecision.com
2. Cable cutter:
 - 2.1 Techni-Tool. www.techni-tool.com
3. Coax stripper:
 - 3.1 Cable X-perts, Inc. www.cablexperts.com
 - 3.2 National Communications, Inc. www.trynci.com
4. Cable meter:
 - 4.1 Hykon Mfg Co. www.hykon.com
5. Instruments:
 - 5.1 DDS Sweeper kit. www.netSPACE.net.au/~tbece
 - 5.2 Reflection test instrument, AIM4170. www.w5big.com
 - 5.3 N2PK Network analyzer www.N2PK.com
6. Shrink fit and other parts:
 - 6.1 Mouser Electronics. www.mouser.com
 - 6.2 Digi-Key. www.digikey.com
 - 6.3 Magnet wire, toroids, etc. www.kitsandparts.com
 - 6.4 Magnet wire. www.bulkwire.com

6.5 Amidon, ferrite and powdered iron cores. www.amidon.com

7. Additional information on the web:

7.1 Information on VHF stubs.

www.ifwtech.co.uk/g3sek/swxfiltr/swxfiltr.htm

7.2 Stub theory. www.k1ttt.net/technote/stubpair.html

7.3 Commercially available stubs. www.qth.com/topten/stubs.htm