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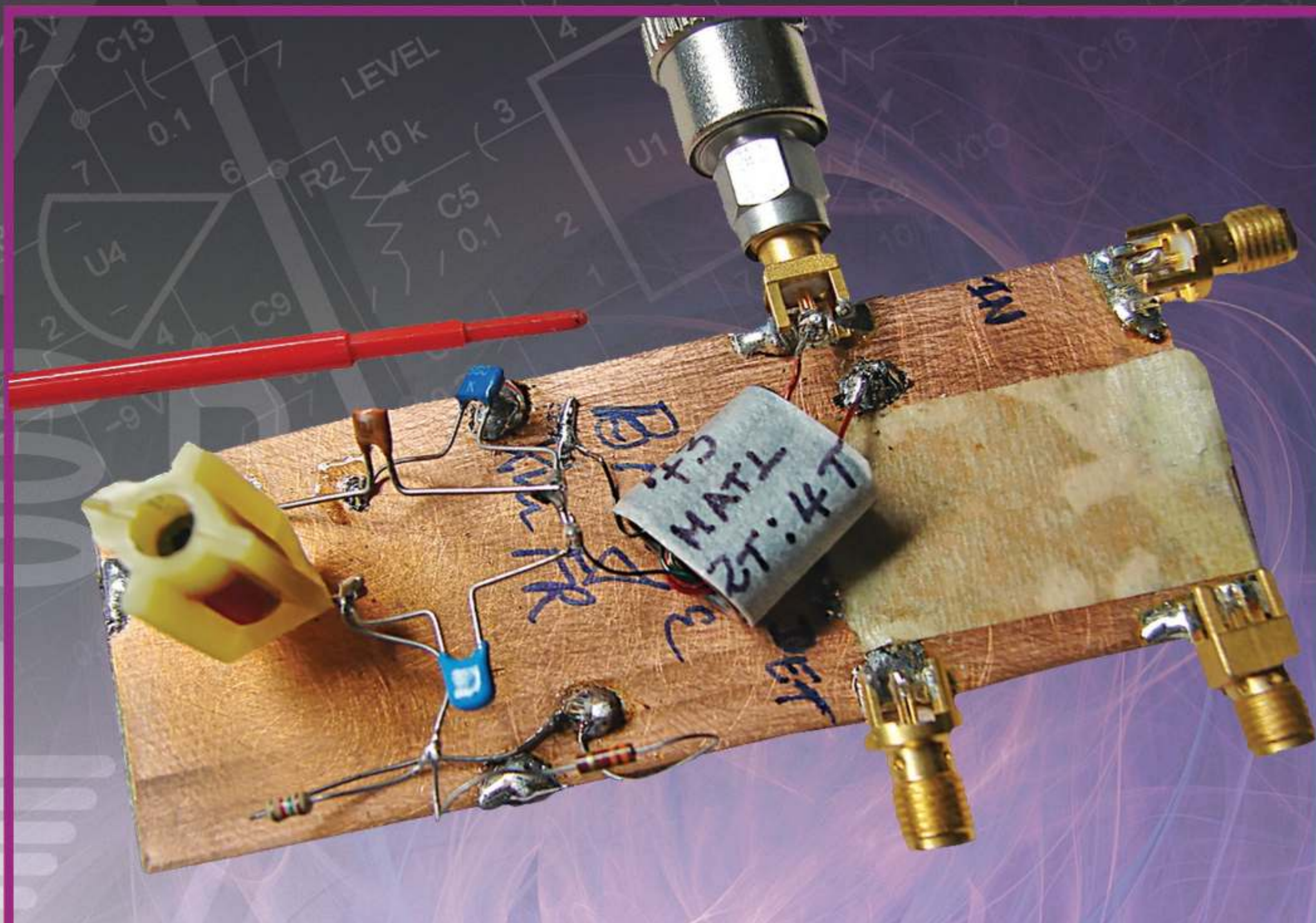

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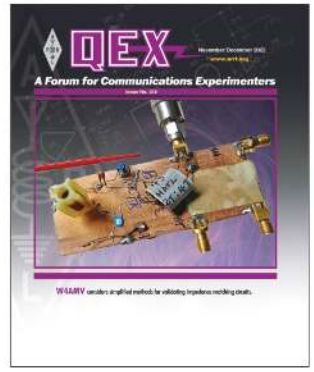
About the Cover

Alan Victor, W4AMV, reviews impedance transforming or matching circuits, and provides various Q definitions used in designing and evaluating these circuits. Distinguishing the operating Q and the design Q reduces the confusion between design and measurement results. W4AMV provides a method of finding the operating parameters of an impedance matching or impedance transformation circuit. A simple single-port measurement is completed using an antenna analyzer, a directional coupler or SWR bridge, an RF signal generator with scope, a spectrum analyzer, a noise bridge, or a VNA. The single one port measurement returns the quality of the match and the operating bandwidth. Alternative approaches are possible if the terminations are not standard values.



In This Issue

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2) document advanced technical work in the Amateur Radio field, and

3) support efforts to advance the state of the Amateur Radio art.

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Kazimierz "Kai" Siwiak, KE4PT

Perspectives

RF Radiation Safety Awareness

We are taking this opportunity to raise awareness about RF safety and exposure to RF radiation. For US amateurs all formal license related interactions with the FCC involve FCC Form 605. This includes applying for an amateur license, changing the licensee's address, renewing an amateur license, upgrading a license, and so on. In the "General Certification Statements" section we find that by signing FCC Form 605, the "Amateur Applicant/Licensee certifies that they have READ and WILL COMPLY WITH Section 97.13(c) of the Commission's Rules (available at web site **http://wireless.fcc.gov/rules.html**) regarding RADIOFREQUENCY (RF) RADIATION SAFETY and the amateur service section of OST/OET Bulletin Number 65 (available at web site **http://www.fcc.gov/oet/info/documents/bulletins/**)." This Certification Statement is not new – it has been on Form 605 for a very long time.

What is new is that OET Bulletin Number 65, along with many ARRL publications, are currently under review to provide updated guidance regarding the rule changes of FCC 19-126 that became effective May 3, 2021. Going forward you will find gentle reminders in ARRL publications, such as:

• Stay safe around antennas and radiating systems,

and, where appropriate, editor-inserted comments in articles like,

• [Be sure to perform an RF exposure assessment with this antenna. — *Ed.*].

Stay aware of RF radiation safety.

In This Issue:

• Kim Swedberg describes the versatility of a lock-in amplifier design.

• Bob Larkin, W7PUA, combines dipoles to produces antennas of very wide bandwidth.

• Alan Victor, W4AMV, simplifies validation of impedance transforming and matching circuits.

• Eric Nichols, KL7AJ, in his Essay Series, tackles quality factor Q.

Writing for *QEX*

Please continue to send in full-length *QEX* articles, or share a **Technical Note** of several hundred words in length plus a figure or two. *QEX* is edited by Kazimierz "Kai" Siwiak, KE4PT, (**ksiwiak@arrl.org**) and is published bimonthly. *QEX* is a forum for the free exchange of ideas among communications experimenters. All members can access digital editions of all four ARRL magazines: *QST*, *OTA*, *QEX*, and *NCJ* as a member benefit. The *QEX printed edition* is available at an annual subscription rate (6 issues per year) for members and non-members, see **www.arrl.org/qex**.

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Very kindest regards,
Kazimierz "Kai" Siwiak, KE4PT
QEX Editor

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Simplified Measurement Methods for Validating Impedance Transforming and Matching Circuits

The emphasis here is measurement simplicity and the desire to obtain validation that our design is functioning as intended.

Key to successful digital and analog circuit performance is the ability to interface diverse circuit functions. This is observed in digital circuits where pulse fidelity must be maintained otherwise signal distortion can readily hamper decision making elements like high speed A-to-D converters. In high speed analog and RF circuits, all functions critically hang on the proper impedance transfer between circuits of varying functions; mixers, amplifiers, oscillators to name a few. These circuit performance attributes such as noise, gain, efficiency, distortion and power are significantly affected by the simple action of transferring one impedance level to another. Some circuits are designed to provide impedance matching between the terminations and a conjugate match is required. While other circuits are used to modify one of the impedances to a different value and not necessarily provide a match. Instead, they provide a specific termination.

The design of matching circuits is addressed in a large volume of literature. A search within the IEEE framework returns over 24,000 papers from the mid 1900s through 2022. Many impedance transforming circuits are narrow band and their terminations are either real or purely resistive or can include a reactance element. In that case the termination is complex. Narrow bandwidth (BW) impedance transforming circuits have closed form solutions that provide for the values of their components in terms of their terminations. Additional requirements for their design are the operating frequency, the required BW or frequency span, designated operating Q and the desired performance of the transform. The performance of the circuit is a measurable value, which includes the transmission loss from the source to the load and a return loss value, which measures the quality of the match between the two terminations. These two measured values are actually intertwined and a single measurement is all that is needed for their characterization. After outlining the steps for impedance transforming or matching, this measurement is discussed with application examples.

The configurations that are popular in narrow band designs include the **L**, π, **T** (Ell, Pi, Tee), capacitor and inductor tap as well as link coupled and auto transformers. A minimum of two circuit elements are required to provide a minimum loss match and three circuit elements extend the possibilities significantly. As many as 28 matching networks are possible **[1]**. These lumped element designs are complemented by their distributive topologies, which use transmission lines. These circuit elements are arranged as cascaded lines with varying impedance and electrical length, or as lines with open or shorted stubs.

Extending the impedance transforming BW is possible by applying a cascade of wider BW or lower Q designs or graphically by applications of the Smith Chart and by means of network synthesis **[2]**. Choosing the appropriate circuit topology and element values is essential as well to readily check and validate impedance transforming circuit operation. The validation techniques discussed here will consider narrow band designs. However, it should be possible to extend the concept to ones with wider BW. Source and load terminations may be real or complex.

The Various Forms of Q

When the concept of Q is applied to impedance transforming and matching circuits, several Q definitions and naming conventions are exchanged and they are not all identical. The desired objective in this paper is to distinguish between the operating Q, Q_o and the design Q. Their values are possibly different and dependent on how the final circuit is applied. References will be made to the singly loaded and the doubly loaded circuit and will show how this idea impacts Q.

The basic definition of Q is the ratio of energy stored per unit cycle compared to the energy dissipated. It is the ratio of the reactance to the resistance and in a series circuit, the series Q or Q_s in **equation (1)** is also referred to as the unloaded Q, Q_{UL} of the reactive component. While in a parallel circuit, **equation (2)** it is still the unloaded Q but it is referenced as the parallel Q, Q_p

$$Q_{UL} = \frac{X_s}{R_s} = Q_s \tag{1}$$

$$Q_{UL} = \frac{R_p}{X_p} = Q_p \tag{2}$$

It is key to note that if a tuned circuit is formed from either the series or parallel reactance defined in **(1)** or **(2)**, at a single resonant frequency, f_o, the series operating and the parallel operating Q, Q_o, are identical. Both circuits possess the same BW at resonance. The BW is the half-power point as measured through the tuned circuit formed and the BW is defined as

$$BW_{3dB} = \frac{f_o}{Q_o}. \tag{3}$$

This idea of equivalent BW is fundamental in impedance transforming circuits as a series circuit composed of resistive and reactive, R_s and X_s components, has an impedance Z_s, which can be converted into an equivalent parallel admittance Y_s with a parallel set of components G_p, B_p **[3]**. This transform technique is referenced as the (Q^2+1) rule as the real parallel resistance R_p or conductance ($1/G_p$) is obtained from

$$R_p = R_s\left(Q_s^2 + 1\right) \tag{4}$$

and the series resistance value from its parallel form

$$R_s = R_p \big/ \left(Q_p^2 + 1\right). \tag{5}$$

Here Q_s and Q_p, the series and parallel unloaded Q, are identified as Q_{UL} from **equations (1)** and **(2)**. The pair of **equations (4)** and **(5)** suggests this impedance transform property. The action of adding a reactance in series with a resistance provides the ability to modify the resistance value, R_s in a lossless manner. Its value is increased to R_p at a specified resonant frequency. Adding a reactance in shunt with a parallel resistance, R_p, will transform its resistance value down to R_s. The pair of **equations (4)** and **(5)** show the Q values for these series and parallel circuits. They are proportional to the ratio of the resistances. These Q values are also referred to as the transformation Q. This ability to transform a series circuit to a parallel equivalent and vice-versa is quite useful. It may be applied to more complex transforming circuits to assist in seeing their impedance transforming operation by successive application of **equations (4)** and **(5)**.

Consider the application of a two element Ell section providing a 1 kΩ termination from a 50 Ω resistive load. The impedance transfer is desired at a center frequency of 7 MHz. **Equation (4)** is applied directly and the required series Q calculated is 4.36. The increase in the resistive value from 50 Ω to 1 kΩ is accomplished by adding a series capacitive reactance obtained from **equation (4)** and finding its value from **equation (1)**. The value of Q_s is substituted for Q_{UL}, The assumption here is the "intrinsic" unloaded Q of the series capacitor added is significantly larger than the desired series Q of 4.36. In impedance matching and transforming circuits, it is desired to have the Q_{UL} much larger than the loaded Q or in this case the series Q so as to minimize losses, see **Appendix I**. Using **equation (1)**, the series reactance and subsequent capacitor value at 7 MHz is $-j$217.94 Ω or 104.4 pF. Next a parallel tuned circuit is required. Converting this capacitor to its parallel equivalent value is obtained from **equation (2)** since $Q_s = Q_p$ at a single frequency. The equivalent shunt capacitive reactance is $-j$229.4 Ω or 99.2 pF and the required shunt inductance to provide resonance at 7 MHz is $+j$229.4 Ω or 5.21 μH. This completes the impedance transform. **Figure 1** highlights the individual 3 step process. The desired circuit outcome is the 50 Ω source is increased in value to 1 kΩ. Looking into the input side of this circuit at node 3 in **Figure 2A**, will provide a 1 kΩ value at 7 MHz. The circuit is said to be singly loaded. The impedance transform is completed and the 50 Ω source now appears to be 1 kΩ. Either Q calculated, Q_p or Q_s results in a value of 4.36 as they are identical at a single frequency. If the Ell circuit is to be analyzed and operation verified or if it is to be used as a matching system, then a 1 kΩ termination would be added to the input side at node 3. The circuit is said to be doubly loaded and the operating Q, Q_o is $Q_s/2$. The purpose in highlighting this simple impedance transform exercise is to identify the various outcomes for Q. Since the circuit Q values can vary, so will its measured BW. In **Appendix II**, a similar process will be exploited to highlight the Pi network.

The Smith chart shown in **Figure 2B** is used to solve the same set of steps just discussed but graphically, and it provides visually the various response Q's just mentioned for a singly loaded circuit. The operating Q_o the transformation Q or the Q_s of the circuit are provided on the chart as Q contours **[4]**. Each sweeping series or parallel resistance and reactance or *RX* contours on the chart represents circuit branch or node Q values. The Ell network is unique in that the Q is constrained and fixed by the load to source impedance ratio. If the Ell network is terminated by its matched value, 1 kΩ in this case, the Ell network loaded Q, Q_L is half the Q value displayed on the chart. If the loaded Q is small, peaking at

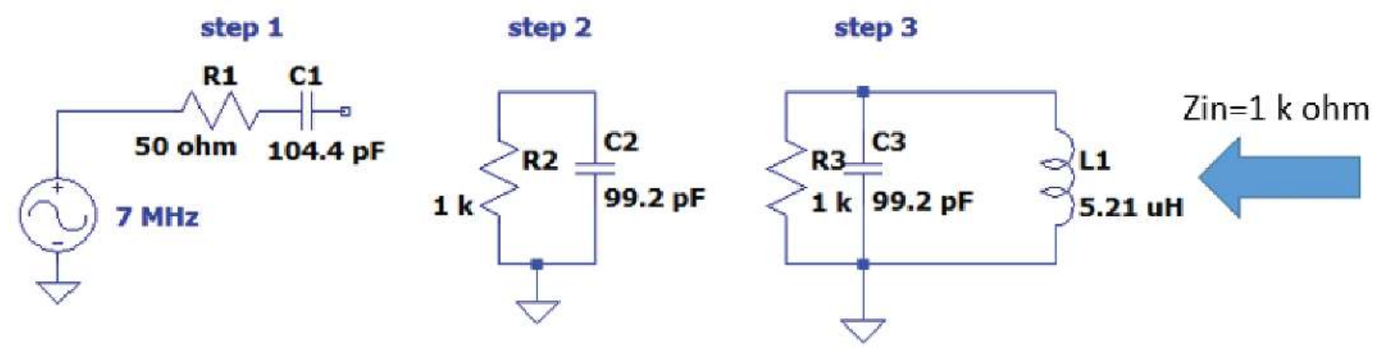
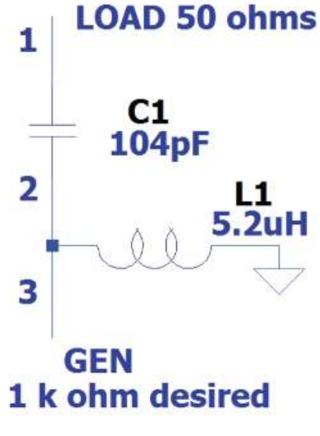
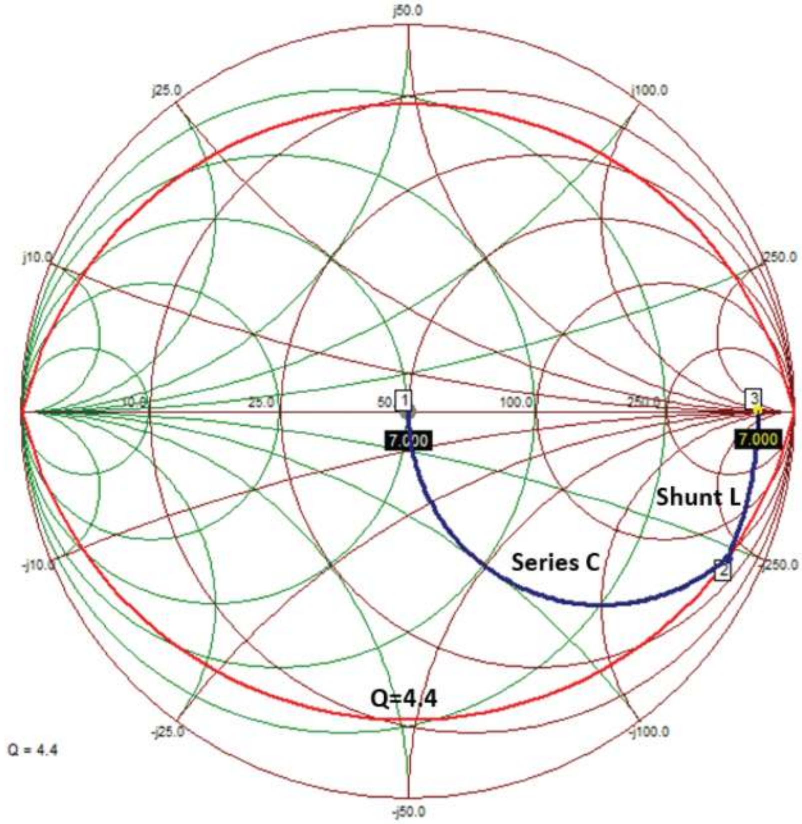


Figure 1 — Three steps are involved to move 50 Ω to 1 kΩ with the Ell circuit.



(A)



(B)

Figure 2 — (A) The Ell match transforms 50 Ω to 1 kΩ and the Q constant contour of 4.4 provides a singly loaded Q readout for this circuit. (B) The 1 kΩ is the required termination value. If the circuit were doubly loaded as in a matching configuration, the resulting operating Q would be half the value of the Q contour.

the resonant frequency is difficult to discern. However, the measured Q value will be reasonably close to the calculated value.

The construction and application of constant Q contours on the chart is applied in the next section. The Q contours are connection points of equal values of *R* and *X*. The chart is an excellent utility for implementing the impedance transforming process. It is quite handy for more complex impedance transforming circuits. The chart shown in **Figure 2B** highlights the Q contour that is calculated based on the *R* transformation required, 1 kΩ from 50 Ω. Namely from **equation (4)**, the Q value is approximately 4.4.

The simple one section Ell impedance transformer plots Q directly on the chart. However, as additional *LC* sections are added, the effective Q becomes more complex. In this case, each individual resistive-reactive combination forms an individual Q, referred to as the branch (series) or nodal (parallel) Q, Q_n. Each Q_n value contributes to the total Q in proportion to their individual values. The total Q is tedious to find for a multiple element circuit. However, key information is available for the total circuit BW by constraining the impedance trajectory movement between various constant Q_n contour boundaries. If the trajectories as shown in **Figure 2B** are constrained to smaller Q contour values, the operating BW is increased as each Q_n value decreases, as well as their total contribution.

In **Figure 1** three steps are involved to move 50 Ω to 1 kΩ with the Ell circuit. First, add a series *C* to 50 Ω, step 1. This moves *Z* to an equivalent value of 1 kΩ with a reactance present. Step 2 is the parallel equivalent circuit of step 1. Step 3 provides the resonant tuned circuit of step 2 by adding a shunt inductance. This step provides a real 1 kΩ value. The end result is the circuit in step 1 with a shunt inductor added as shown in the final circuit of **Figure 2A**.

Method of Testing

Characterization of the matching network or the impedance transforming circuit involves looking at the insertion loss, the quality of the match such as the SWR or the reflection coefficient and the operating BW. Two-port testing will accomplish this task, however there is a simpler technique that requires only a single one-port measurement. Since the impedance transforming system is a passive circuit and is reducible to a single tuned circuit at a specific frequency, the transmission and reflection coefficients are dependent on each other. The conservation of energy stipulates that the total power incident on a passive circuit equals the power absorbed, plus any power flowing out from the circuit. When the network is passive, the dissipated power is greater than or equal to zero. Thus for a passive two port the following relationship holds **[5]**, **[6]**:

$$|S_{11}|^2 + |S_{21}|^2 \leq 1 \qquad (6)$$

Equation (6) permits finding the transmission loss, S_{21} from a single one-port measurement, S_{11}.

If the passive matching circuit is lossless and properly terminated, that is designed to work in the Z_o of the system, then S_{21} must be unity and hence S_{11} must be zero, that is, a perfect match. Correspondingly, if the loss is excessive, S_{21} will tend to zero, the reflection is high and S_{11} will converge on unity. The relationship between the measured center frequency return loss in dB versus the corresponding frequency where the 3 dB down power point return loss occurs, is shown in the graph of **Figure 3**. If the match at center frequency is excellent, for example 20 dB or better return loss, then

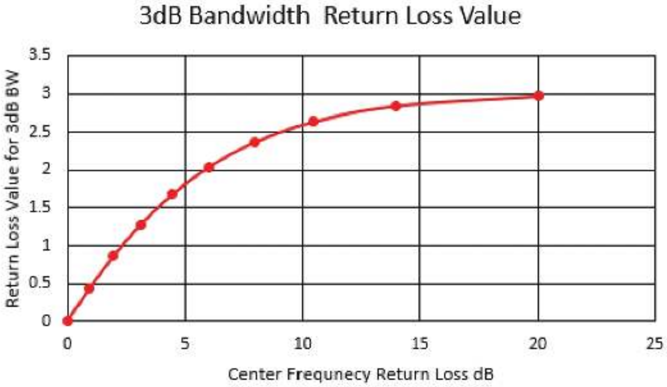


Figure 3 — Plot of the center frequency response return loss in dB versus the corresponding half power, 3 dB down points of the matched circuit. This data is readily available from a single one port return loss measurement of the circuit.

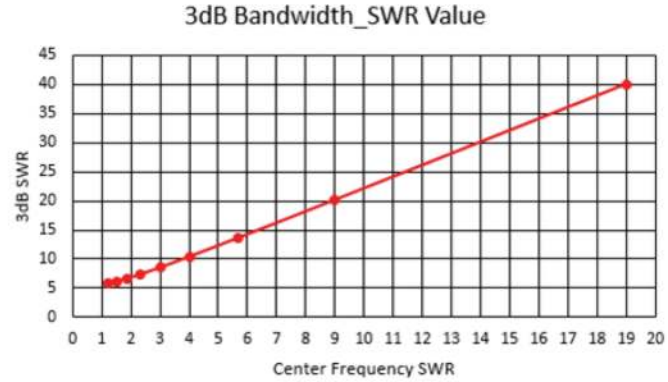
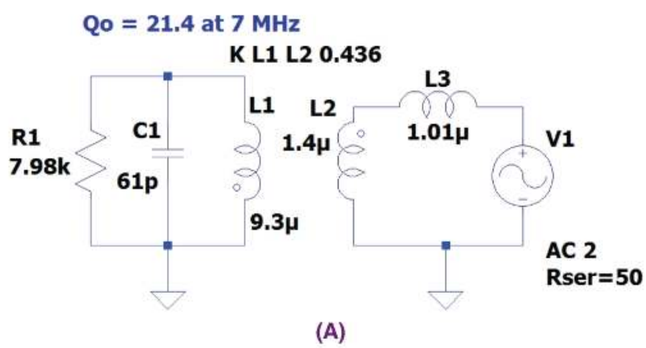
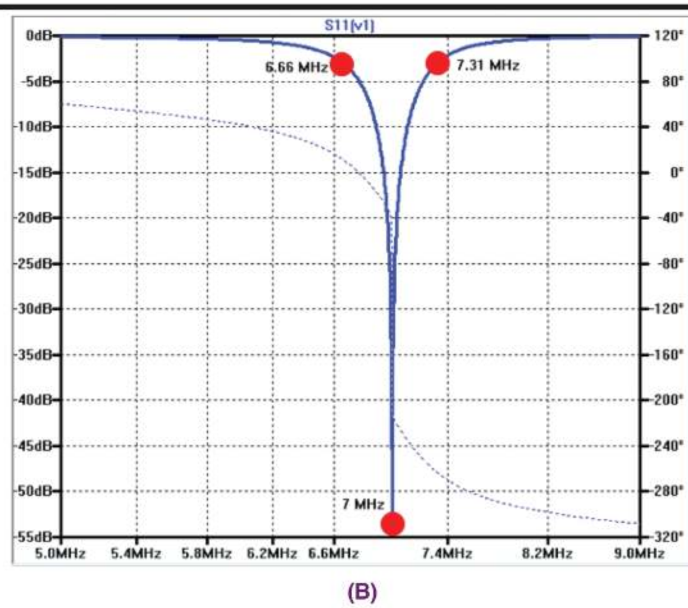


Figure 4 — Plot of the center frequency response SWR versus the corresponding half power or 3 dB down points of the matched circuit, SWR. This data is readily available from a single one port SWR measurement of the circuit. For a 1:1 center frequency SWR, the 3 dB points are at 6:1 SWR.





(B)

Figure 5 — (A) A link coupled circuit serves as an impedance transformer and provides a band-pass response. (B) — Evaluation of the match quality and operating BW are provided by a single one port measurement of the link coupled circuit.

the frequency where the 3 dB down power point return loss occurs are the points where the return loss intersects 3 dB. There are two points, one above the center frequency and one below. Then the operating Q is obtained from **equation (3)**. As the center frequency return loss decreases, so do the points for the 3 dB down frequency response locations. For example, if the center frequency return loss is only 6 dB, the corresponding 3 dB down return loss frequency points are 2 dB. In terms of SWR, the limiting value for a perfect center frequency match is an SWR of 1:1. This correlates to the 3 dB down frequency response points at a set of SWR frequencies intersecting the 6:1 SWR value, see **Figure 4**.

It is key in the measurement to check and calibrate the instrument used prior to assessing the return loss or the SWR values. Devices such as the directional coupler, SWR bridge, and vector network analyzer (VNA) require a known quality standard for initial measurement. Knowing the limits of your measurement device is key. What are the lowest and highest SWR or maximum and minimum return losses achievable in the measurement? As an example, a directional coupler could very well discern a 40 dB return loss while not able to achieve a return loss of zero dB with a perfect reflection termination attached. Coupler losses will limit how accurately a poor return loss will be displayed. Although simplified, a correction technique attaches a perfect reflection, such as a short circuited plate and normalizes –3 dB BW measurements to that value. A variety of measurement instruments are briefly outlined and discussed in **Appendix III** and shown in **Figure A3**.

Examples

A link coupled impedance transforming circuit is a good example of this measurement application **[7]**, see **Figure 5A**. As a variation on the usual link coupled configuration, consider a case where the smaller link coupled coil is placed in series with a second series inductor of 1 μH and the load. This series inductor is not coupled to the link coil or to the primary main coil. See **Figure 5A** schematic. The design provides an impedance transformation from 50 Ω to nearly 8 kΩ at 7 MHz and the primary and secondary *L* are 9.3 μH and 1.4 μH. The coefficient of coupling is 0.44 and not difficult to obtain if the secondary link coil is physically overlapping the primary. The addition of a series inductor onto the link provides the increased transformation,

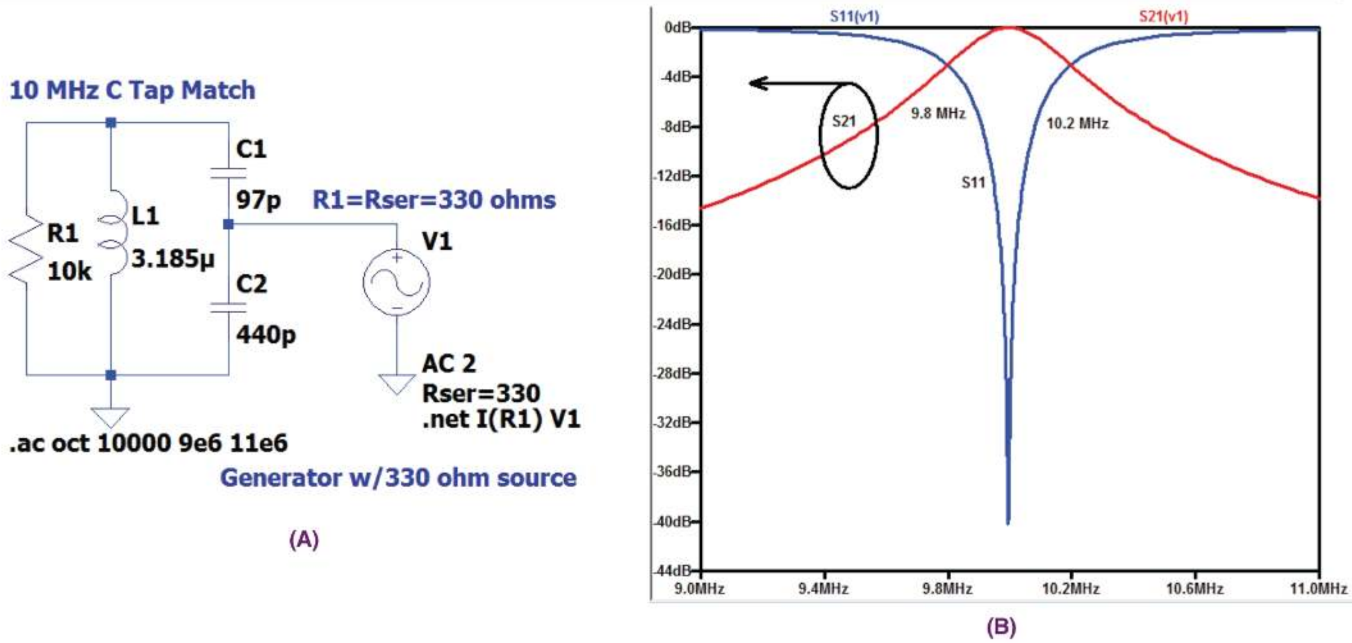


Figure 6 — (A) The capacitive tap matching network frequency response is doubly loaded. One load is provided by the shunt 10 kΩ and the other by the transformed 330 Ω to match. (B) The circuit BW response is 400 kHz and at a 10 MHz center frequency a doubly loaded Q of 25. Hence, in good agreement with the design value for Q_o of 50.

however, the operating BW of the circuit is narrow and the Q_o is 21.4. Since the center frequency match is excellent, searching for the 3 dB BW points is simply a process of finding the set of 3 dB return loss points. These are highlighted at 6.66 and 7.31 MHz in **Figure 5B**, a frequency span of 650 kHz. Application of **equation (3)** returns a Q_o of 10.77, which needs to be doubled as the calculated Q_o is based on a singly loaded condition. Hence, the measurement returns a Q_o of 21.54 and is in good agreement with the calculated value of 21.4.

What if the Impedance Set is Not 50 Ω?

A single one port measurement is desirable for the characterization of impedance matching circuits. Again, the emphasis here is measurement simplicity and the desire to obtain validation that our design is functioning as intended. However, many cases require a source and load that are not 50 Ω or the matching system does not provide a Z_o equal to that of the measuring system. The use of minimum loss pads or modification of the source or load from 50 Ω to the desired impedance will reduce the dynamic range of the return loss measurement. The introduction of measurement error is possible without properly accounting for the impedance modifications. An ideal transformer is a great solution to this dilemma and can offer one solution. Another possibility is introducing a second matching circuit to provide the desired interface. Yet another is the use of an oscilloscope and a signal source. The oscilloscope application can verify the singly loaded Q value by a performing a response measurement of the –3 dB voltage points or where the voltage across the match circuit is reduced by 0.7 times the peak response at the design center frequency. The impedance transform circuit is only loaded by one of its designated terminations, namely at the source. This is not a matched system; it is a termination circuit. However, the design Q for this termination circuit is verifiable. When a proper termination is added to the load side, the prior measured BW should be doubled.

A capacitive tap matching network is a good example of an inter-stage impedance transform circuit, which acts like a transformer. The circuit can increase the value of an impedance in proportion to the ratio of a series pair of capacitive reactance components. The ratio between the resistor values is designated as N. The capacitive tap is able to handle the interface between widely different terminations. In this example, the desired Q loaded or Q_o is selected to be large, 50. The next section will show that this is not an easy Q value to obtain, and the consequences. The source and load impedances for this example are R_1 = 330 Ω and R_2 = 10 k Ω respectively. The solution to the matching system is obtained by successive application of the parallel to series equations and solving for the component values of the match in terms of known values **[3]**. Validation of the design method is ensured by successive application of the (Q^2+1) method discussed earlier. Starting with the step up in impedance, a shunt *C* is added in parallel with the smaller output impedance, R_1. The Q associated with this parallel *RC* is designated Q_p. The value of R_1 is transferred to a smaller series value and the tapped capacitor components are all in series. The ratio of their composite series reactance, which is increased in value to the smaller series R_1 value, provides for a larger Q and hence a substantial increase in R_1. Now R_1 is equal to the desired R_2 and in proportion to the desired operating Q. The center frequency, the operating Q and hence the desired BW are all given as follows: f_o = 10 MHz, and the operating Q desired is Q_o = 50, hence the BW is from **(3)**, BW_{3dB} = 300 kHz. The capacitive tap circuit values are found as follows. The equation set for *C* tap impedance transform with operating $Q_o \geq$ 10 below, and is referred to later as *Table 1*:

$$C_t = 1/(2\pi \mathrm{BW} R_2)$$

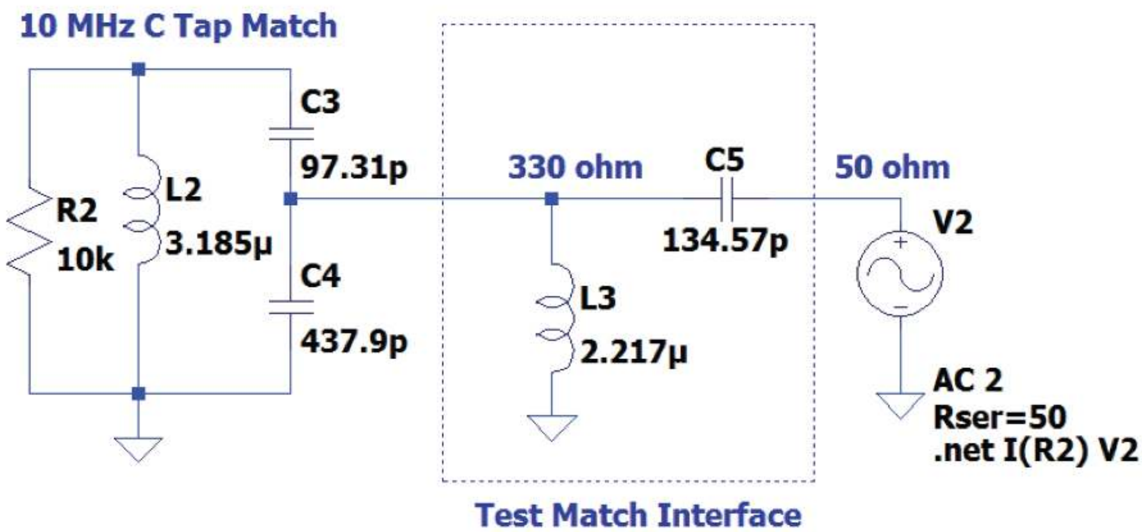


Figure 7 — A pre match front end assists in managing this limitation of Figure 06A and provides a 50 Ω interface.

$$L_t = \frac{1}{(2\pi f_o)^2 C_t}$$

$$Q_o = \frac{f_o}{BW}$$

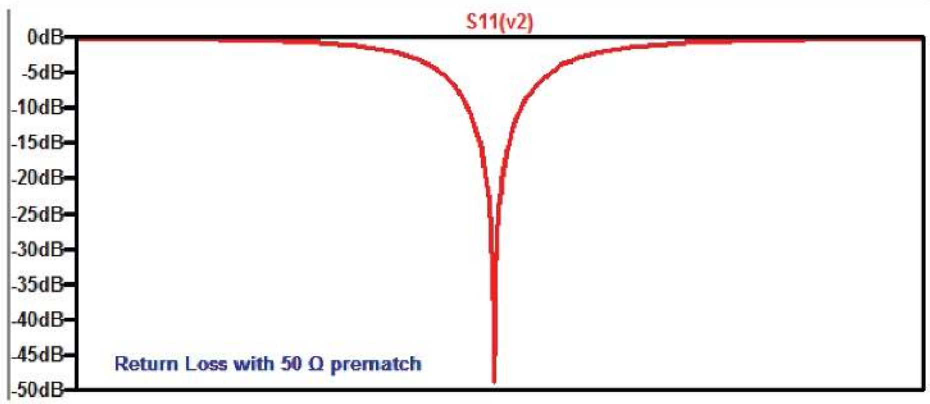
$$N = \sqrt{\frac{R_2}{R_1}}$$

$$Q_p = \frac{Q_o}{N}$$

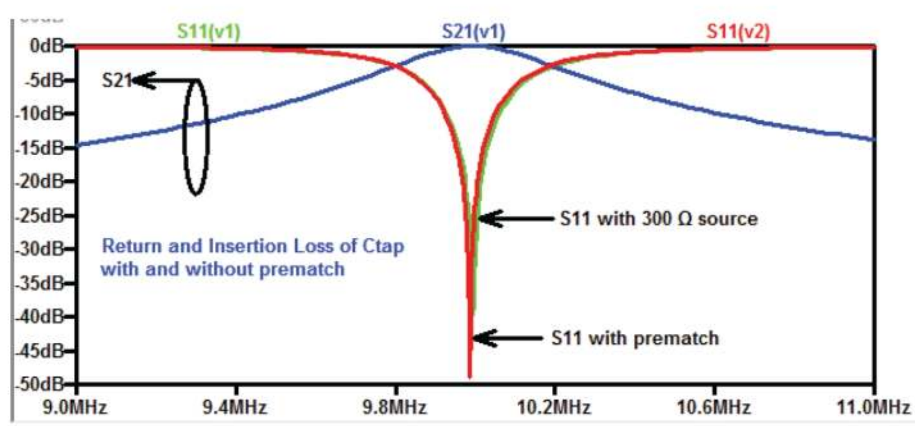
$$C_2 = NC_t \text{ and } C_1 = \frac{C_2}{(N-1)}$$

This completes the closed form design sequence. Following these design steps, the circuit schematic is shown in **Figure 6A**, and the circuit response is shown in **Figure 6B**, which results in using an N of 5.5 and Q_p of 9.08. The capacitor tap provides the impedance step up and resonates with the tank inductance of 3.185 µH at 10 MHz. The resulting BW and Q should coincide with a 10 MHz center frequency and a 200 kHz BW. Hence, a doubly loaded circuit will have a BW of 400 kHz.

The generator source for this simulation sweep is 300 Ω. However, the desire is to use a 50 Ω source, so a pre match is required, **Figure 7**. Using the Ell network will provide a 50 Ω match to 300 Ω with a series Q of only 2.24, see **equation (4)**. This Q is much less than the *C* tap matching Q, hence, it will have a minor impact on the total Q and the measured BW. The design of the Ell proceeds as in the prior example. The series *C* required is 135 pF while the shunt *L* is 2.2



(A)



(B)

Figure 8 — (A) The return loss response without the pre-match. (B) — The return loss response with the pre-match is indistinguishable from Figure 8A. This is the case where the Q of the pre match is significantly less than the circuit under test. If this is not the case, other measurement methods are possible.

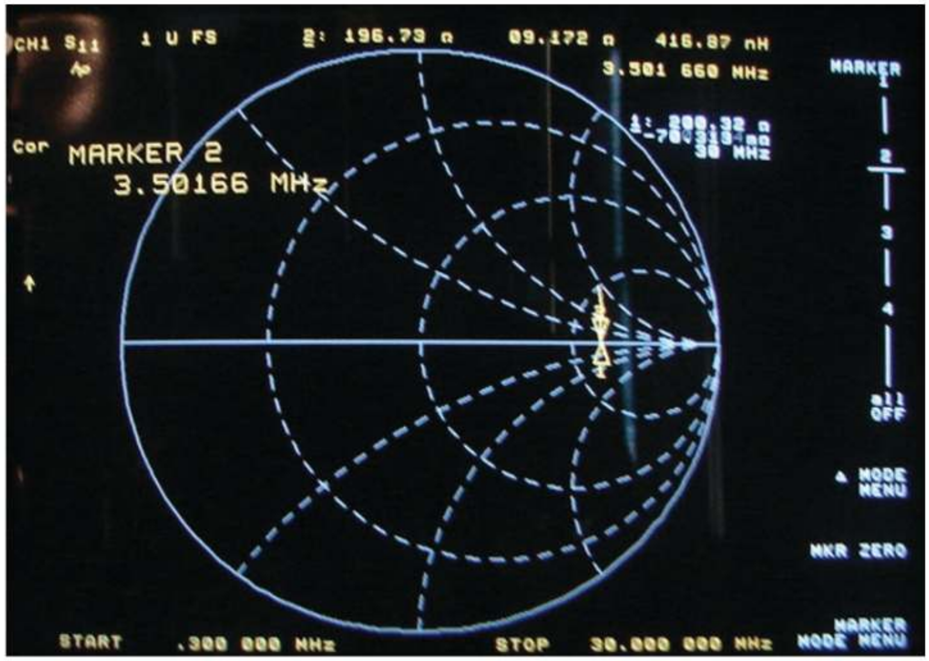
μH. This circuit provides the pre match and the resulting cascade of matching circuits and the response is shown in **Figure 8A**. The response measurements without the pre match, **Figure 6B**, and with the pre match, **Figure 8B**, are indistinguishable. Again, the advantage of the pre match is it brings the measurement point down to 50 Ω and facilitates measurements with standard test equipment.

If a pre match circuit cannot be constructed in a manner that permits its operating Q to be less than the circuit under test, then there are several other possible approaches. Lumped element transformers of the unbalanced to unbalanced (UNUN) form provide one solution. A 2:1 turns ratio transformer is constructed with a core of material 73. This transformer provides either a 200 Ω or 12.5 Ω Z_o system, see **Figures 9A** and **9B**. The transformer permits testing a match in a single one-port measurement. A family of these transformers would provide the ability to test a range of transformed impedance values. For example, a 4:1, 9:1, 16:1 and 49:1 are all popular designs that permit the application of a Z_o of 50 Ω for testing while providing a higher or lower termination to be applied to the circuit under test. As an example, a 200 Ω to 3.9 kΩ match is discussed next applying this technique.

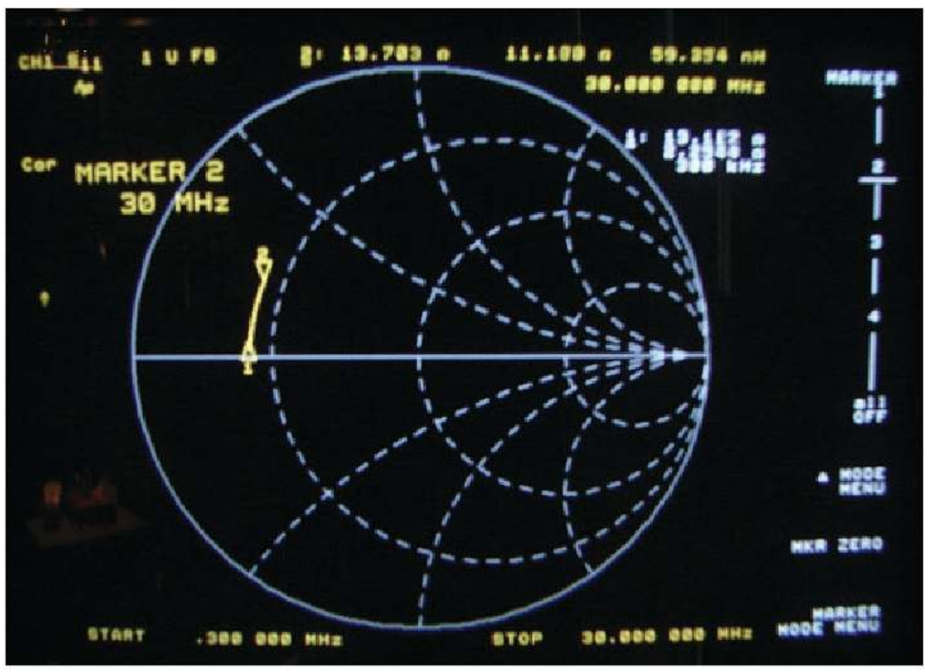
The transformer used for **Figure 9A** is constructed on type 73 material with 2 and 4 turns of #28 AWG enamel wire. The sweep range shown is 300 kHz to 30 MHz.

An application of the 4:1 impedance ratio transformer is used in the evaluation of an Ell match operating at 40 meters. A transform from 3.9 kΩ to 200 Ω is desired. The design Q is set by the constraints of the resistor ratio, $\sqrt{3900/200}$ or 4.4. Evaluation circuit for this match is shown in **Figure 10A** and its operating Q response is in **Figure 10B**.

The Ell match with low operating Q is quasi low pass. There is not significant peaking in the frequency response. Nevertheless, the response below the center frequency for a low pass impedance transform will display a return loss value that follows the graph of **Figure 3** and is useful for checking operating Q. Since the return loss is in excess of 20 dB, the 3 dB return loss value is located at 5.35 MHz, see **Figure 10B**. The Q_o value is 713/(713-535) or 4 and is the desired Q_o requirement. Since the half BW is taken directly from **Figure 10B**, there is no need to double the resulting value.



(A)

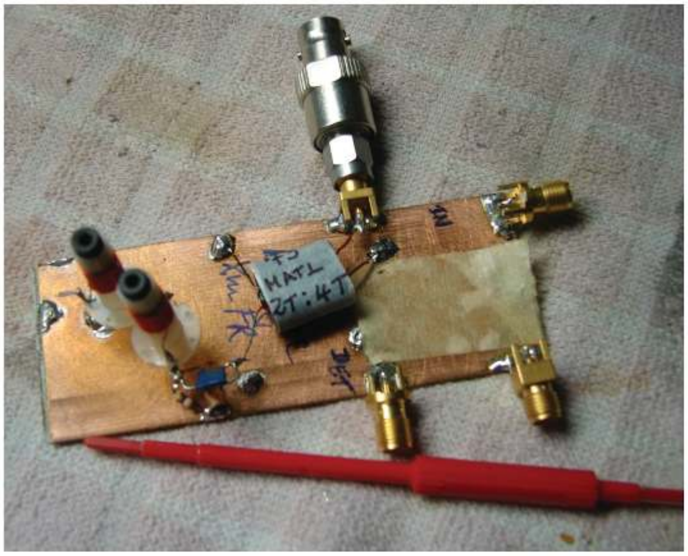


(B)

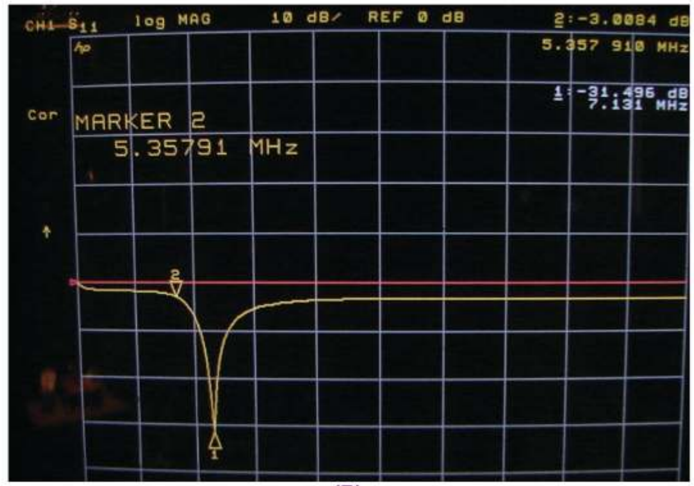
Figure 9 — (A) A 4:1 transformer will serve as a 200 Ω source from 50 Ωs. (B) — A 12.5 Ω source when terminated into 50 Ω is possible with the same transformer.

Why the Q Measurement is Useful?

The return loss is an excellent indicator of the quality of the match. Ideally, the return loss should be 20 dB or better which equates to an SWR of less than 1.25:1. The power loss is small and the efficiency should be high. However, the return can be masked by dissipative losses. Resistance in circuit components is real and is part of the matching system. Unfortunately, that resistance which contributes to the quality of the match also contributes to the loss of power. Obtaining a quality match will be pointless if the match is aided by dissipative circuit elements. Identifying that condition is possible by looking at the operating Q of the match as well as the return loss. If the BW of the match or the impedance transform does not come close to the desired design value

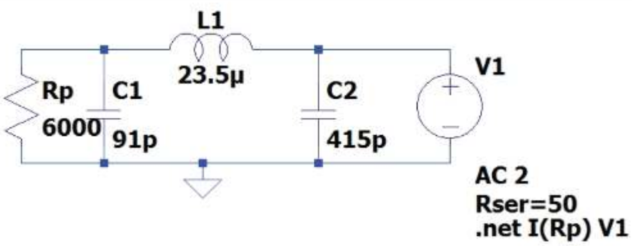


(A)

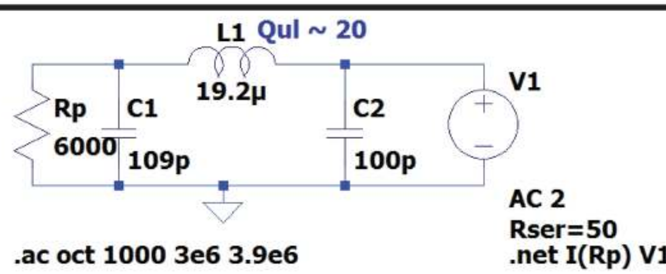


(B)

Figure 10 — (A) A 4:1 transformer is constructed and operates as a 200 Ω source in a 50 Ω measurement system. (B) — The Ell network return loss is swept and the operating Q and quality of the match provided by the response.

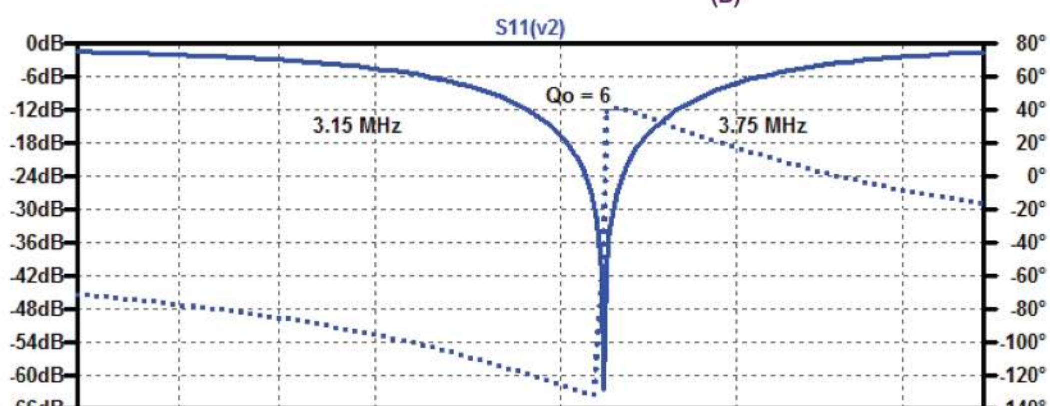


(A)

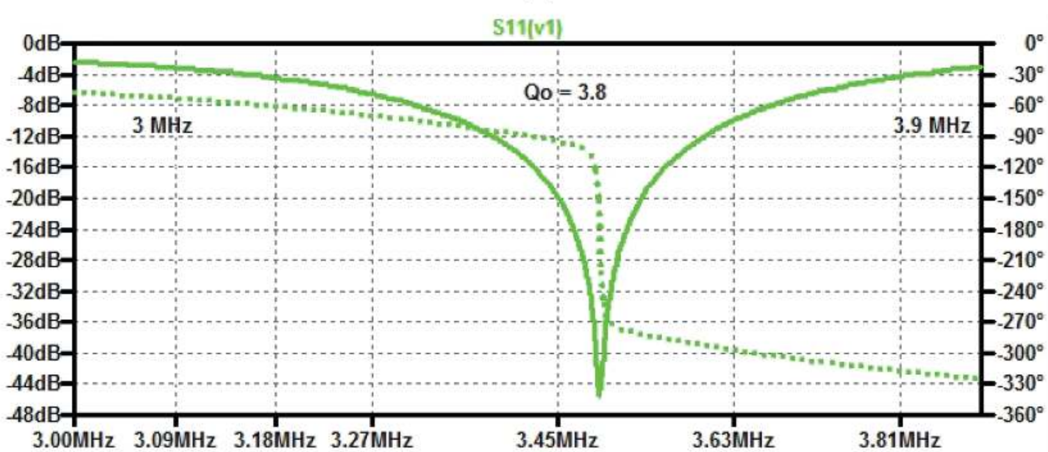


(B)

Figure 11 — In (A) and (B) the Pi circuits are used to transform 50 Ω to 6 kΩ, which appear excellent; however, inspection of the one port BW in (D) is significantly less than (C) and the disparity is due to circuit losses. (11C) is the doubly loaded lossless element demonstrating an operating Q, of 6; (D) is the same topology network as (C) however, the inductor is not lossless and its Q value affects the final response.



(C)



(D)

while meeting the desired match quality, it could be due to excessive low Q elements in the matching circuit. **Figures 11A and 11B** illustrate these cases. A Pi match is designed at 3.5 MHz to transform an antenna *Z* of 50 Ω to a tube with plate resistance of 6 kΩ.

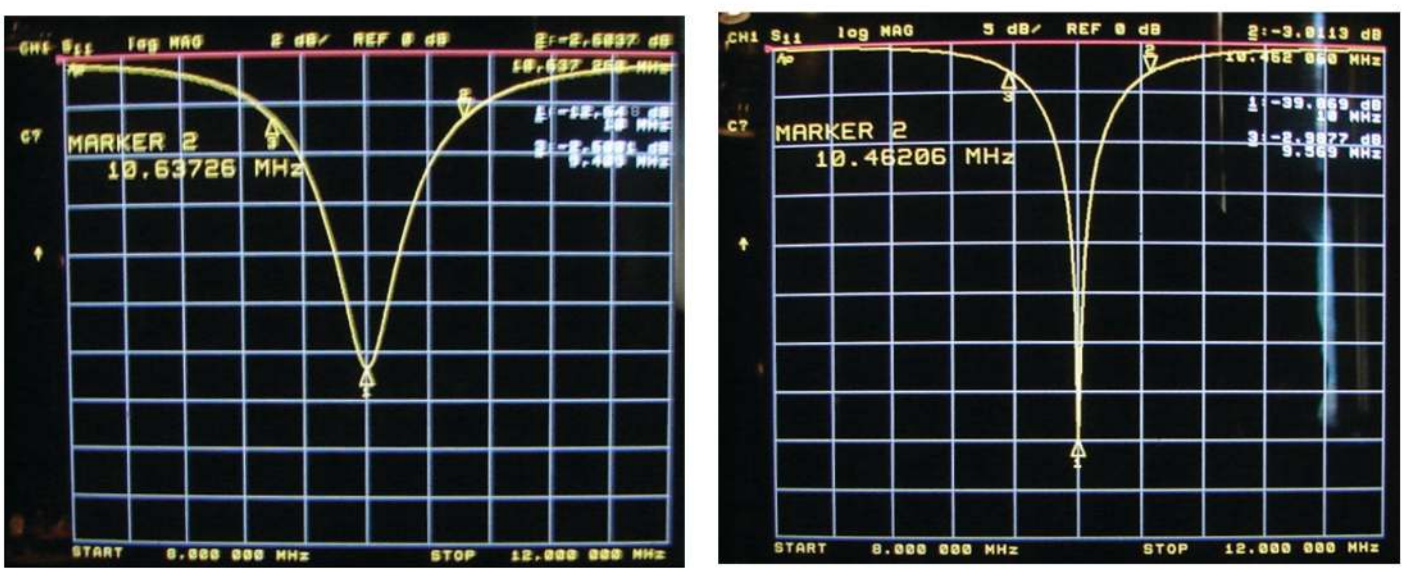
The Pi circuit of **Figure 11A** is used to transform 50 Ω to 6 kΩ. The Pi match of **Figure 11B** also appears excellent. However, inspection of the one port BW in **Figure 11D** is significantly less than **Figure 11C** and the disparity is due to circuit losses. The doubly loaded lossless element Q's in **Figure 11C** demonstrate an operating Q, of 6. **Figure 11D** has the same topology network as **Figure 11C**, however, the inductor is not lossless and its Q value affects the final response.

In the circuit, **Figure 11A,** this is completed with high unloaded Q components. The desired circuit design Q is 12 as provided by design tables **[8]**. This is a singly loaded Q value. The Pi match represents a circuit which is either doubly loaded and is designed as an impedance matching network or, the Pi circuit will operate as a termination circuit, singly loaded. Interesting to note, many of the design notes and approaches to the component values for tube and solid state amplifiers are based on a singly loaded circuit. This makes sense as the Pi network is used as a termination circuit, not necessarily as a conjugate match circuit. However, the design equations use a doubly loaded schematic configuration and neglect the second termination or the load. This is the case as provided for in **[8]**. This introduces confusion. A treatment of this problem and a universal solution is provided in **[9]**. An abbreviated detail of the Pi network arrangement and dissection is provided in **Appendix II**. The component element values are highlighted at each step.

Continuing with the example of **Figure 11**, as previously discussed, the Pi network design Q is not necessarily the same as its operating Q, Q_o of 12. The actual $Q_o = 6$ is doubly loaded. The (Q^2+1) formulation and application as discussed earlier when followed will demonstrate this result and it will become apparent, there are two parallel *RC* circuits. One associated with the tube plate load resistance, if a physical *R* is there and the other a transmission line termination. Hence, there are two possible parallel *RX* circuits that are transformed to their series equivalent values and so two Q values are present, see the circuits in **Figure 11**. The series equivalent for R_p of 6 kΩ at the source side and the series equivalent of R_{SER} of 50 Ω are added as well as their reactances. The net is a doubly loaded Q value of approximately ½(12+0.5) or 6.3. The response Q as displayed in **Figure 11A** is the final operating Q and not the design value of 12. If, the source termination is absent and the Pi network is to serve as only a load termination, then $Q_o = 12$.

As shown in **Figure 11A**, the quality of the match is good and the return loss is better than 40 dB. Hence the BW is found by the measurement of the 3 dB return loss frequency points, 3.15 and 3.75 MHz. The center frequency is 3.5 MHz. This equates to a doubly loaded Q of 6. The circuit in **Figure 11B** also demonstrates an excellent return loss, better than 40 dB. However, the design BW is not achieved and is wider. The element values are contorted to achieve the match, but without being the wiser, one might expect some other problem is to blame and all is well! The issue could be detected in a two port measurement. The circuit of **Figure 11B** would display higher S_{21} or transmission loss since the circuit components had a lower unloaded Q. However, the single one port measurement is more convenient and at least would have prompted the builder or designer to dig deeper.

Another case is the design and construction of a *C* tap circuit to match a ceramic filter to an integrated mixer, see **Figure 12C**. The filter required a 330 Ω termination while the mixer output *Z* is fixed at 1.5 kΩ by its internal termination. To study this case, the previous 4:1 transformer is used and the 330 Ω is changed to 200 Ω as a means of checking the design. The *C* tap design uses the equation set of *Table 1*. The desired BW is 500 kHz at a center frequency of 10 MHz, hence a Q_o of 20. The resistance ratio defines N, equal to 2.73 and the resulting L_t, C_t, and *C* tap values are as follows: 1.2 µH, 212 pF, C_1 = 330 pF and C_2 = 578 pF. The inductor is a tunable slug form and the 578 pF consisted of 560 pF in



(A) (B)

Figure 12 — (A) The initial return loss response of the match was 12.5 dB at 10 MHz. This return loss is less than desired and the BW defined by a return loss of 2.5 dB is too large resulting in a design Q of 16 instead of 20. (B) — The termination is increased to 3.3 kΩ so that the inductor finite Q and its shunt *R* provide the desired match termination of 1.5 kΩ. The return loss is now over 30 dB and the 3 dB return loss points coincide with the desired design Q of 20.

parallel with 18 pF. It is assumed that the inductor unloaded Q is sufficiently large (a bad assumption). Upon construction and measurement, the results were disappointing. Immediately, the following is noted. The tunable inductor L value had to be increased in value to bring the circuit into resonance, the return loss is just under 12 dB and the BW is too wide. See **Figure 12A**. If the inductor has significant series loss, the resonant frequency is increased. Hence, the tunable L value inductance had to increase to compensate. Furthermore, the match is significantly in error as the shunt R_p of the inductor along with the 1500 Ω fixed shunt termination resistor present will create a match error and a decrease in operating Q. This surprising chaos of problems is all due to just one issue: insufficient inductor unloaded Q. Removal of the proper shunt termination R of 1.5 kΩ and replacing it with 3.3 kΩ, resulted in the desired response. See **Figure 12B**. Although an impedance match is achieved, the required termination is improper! The desired termination is 1.5 kΩ, not 3.3 kΩ. The only way to correct this issue is by accepting a lower operating design Q or selecting a higher unloaded inductor Q. **Figure 13** shows a capacitive C tap match constructed and a 4:1 impedance transformer provides the required 200 Ω source to test the match as a one port.

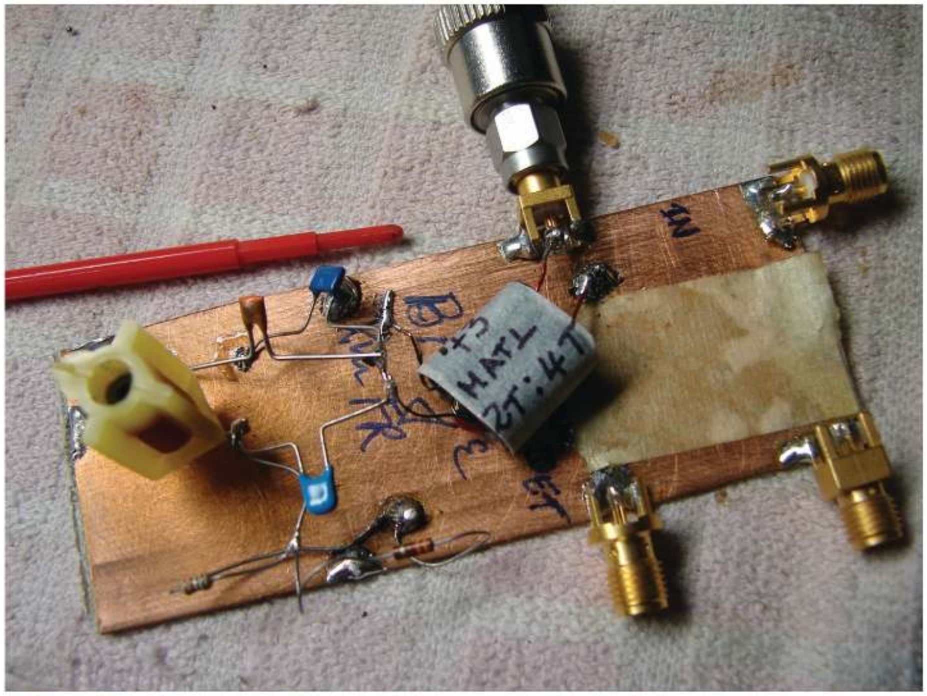


Figure 13 — A capacitive C tap match is constructed and a 4:1 impedance transformer provides the required 200 Ω source to test the match as a one port.

Conclusions

A review of various impedance transforming or matching circuits are provided. There are various Q definitions and these are used in both designing and evaluating the operation of these circuits. When transforming circuits are used, the Q value provided is singly loaded and agrees with the Q values assigned to simple transforming networks like the Ell. However, circuits applied in matching are doubly loaded, both a source and a load termination are present. Therefore, the evaluation of their measured Q must be doubled to be in line with the assigned Q in their design. Distinguishing the doubly loaded Q and the singly loaded Q, or the operating Q and the design Q, should reduce the confusion between design and measurement results. Also, it is key to realize, some so called matching circuits are not providing a match at all, but instead are serving as termination networks.

A method of finding the operating parameters of an impedance matching or Z transformation circuit is provided. It is a simple measurement requiring only one single port measurement and is completed with any of the following instruments: an antenna analyzer, a directional coupler or SWR bridge with good measurement accuracy **[10]**, an RF signal generator with scope, a spectrum analyzer, a noise bridge, or a VNA. The measurement is particularly simple if the load termination is 50 Ω as this directly interfaces with the most standard test equipment. The single one port measurement returns the quality of the match and the operating BW. If the terminations are not standard values or the same as provided by the test equipment, several alternative approaches are possible. These include a pre match circuit, a transformer or UNUN, or a scope and an RF source.

Appendix I

The insertion loss of a tuned circuit is directly related to the unloaded and loaded Q. Consider the series tuned circuit shown in **Figure A1**.

The value of R_s is contributed partly from the loss of the inductor and capacitor and forms a voltage divider with R_L. At resonance, if R_s is small, the voltage available from the source appears totally across R_L. A small R_s per **equation (1)** or **(A2)** is provided by a large Q_{UL}. As R_s increases, the voltage transfer to R_L decreases and the loss is higher. The loaded Q and the unloaded Q, Q_L, and Q_{UL}, which are

$$Q_L = \frac{X_T}{R_L} \tag{A1}$$

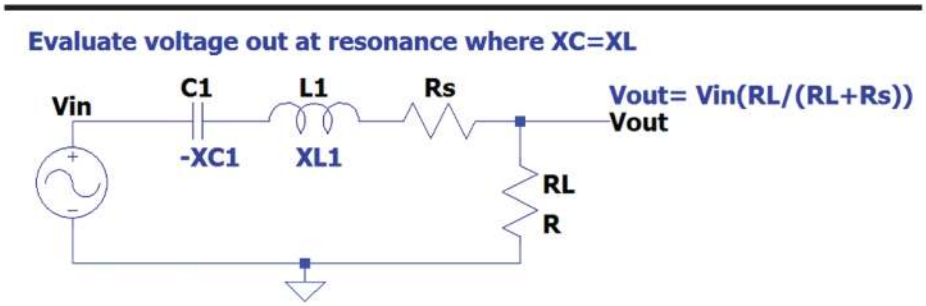


Figure A1 — A series tuned circuit identifies the loss relation between the loaded and unloaded Qs.

and

$$Q_{UL} = \frac{X_T}{R_s} \tag{A2}$$

are written in terms of the voltage divider rule as

$$V_o = V_{in}\left(\frac{1}{1+\frac{R_s}{R_L}}\right). \tag{A3}$$

Using **equations (A1)** and **(A2)**, **equation (A3)** is recast in terms of Q.

$$V_o = V_{in}\left(\frac{1}{1+\frac{Q_L}{Q_{UL}}}\right) \tag{A4}$$

Hence, lower loss occurs in matching and impedance transforming circuits where a large ratio of Q_{UL}/Q_L is provided.

Appendix II

A detailed breakdown of the Pi network as a transform circuit or as a matching circuit is outlined here. More details and further insight is provided in **[9]**. The Pi circuit can be disassembled and put into either a series (steps **Figures A2A** and **A2B**) or a parallel form (steps **Figures A2C** to **A2F**). In either case, it is noted that the Pi circuit is after all, a single tuned resonant arrangement of *L* and *C* with resistive loading from the source (input) and the load (output). If the source or load are absent, then the frequency response is singly loaded, otherwise it is doubly loaded. Proceeding from **Figures A2A** to **Figures A2B**, find the series equivalents for 6 kΩ in parallel with 91 pF and 50 Ω in parallel with 415 pF.

Apply **equation (2)** at 3.5 MHz. The load side reactance of 415 pF is 109 Ω. Then the Q_p value on the load side is 50/109 or 0.46. The same equation is applied to the source side; the reactance of 91 pF is 500 Ω. Then the Q_p value on the load side is 6000/500 or 12. The source side Q dominates and one would expect this value to set the final operating Q, Q_o. The next step is to convert the Pi circuit from **Figure 2A** to a complete series circuit as seen in **Figure 2B**, using **equation (A5)**.

$$R_s = R_p \big/ \left(Q_p^2 + 1\right) \tag{A5}$$

or the series *R* values at each side are: 6000/145 = 41 and 50/1.21 = 41.

As expected they are equal and a match is provided. Furthermore, the series equivalent reactances from each side must be found. Their total must equal the remaining series inductance from the original Pi circuit of **Figure A2A**. Since the series and parallel Q values are identical at resonance, apply **equation (1)** to get each series reactance. On the source side,

$$X_{s1} = 41 \bullet 12 = -j496$$

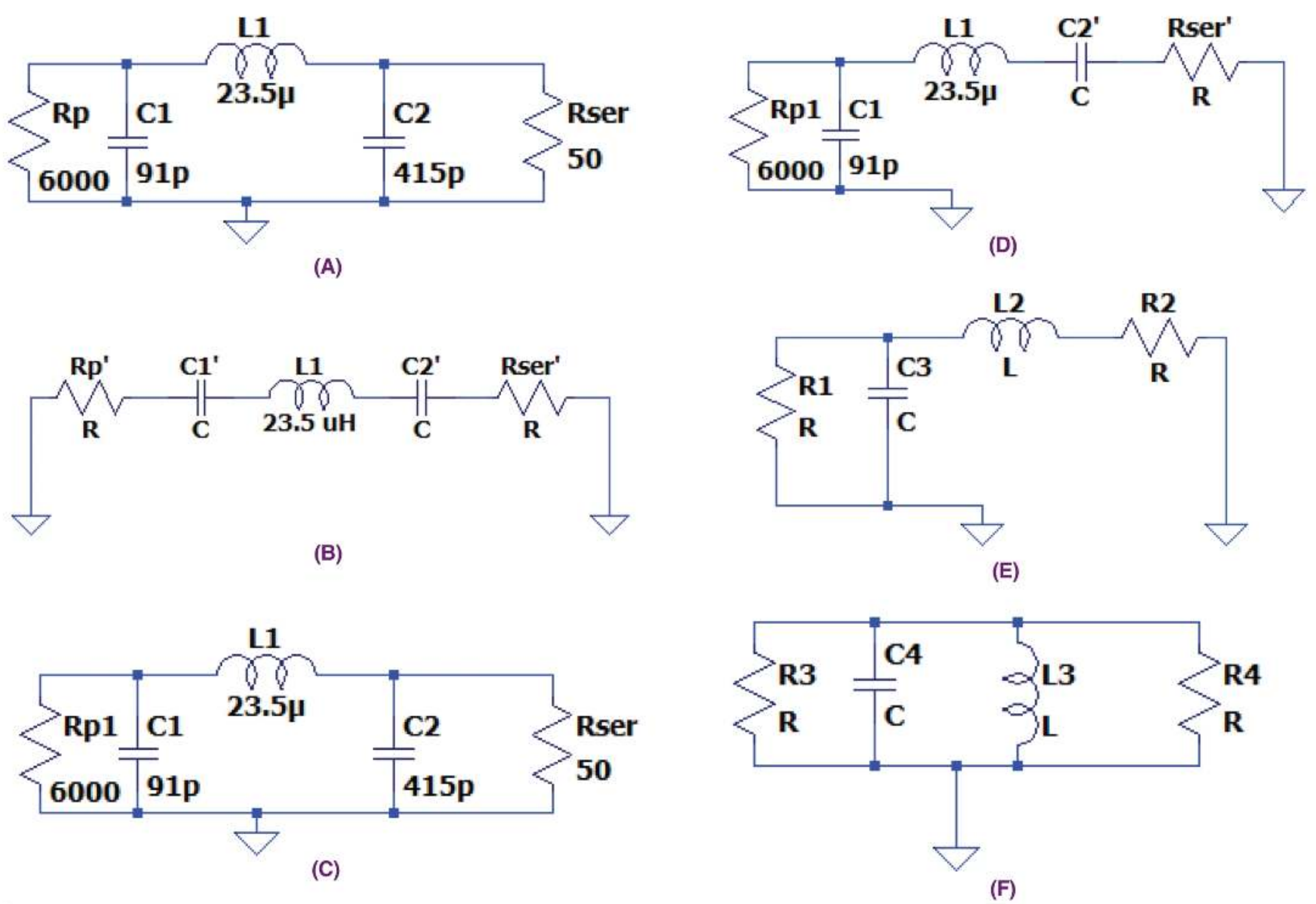


Figure A2 — Circuit A is the final resulting Pi network. (B) is Circuit A replaced with parallel to series equivalent circuits for the source and load side of the A network. Circuit (C) is the final resulting Pi network. (D) is a transform from parallel to series form for the load side. (E) consolidates the series load side reactances. (F) transform the resulting consolidation of reactances to their parallel form.

While on the load side,

$X_{s2} = 41 \bullet 0.46 = -j19.$

Their sum is $-j515$, which cancels the positive reactance of the 23.5 µH inductor. The Q_o is given by the total series reactance compared to the total series resistance, 515/82 ~ 6. Hence, the actual BW if measured as a doubly loaded circuit would also be 6. However, as used in a power amplifier tube circuit, the tube termination or 6 kΩ is absent. The BW then would be dictated by a Q_o of 12. A similar result should be expected if the circuit progression were to start with **Figure A2C** circuits through **Figure A2F** below and noting that **Figure A2B** and **Figure A2D** are equivalent. This process follows next.

Step 2) in creating the parallel tuned circuit from the Pi circuit is the same as before. Converting the load R and the shunt C of 50 Ω and 415 pF to their series form. As already discussed, the values are 41 Ω series with $-j19$ Ω. Next combine the 23.5 µH series inductor with the series capacitor and the total resulting reactance's add, $-j19 + j515$ resulting in $+j496$ Ω. This is circuit **Figure A2E**. All that remains now is to convert the series R_L of **Figure A2E** to its parallel equivalent and note if the completed parallel circuit provides a match as well is a resonant tuned circuit. The series circuit is 41 + j496 Ω, hence the Q_s is 12.12 and the parallel R, which result is from **equation (A5)**: $R_{p_s} = R_s\left(Q_s^2 + 1\right)$.

This is approximately 6 kΩ and the X_p that results is $j500$ Ω. Thus the 91 pF is tuned and its reactance canceled at 3.5 MHz and the circuit is again matched. The Q_o as before is approximately 12 if singly loaded, otherwise it is 6 if doubly loaded.

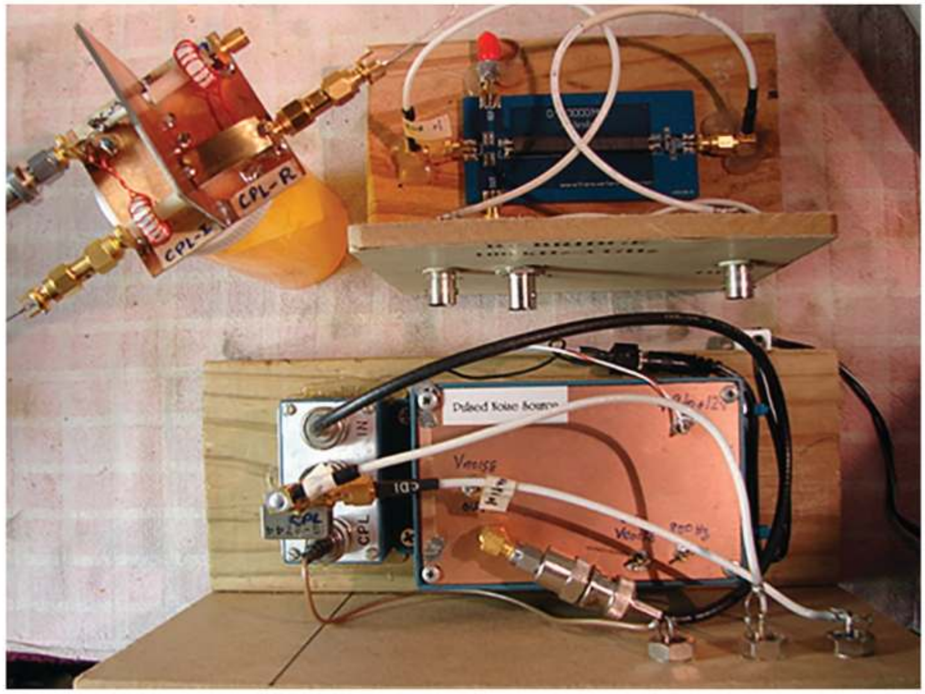


Figure A3 — Measurement instrument examples include directional couplers (reflectometer) driven by a diode noise generator (bottom) or signal sources such as a GDO (grid dip oscillator) or RF signal generator. A tandem directional coupler (upper left) and an antenna bridge (center top) can both be used with an oscilloscope as the display instrument. All of these devices are the heart of an SWR meter.

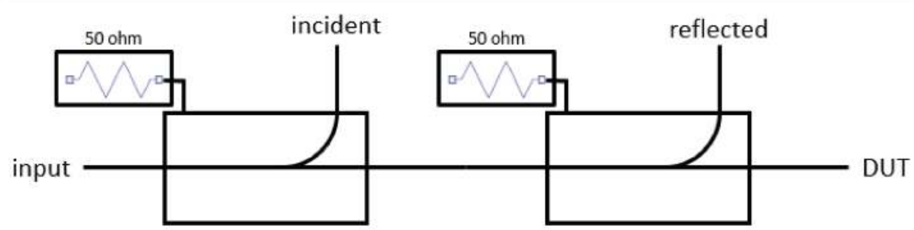


Figure A4 — Directional couplers with internal terminations are cascaded to make a coupler with dual ports. One port provides the incident voltage measurement while the other, the reflected. These couplers are an asset to an accurate SWR meter.

Appendix III

A variety of measurement instruments are possible, see **Figure A3**. These instruments provide the ability to determine the SWR or the return loss accurately. These include a reflectometer, **Figure A4**, a RF bridge and a tandem directional coupler **[11]**, **[12]**, **[13]**. Some form of calibration is usually required. The primary interest in calibration is to determine the maximum and minimum capability of recording SWR or reflection coefficient values. A coupler Z_o of 50 Ω will dictate the lowest SWR reading with a 50 Ω termination and this termination is used as one calibration point. The other calibration point is a perfect reflection termination and a short circuit such as a copper plate across the measurement port to ground will do. The calibration completed, reference points established will highlight a 1:1 SWR and an infinite SWR. The measurement of the quality of the match and the BW leading to the Q_o value is now straightforward.

The output from a test set is shown in **Figure A5**. A noise generator drives a pair of couplers configured to sample the incident and reflected noise power, **Figure A4**. The display is on a spectrum analyzer, but a shortwave receiver could be used with the AGC turned off and the audio output sampled. The analyzer display is a noise voltage and the maximum value is the short circuit reference value at −52 dBm. The center frequency reflected power is −71 dBm, nearly a 20 dB return loss. The 3 dB BW points can be determined from the markers set at −55 dBm, a 3 dB delta. The BW is 1.26 MHz and therefore a Q_o of 5.6 and upon doubling a design Q of 11.2, quite close to the design value of 12.

[Photos by the author]

Alan Victor, W4AMV, was licensed in 1964. He operates mostly CW using an all homebrew station and enjoys design, construction and restoration of communication and test equipment. Alan worked in both the communication and semiconductor engineering fields. He received his PhD in electrical engineering from North Carolina State University and is currently involved with their mentorship program assisting new graduates in their engineering studies.

Figure A5 — A wideband noise generator using a Zener diode operating near breakdown voltage provides an excellent source. The Pi match network response is displayed. The match frequency is 7 MHz and the Qo = 5.6. The markers 2D and 3D provide the 3 dB return loss value and hence the BW. The design Q value is 11.2 and in close agreement with the measured response for the doubly loaded Q.

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[13] Tandem couplers: **k6jca.blogspot.com/2015/01/notes-on-directional-couplers-for-hf.html**.

Errata – *QEX* July/August 2022

In "Tuned Transformers," by Gérald Julien Lemay, VA2GJ, an error crept into an equation on p. 31 column 1, lines 16-17, and repeated on p. 32 at the top of the third column. The corrected equation in both places is:

$(3/24)^2 Z_L = Z_L/64$.

Also, the label (6) should be on the same line as the equation. Thanks to James Whitfield, N5GUI, for spotting the errors.

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Versatile Lock-In Amplifier Design

A lock-in amplifier can be adapted for QRSS CW on the 2200 m band.

When it comes to extracting extremely weak coherent signals from background noise, nothing performs better than the venerable lock-in amplifier and no other technique likely ever will. The lock-in amp is standard instrumentation in the scientific community, used in everything from medicine to metallurgy. Its use is almost unheard of in amateur radio.

However, there is a price to pay for using the lock-in method: it is an extremely narrow band method. It is unsuitable for high-speed data communication, but is excellent for detecting man-made and natural unmodulated radio frequency signals.

This project describes a lock-in amplifier specifically designed for ULF detection but can easily be adapted for other uses, namely, QRSS CW communications on the 2200 m band. The goal for this project is to design and build an affordable versatile lock-in amplifier using easily accessible parts. The parts used for the prototype consist of logic gates, D-flip-flops, and op-amps, which are common chips in most electronic labs. The frequency that is of interest for this prototype lock-in amplifier is 1.5 kHz. The first 1.5 kHz prototype is shown in **Figure 1**.

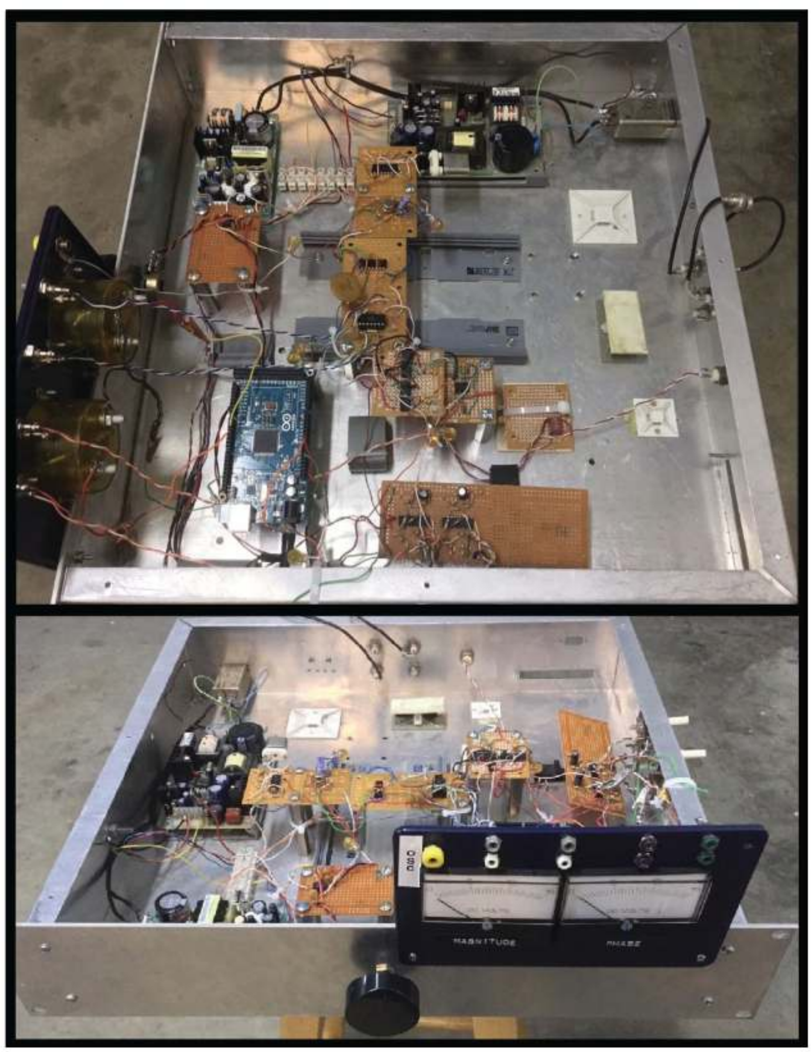

Figure 1 — Prototype ULF lock-in amplifier.

Block Diagram

The following describes two channels in phase quadrature. The block diagram for the prototype lock-in amplifier is shown in **Figure 2**. The lock-in amplifier starts out by generating a square wave using the NE555P

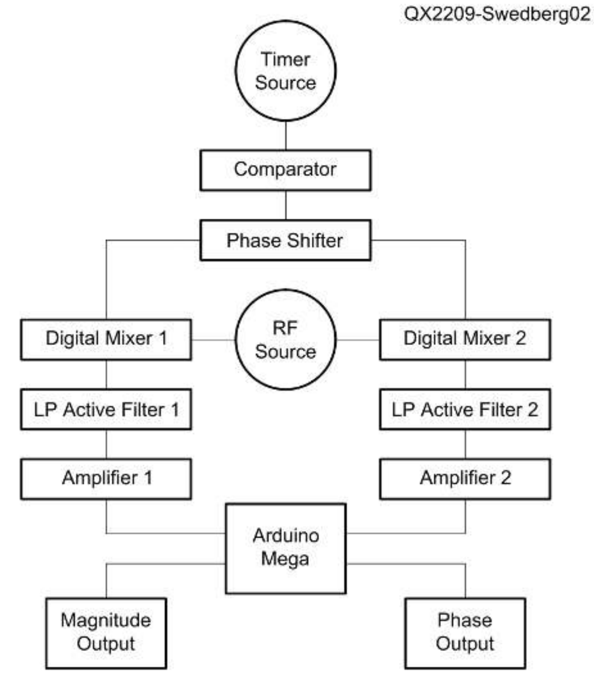

Figure 2 — Block diagram of the lock-in amplifier.

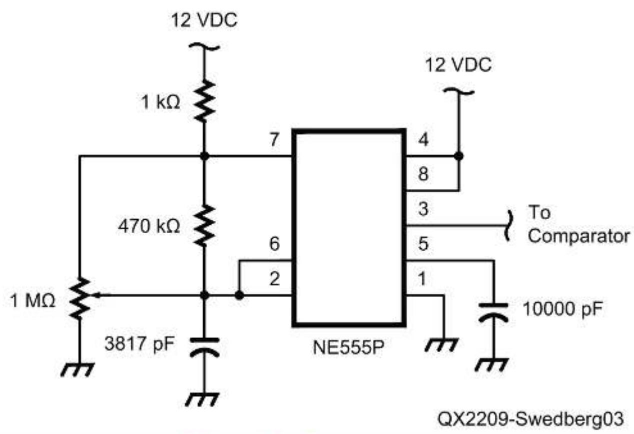

Figure 3 — Timer source.

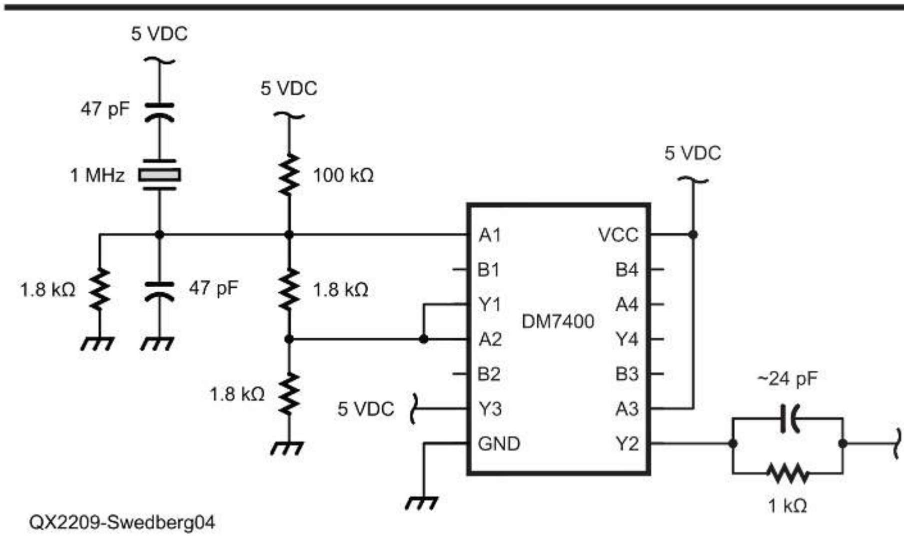

Figure 4 — TTL crystal-oscillator source.

timer chip. The square wave then passes through a comparator and a phase shifter. The output is then passed through a digital mixer to mix the signal with a RF source. The IF signal is then passed through a low-pass filter, which eliminates all frequencies, except for a near-DC voltage component. A series of high-gain DC amplifiers are then used to amplify the signal. Treating the two DC output signals as a Cartesian point at an instantaneous point in time, the magnitude and phase for this point is calculated using an Arduino Mega.

Timer Source

The schematic for the timer source is shown in **Figure 3**. The timer source utilizes a NE555P timer chip with a VCC of 12 V. The timer chip is configured as an astable circuit with a 12 V peak-to-peak square wave output. The frequency of the square wave output can be varied from 2.5 kHz to 21 kHz by changing the 1 MΩ variable resistor. The source frequency needs to be set to twice the desired frequency, because of a frequency divider that occurs when passing through the phase shifter circuit. The output of the source goes directly into a comparator.

Other Input Source Options

There are several other input source options. One of these options is to use a digital TTL crystal oscillator. This circuit diagram is shown in **Figure 4**. It uses a DM7400 IC with a 1 MHz crystal. A second input source option is to use an analog sinusoidal signal. This is an option because the input source is put into a comparator that will convert the sinusoidal signal into a square wave. However, to vary the input frequency easier, the timer input source was chosen.

Comparator and Phase Shifter

The schematic for the comparator and the phase shifter circuits are shown in **Figure 5**. They are both powered by a 5 V VCC source. The comparator circuit uses an LM533 op-amp chip and a NAND gate from the 7400 IC configured as a NOT gate. If a sinusoidal input source is used, the comparator turns the input signal into a square wave. Using a square wave input source with the comparator will not impact the output of the lock in amplifier. From the comparator, the output is sent through a phase shifter to output two signal pairs.

The first signal pair is created by directly feeding the comparator output from pin Y4 on the 7400 IC into CLK2 input on the DM7474 D-flip-flop IC. The D2 and I2 pins are tied together, and the CLR2 and PR2 pins are tied to the 5 V VCC supply. The output signal pair is taken from pins Q2 and I2.

The second signal pair is created by first sending the output from the comparator into another NAND gate from the 7400 IC. As before, the NAND gate is configured as a NOT gate. The output from the NAND gate is then sent into the CLK1 on the other D-flip-flop on the DM7474N IC. Again, the D1 and I1 pins are tied together, and the CLR1 and PR1 pins are tied to the 5 V VCC supply. The output signal pair is taken from pins Q1 and I1 and has a 90° phase shift with respect to the first signal pair.

Digital Mixer

Several variations of Gilbert cell circuits were used to try and create an analog mixer. However, it was decided to use a digital mixer instead due to the troubleshooting involved with using the analog mixers. The schematic for a digital mixer is shown in **Figure 6**. The digital mixer circuits use the CD4066BE CMOS quad bilateral switch to mix an input pair with the RF source. The CD4066BE runs off a VCC bipolar supply of ±5 V. One digital mixer is required for each input pair. Each output is then sent to its own low pass active filter.

Active Low Pass Filter

The schematic for both active low-pass filters for the two signals is shown in **Figure 7**. The active low pass filters consist of the RC4558 op-amp IC with a bipolar supply of ±5 V. The cutoff frequency of each filter was set to approximately 67 Hz. For each filter, the input signal is sent into the op-amp's non-inverting input. The inverting input is set up with a feedback loop that will set the gain to 2. The output for each filter is essentially a DC voltage. The output for each filter is then sent to an amplifier circuit.

Amplifier

The amplifier circuit (**Figure 8**) consists of a pair of op-amp ICs, which can be two TL072, two TL082, or one of each. The TL072 and TL082 ICs both contain two op-amps each. The op-amp ICs are powered by a bipolar ±15 V supply. Three inverting op-amps are set up in series with a gain of 10. An external feedback loop between the output of the third op-amp and the inverting input of the first op-amp increases the gain by 100. This increases the output from the low pass filter by 100,000.

To keep the amplifier output between 0 and 5 V, a voltage offset is created using a summing amplifier. The output of the circuit and a 5 V supply are each passed through a 100 kΩ variable resistor. Then, the output is fed into a non-inverting amplifier. The gain and offset of the output can then be adjusted by adjusting the variable resistors. Once the output is adjusted to not exceed the range from 0 to 5 V, the two outputs can be connected to the analog pins A0 and A1 on the Arduino Mega.

Arduino Mega

It is important to not connect the amplifier outputs to the Arduino Mega until it is certain that the amplifier outputs do not exceed the range of 0 to 5 V. The amplifier outputs are connected to the analog pins A0 and A1 on the Arduino Mega. The Arduino code is shown in **Table 1 – Arduino code**. The code inputs the signals as Cartesian coordinates

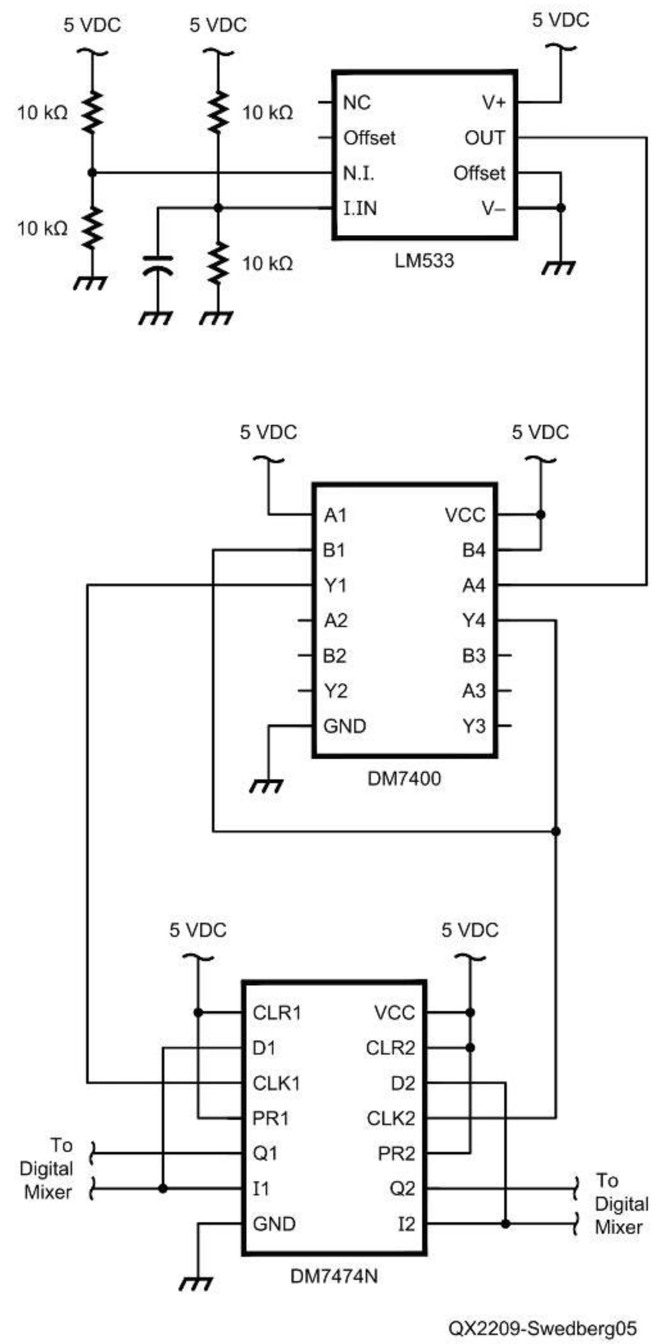

Figure 5 — Comparator and phase shifter.

Table 1 – Arduino code.

```
#include "math.h"
                                        // Define Variables
int X_pin, Y_pin, magn_pin, Angle_pin;
double X, Y;
float magn, Angle;
double pi = 3.1415926535;
void setup()
{
Serial.begin(9600);                     // put your setup code here, to run once
                                        // Set up input and output pins
pinMode(A0, INPUT);                     // X Input Pin for Q
pinMode(A1, INPUT);                     // Y Input Pin for I
pinMode(5, OUTPUT);                     // Magnitude Output Pin
pinMode(6, OUTPUT);                     // Angle Output Pin
}
void loop()
                                        // put your main code here, to run repeatedly
X_pin = analogRead(A0);                 // Read pin values from input pins
Y_pin = analogRead(A1);                 // (analog pins read values from 0 to 1023)
X=X_pin * (5.0/1023.0);                 // Convert X and Y pin values to voltages
Y = Y_pin * (5.0/1023.0);
magn = sqrt(X*X+Y*Y);                   // Convert X and Y to polar coordinates
Angle = abs( atan2(Y, X) * (180.0 / pi) );
                                        // Convert magnitude voltage to a pin value
magn_pin = magn * (255 / 5.0);          // (analog pins write values from 0 to 255)
analogWrite(5, magn_pin);               // write magnitude pin value to output pin
Angle_pin = Angle * (255.0 / 180);      // Convert Angle to a pin value
analogWrite(6, Angle_pin);              // write angle pin value to output pin
}
```

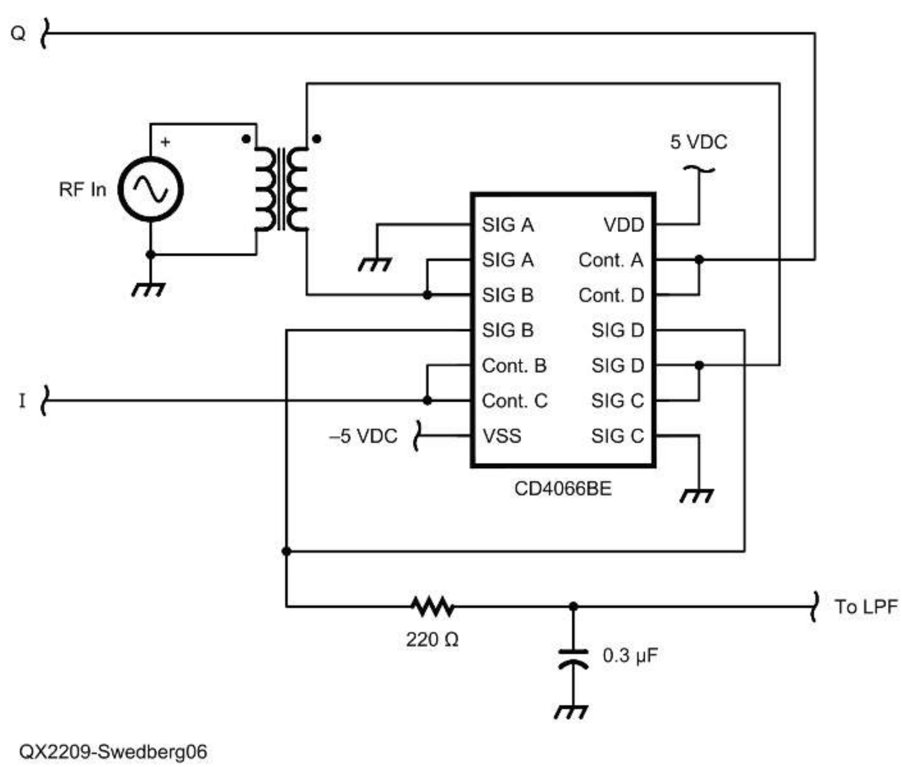

Figure 6 — Single digital mixer.

and uses A0 as the x – coordinate and A1 as the y – coordinate. The coordinates are then converted into polar form. The magnitude and phase outputs are on pin 5 and pin 6, respectively, as voltages.

Future Changes

The next prototype will include changing the 555 timer source input to make it easier to tune to a predetermined frequency. Currently, with the large range of frequencies, it can be difficult to get a square wave at the desired frequency. To help solve this, the plan is to make the range of possible frequencies smaller with the desired source frequency closer to the center. The desired source frequency may also be changed to a lower frequency to help stabilize the 555 timer source.

To make it easier to calibrate the final product, the fourth op-amp that is responsible for setting the offset in the amplifier circuit will be changed. The signal from the third op-amp will be fed into the inverting input of the fourth op-amp, instead of the non-inverting input. The 5 V input will continue to be fed into the non-inverting input of the fourth op-amp. Currently, each variable resistor for the amplifier circuit adjusts the offset and gain. With this change, the offset or the gain can be adjusted with a variable resistor without changing the other.

Another change will be to make the final prototype as small as possible, using fewer circuit boards and wire connections. Since a square wave will most likely be used for the final product, part of the comparator circuit is not necessary and will probably be removed.

Calibration

To calibrate the lock-in amplifier, a switch will be added to the final product to allow the shifted signal to change between a 0° phase shift and a 90° phase shift. With the 0° phase shift, the shifted signal should be identical to the non-shifted signal. Using an oscilloscope and the variable resistors for the amplifier circuit, the gain and offset of the two signals can be adjusted until the signals have the same gain and offset. The offset should be set to exactly 2.5 V, so that 0 V with no offset becomes 2.5 V with the offset. The gain should be adjusted, so that the signal does not exceed the range of 0 to 5 V, which is the allowable voltage range for the Arduino Mega. Once the signals are nearly identical, the switch can be turned back to the 90° phase shift position. When fully calibrated, the magnitude output from the Arduino should be held constant, while the phase angle output should vary.

Kim Swedberg graduated from the University of Alaska Fairbanks with a degree in electrical engineering, and a concentration in computer engineering. At the time of this writing, Kim worked at AlasKit Educational and Scientific Resources and Timbreland North Audio designing and building affordable equipment that is focused on radio wave detection from the ELF to the VLF range. Projects include building a lock-in amplifier and a signal strength meter. Currently, Kim works in the avionics insdustry as a trainee at Aircom Avionics at the Fairbanks Airports.

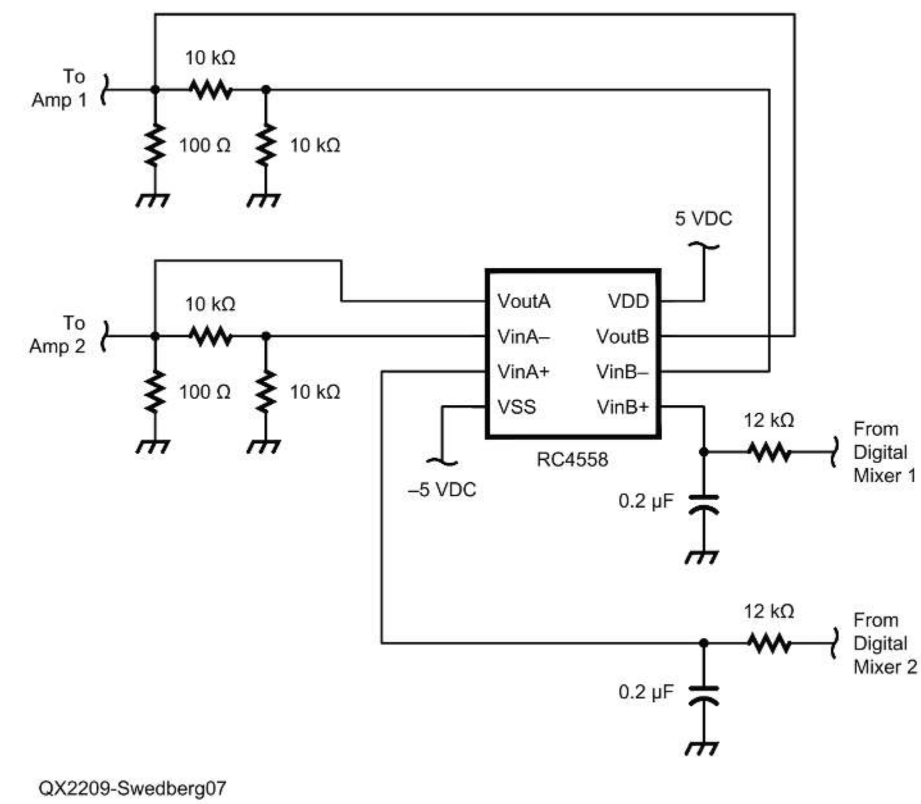

Figure 7 — Active low-pass filters.

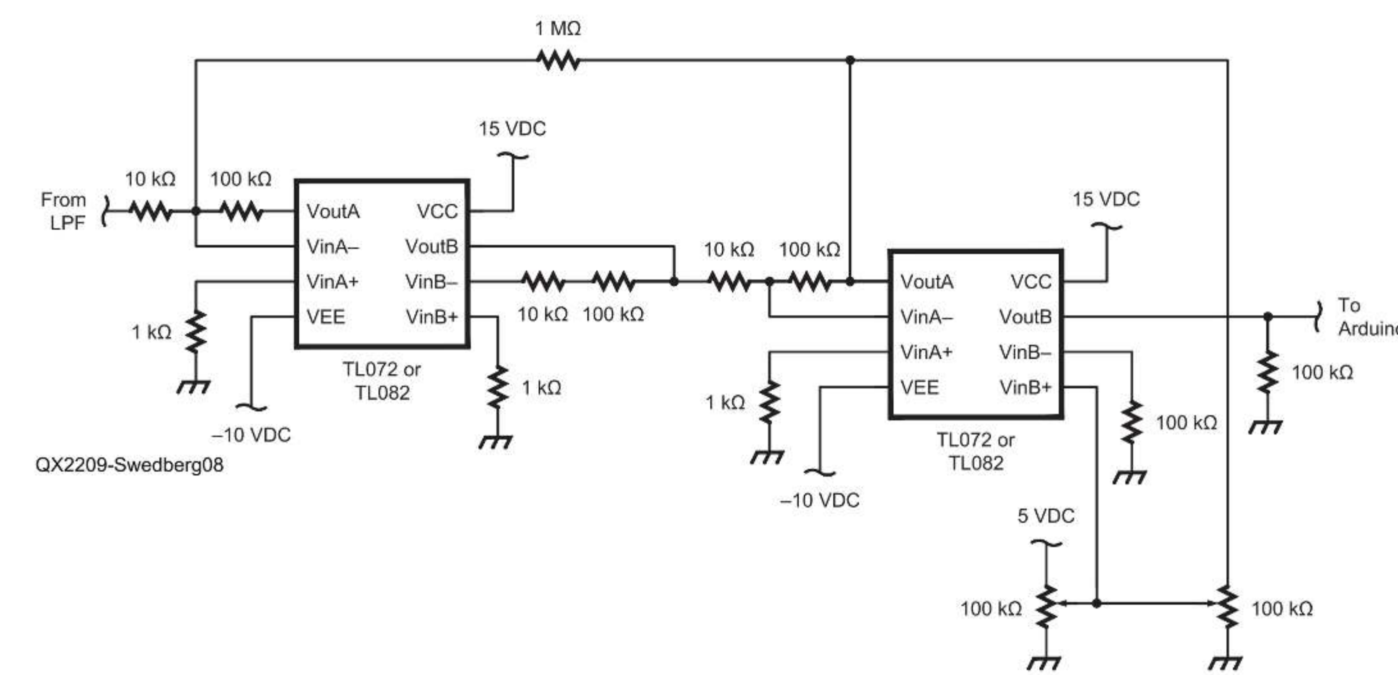

Figure 8 — Single amplifier circuits.

Bob Larkin, W7PUA

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Broadband Tapered-Length Fan Dipole Antennas

Careful choice of wire lengths and number of dipoles produces antennas of very wide bandwidth.

Techniques for increasing the bandwidth of dipole antennas have been applied since the earliest days of radio. Many forms of fat dipoles have been discovered and used. The bi-conical dipole, consisting of a pair of cones, has good bandwidth response and is a good example of this type of antenna. A related antenna is the bow-tie dipole that gives up some performance for a simpler mechanical structure. A recent two-part *QEX* article **[1]** by Jacek Pawlowski, SP3L, demonstrates the ability of multi-wire structures to increase antenna bandwidth enough to cover all amateur bands from 14 to 29.7 MHz. Many of these use wires of similar length, adding bandwidth by creating a "fat" structure.

A different technique for using the dipole antenna over a wide range of frequencies is the well known ham-band Fan Dipole. The technique uses multiple dipoles, cut to various widely spaced bands and fed in parallel at the center. It can be a simple and effective way to cover several ham bands. However, the SWR is generally very high between those bands. In that sense it does not get categorized as a broadband antenna. The similarities and differences with the Fan Dipole of this article are discussed below.

Bandwidth widening by stagger tuning of two inverted **V** antennas was introduced by James Lawson, W2PV, **[2]**. Lawson showed the advantages of two inverted **V** dipoles of slightly different lengths, emphasizing the need for low coupling between dipoles that was achieved by wide physical separation. He demonstrated the ability to use this approach to cover the 13% bandwidth between 3.5 and 4.0 MHz. The "Stagger-Tuned Dipole" **[3]** of Mason Logan, K4MT, added to Lawson's work by analyzing the impedance of two 90° separated dipoles. This used lumped element approximations with the conclusion that good 3.5 to 4.0 MHz coverage was practical.

A variation on the two staggered parallel-fed dipoles is described by Rudy Severns, N6LF, where the second dipole is not attached at the center feed, but is coupled electromagnetically **[4]**. The impedance match is, again, good over 3.5 to 4.0 MHz.

Dave Leeson, W6NL, **[5]** summarizes the possibilities of extending the bandwidth of a single center-fed dipole by use of various networks on the feed line. This includes the transmission line networks pioneered by Frank Witt, AI1H, **[6]**. These again succeed in covering the 3.5 to 4.0 MHz bandwidth with an SWR below 2.0.

The antenna designs covered in this article are extensions of the Lawson and Logan staggered-length dipoles. However, these will use more than two dipoles, arranged into a fan shape, to minimize coupling between the dipoles. Fortunately,

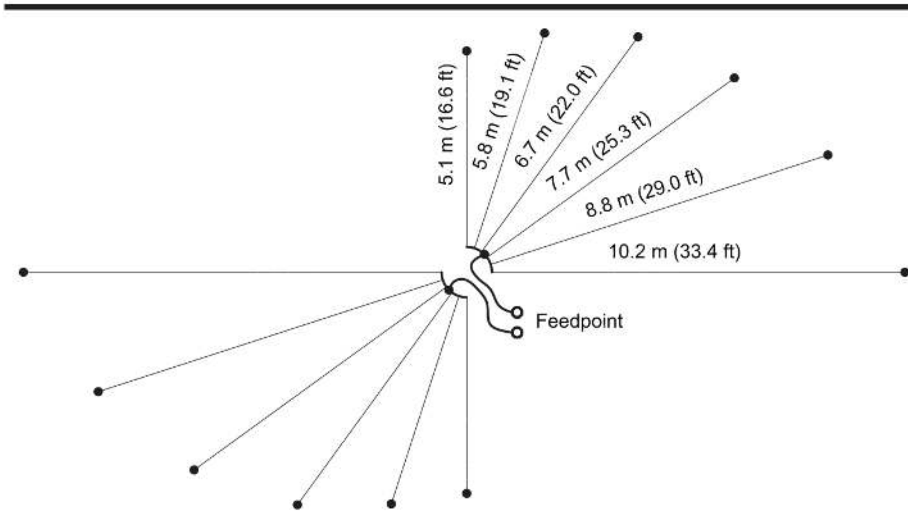

Figure 1 — Layout for the tapered Fan Dipole antenna for 40 m and higher, with wires radiating from the center at 18° intervals. Dimensions for half-dipole wires in m (ft) are shown for a version useful for the 40 m band and higher frequencies.

there is a happy compromise, at about six dipoles, where the coupling is small enough to not produce large impedance spikes between the resonant frequencies. In turn, that allows a reasonable match across a 3:1 frequency range. Once a 3:1 frequency range is achieved, it is further widened since the dipoles all go into a 3/2-wavelength mode. This causes the impedance pattern to roughly repeat at three times the frequency. This odd-multiple trait continues at still higher frequencies, resulting in a useful wideband high-pass response. This is much wider bandwidth than provided by the simple dipole antenna that was the building block.

Multiple Parallel Connected Dipoles

Figure 2 of Lawson's paper **[2]** shows the futility of building a broadband antenna with two tightly coupled dipoles connected at the center. Figure 5 of the same paper suggests that the coupling of the wires remains problematic until the angle between the two dipoles is close to 90°. But, this does not rule out the possibility that an antenna having more than two dipoles could provide a wideband response, even with coupling between the dipoles. I started experimenting with four dipoles operating at about 700 MHz for convenience — the antenna was small enough to hold in my hand. These dipoles were separated into a fan. By trimming the lengths and bending the wires, it was apparent that the widest bandwidth occurred when the total spread of the fan was around 90°. The important discovery was that, if there were enough dipoles, the SWR spikes between dipole resonances were greatly reduced. The SWR response had become truly wideband.

We also know that as the dipoles increase in number, they also couple together more tightly, working to our detriment. Experiments gave a good compromise at six dipoles spaced 90/(6–1) = 18° apart. There may be better combinations, but this seemed like a good starting set of parameters and the ones to explore with *NEC2*.

This geometry is shown in **Figure 1** and the resulting computed SWR is in **Figure 2**. The region from 7 to about 20 MHz uses the wires as half-wave dipoles. The region from 20 to 33 MHz is using the wires as three-half-waves dipoles and so forth for higher odd multiples.

The *NEC2* computed response is remarkably different from the same dipoles in a parallel tightly-coupled configuration (non-fan). Above the resonant frequency of the longest dipole, the SWR does not exceed about 3:1. This contrasts with the sequence of very high SWR regions for the parallel non Fan Dipoles. The center feed-point impedance is higher with the lowest peak SWR corresponding to a feed-point impedance around 175 Ω.

Additionally, the bandwidth of the fan geometry, operating as half-wave dipoles, has increased to almost 3:1. This is important since dipoles will also operate in a 3/2-wavelength mode, tending to cause the SWR at the feed-point to repeat at 3 times the half-wave mode frequencies. To continue this idea, the dipole response repeats at all odd multiples of a half wave and this produces contiguous regions of reasonable impedance match. The final result is that the feed-point SWR has a high-pass like shape and provides very wide bandwidth, with a lower limit at the frequency where the longest dipole is about a half-wavelength.

This all says nothing about the beam pattern that is changing with frequency and will be discussed below. But, an SWR of 3:1 is an interesting level. It is low enough to be used effectively for reception in the HF range. The moderate amounts of attenuation work on both signals and external noise, that dominates internal receiver noise, resulting in almost no degradation in signal-to-noise

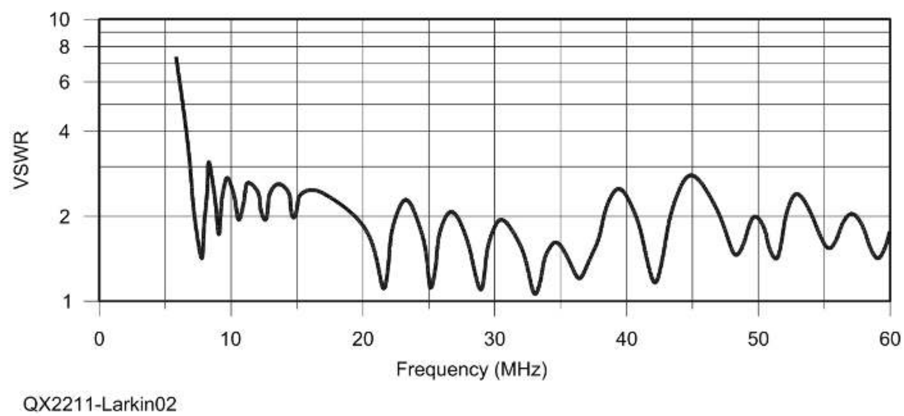

Figure 2 — SWR of the antenna of Figure 1 calculated by *NEC2*. The reference impedance is 175 Ω.

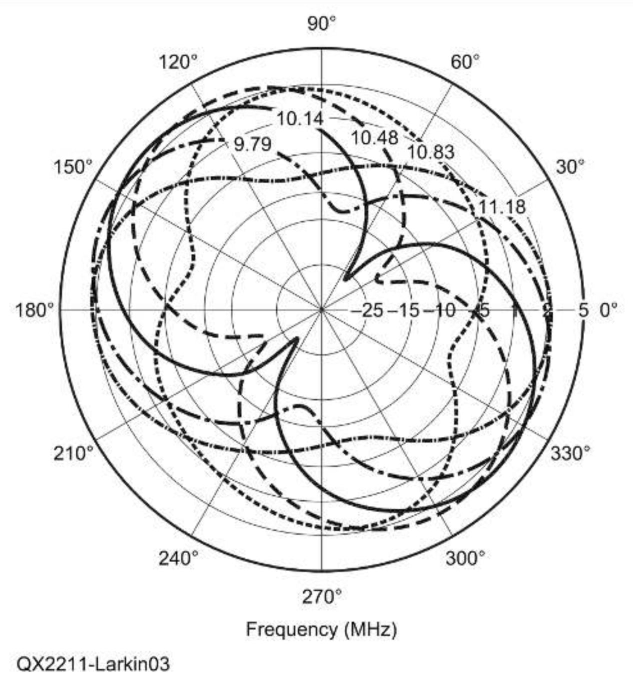

Figure 3 — The *NEC2* calculated horizontal radiation patterns for the antenna of Figure 1 with gain in dBi. The longest dipole lies along the *x*-axis and the shortest along the *y*-axis.

ratio. For transmitting, many transmitters will not produce full output with that much SWR. However, it is in the range where matching is easily done by antenna tuners.

Radiation Patterns

The beam patterns for the multi-wire Fan Dipoles vary considerably with frequency. In general, two or three of the dipoles carry most of the excitation current. Because the lengths of the multiple dipoles are different, these currents are not at all in phase. Since there will be partial cancellation of the fields, the magnitude of these currents will be greater than that seen in a simple dipole. The magnitude and phase of the currents change rapidly with frequency as do the resulting radiation patterns. An example of this is **Figure 3**, showing the azimuth pattern in dBi for 5 frequencies chosen to lie between the second and third dipole resonant frequencies. This is for the six dipole antenna of **Figure 1**. The longest dipole lies along the *x*-axis and the shortest along the *y*-axis. These five patterns cover only the range of 9.79 to 11.18 MHz, which is about the spread between the resonant frequencies of the second and third dipoles. These are only about 0.35 MHz apart but the orientation of the radiation peaks vary considerably with different frequencies. It would be interesting to explore the factors determining radiation pattern but for now we'll settle for the observation that the patterns vary but are generally greatest in the quadrants that are broadside to the six dipoles.

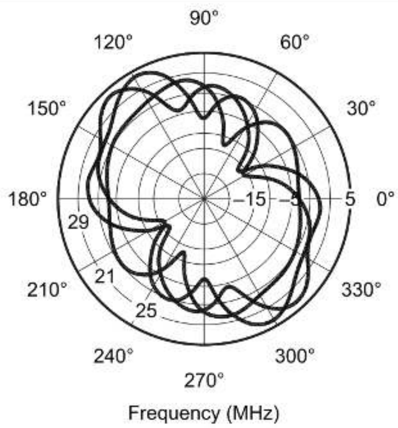

Figure 4 — Horizontal radiation pattern for the antenna of Figure 1 at frequencies where the wires behave as three-half-wave dipoles (20 to 33 MHz).

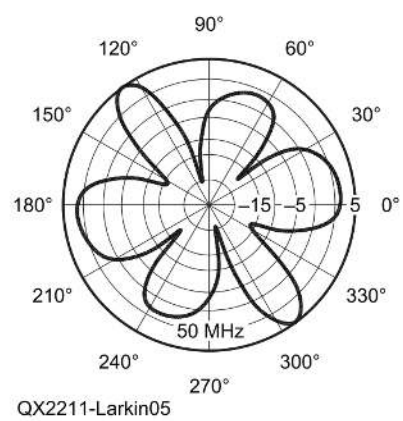

Figure 5 — The horizontal radiation pattern for the antenna of Figure 1 at 50 MHz where the wires behave as 5/2 wavelength dipoles.

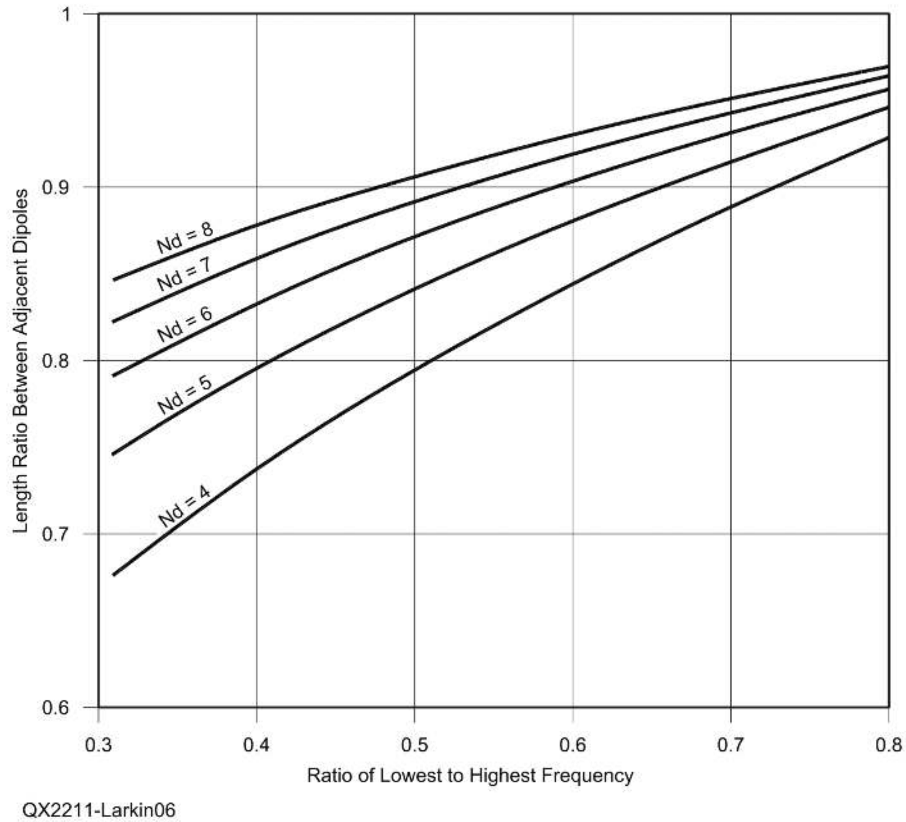

Figure 6 — Length Ratio as a function of the highest/lowest frequency band to be covered, the "Frequency Ratio," and the number of dipoles, *Nd*, being used.

Figure 4, the second of the radiation pattern plots for the same six dipole antenna, covers the 15, 12 and 10 meter bands (21, 25 and 29 MHz) where the wires are operating in a 3/2-wave dipole mode. This results in a more complex structure of radiation lobes. Even more complex is the 6-meter (50 MHz) pattern of **Figure 5** that is operating in the 5/2 wave dipole mode. As a general purpose antenna these have useful patterns even though it would be difficult to put lobes in specific directions. Most stations do not have control of the location of support trees, either! Note that the gain reaches 5 dBi in certain directions.

The tapered fan structure does not behave as a group of independent dipoles as the frequency is shifted. Some current flows in non-resonant dipoles and this alters the radiation pattern from that of a simple dipole. This can be beneficial, as it prevents the dipole peak pattern from rotating all the way through 90°. In addition, the end nulls we associate with dipole radiation are often not as deep as they are for simple dipoles.

Dipole lengths

The ratio of the shortest to longest lengths, for the Fan Dipole of **Figure 1**, was set to 0.5. With six dipoles this yielded the useful broad-band SWR curve that we saw in **Figure 2**. Other lengths and numbers of dipoles could yield other useful results, and experimenting with these parameters is encouraged. The lengths of the successive dipoles in my experimentation have been geometric progressions, meaning that each dipole has a length equal to the previous dipole's length multiplied by a "Length

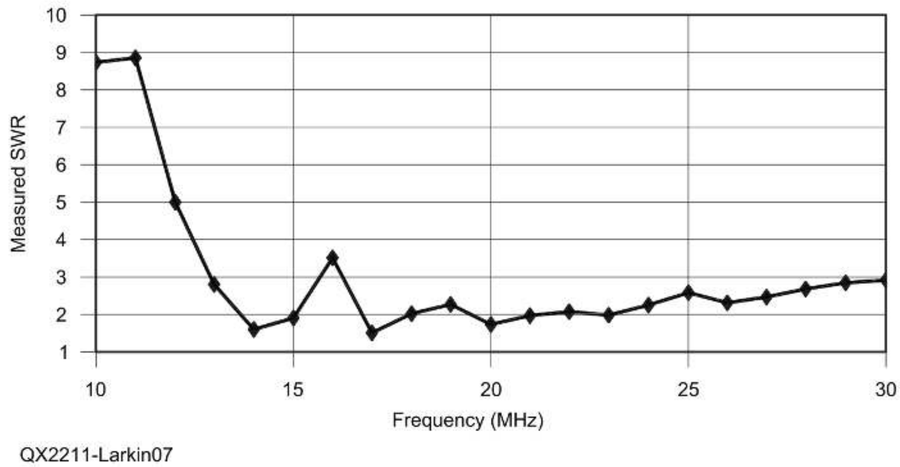

Figure 7 — SWR of the Tapered Fan Dipole described in the Construction section.

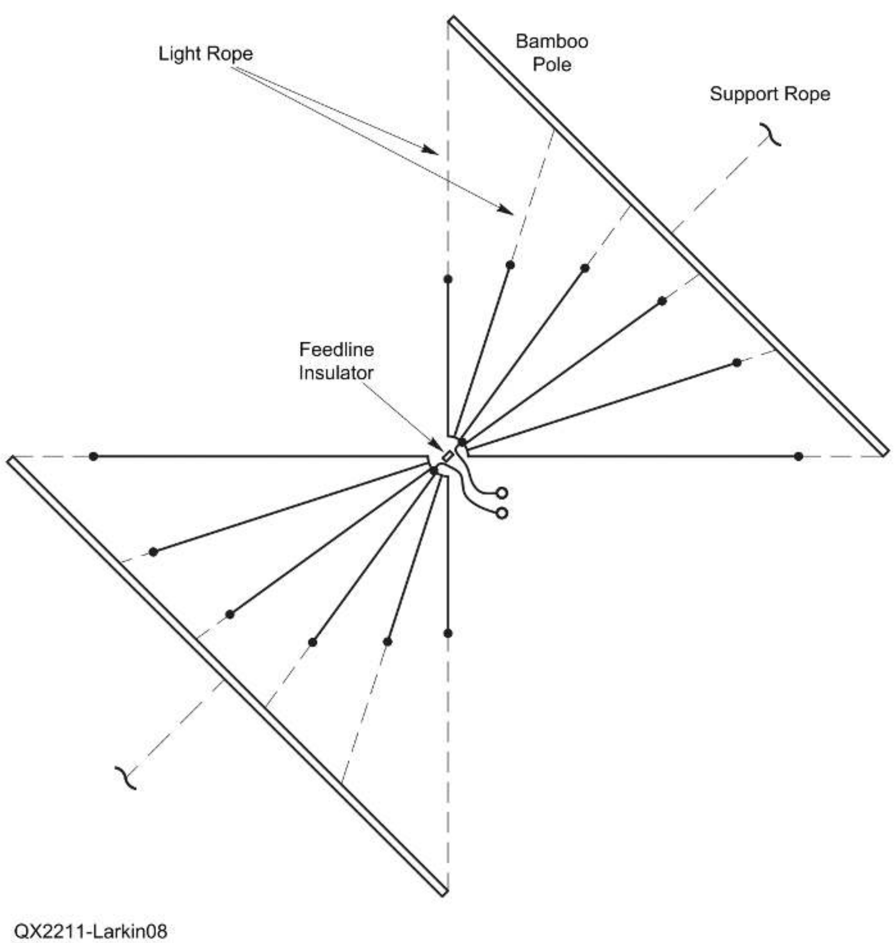

Figure 8 — Construction of the Tapered Fan Dipole supported by two halyard lines in trees. Bamboo poles were used for the end connections. Extra lines (not shown) keep the fan horizontal.

Ratio" (*LR*) constant. For the example antenna, the Length Ratio was 0.87. The half-lengths of the six dipoles in meters (feet) are 10.2 (33.4), 8.8 (29.0), 7.7 (25.3), 6.7 (22.0), 5.8 (19.1), and 5.1 (16.6). These lengths are based on a starting point for the longest dipole at the classic half-wave 468/*f* formula, where *f* is the frequency in MHz and the result is in feet. For the tapered Fan Dipole, this results in a SWR at the lowest frequency of about 3.0.

The Length Ratio can be found from $LR = \left(f_2 / f_1 \right)^{1/(Nd-1)}$, where f_1 and f_2 are the frequencies corresponding to the longest and shortest dipole lengths for the fan and *Nd* is the number of dipoles. **Figure 6** shows the Length Ratio for different values of Frequency Ratio, f_2/f_1, and number of dipoles, *Nd*, from 4 to 8.

The Frequency Ratio is nominally set to 0.5. There are two effects to be worked with. If the Frequency Ratio is too close to 1.0, say 0.8, the antenna behaves more like a fat dipole and has a single resonant frequency. If the ratio is too low, say 0.35, there will be high SWR at frequencies between each pair of low SWR frequencies, without any increase in total bandwidth. It is worth experimenting with values around 0.5 to find the best answer, depending on the application.

Construction

A Fan Dipole was constructed with the lowest frequency set to about 12 MHz. The longest half dipole, measured from the center to the end of each wire, was 5.8 meters and the Length Ratio was 0.85. This is shorter than the version of **Figure 1**, but fit the space that I had available. For convenience, this antenna was fed directly with 50 Ω coax and measured in a 50 Ω system. This was a compromise impedance level, but the highest observed SWR was 3.5 at around 16 MHz as seen in **Figure 7**.

The antenna dipoles were horizontal and supported between a pair of trees at about 8 m (26 ft). The #18 AWG dipole wires were stretched between bamboo poles. The mechanical arrangement for this is shown in **Figure 8**. It all worked fine for me since I had a clear area available that was large enough to assemble the wires on the ground. With bending of the bamboo, it was most helpful to have the entire antenna assembled a few feet above the ground.

For VHF and above, rods and wires are easily center supported. Somewhere above a few hundred MHz, wires can be simply soldered to a center support. To test this idea,

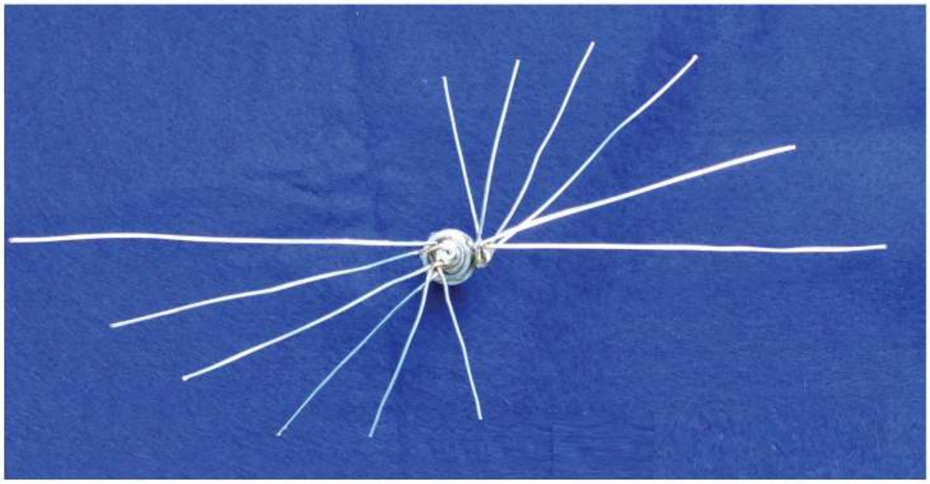

Figure 9 — 550 MHz (lowest frequency) Tapered Fan Dipole built on a BNC panel connector is well suited to experimentation.

a six tapered-length dipole fan was built on a BNC connector (**Figure 9**). The longest half dipole had a length of about 105 mm (4.13 in) and the remainder of the lengths corresponded to a multiplying ratio of 0.85. This was fed directly to a 50 Ω system where the highest SWR observed up through 2000 MHz was 3.6. The lowest frequency for this level of SWR was 550 MHz.

Comparison with Ham-Band Fan Dipoles

The notion of using multiple dipoles, cut to the centers of various bands and connected together at the center, goes back at least to 1937 **[7]** and has been widely used and described in literature **[8]**. Those antennas use the wide separation of ham bands, along with the impedance characteristics of the center-fed dipole, to ensure that only a single dipole is being used at once. This means that at the frequencies of the ham bands, significant current is flowing in only one of the dipoles at a time. This can work well for operation on the individual bands but at in-between frequencies the SWR is very high.

In contrast, for the tapered Fan Dipole, the dipoles do not operate individually since they are too close in resonant frequencies. The resulting current patterns, described under "Radiation Patterns," result in an SWR curve that does not have high peaks.

Extensions

One way of making the tapered-length Fan Dipole easier to construct is to use an inverted **V**. arrangement with a single

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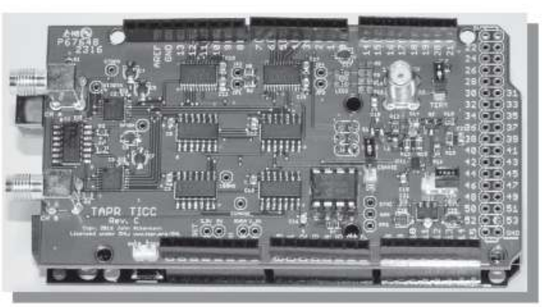

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center support. This was the method used by Lawson, W2PV, for two dipoles **[2]**. The amount of coupling between dipoles is different than for the horizontal fan. This may cause a change in the best number of dipoles and the Length Ratio.

Another interesting possibility is a half fan over a ground plane. Most likely vertical polarization would peak over the frequency range when the longest element was about 45° from vertical. The radiation pattern would have a few quirks. It might be an alternative to the classic discone for some uses.

The horizontal fan as described in this article could be stacked vertically to achieve some power gain along with narrowing of the vertical pattern. That arrangement would be particularly attractive at higher frequencies where the mechanical issues are easy.

Conclusions

An arrangement of multiple dipoles, all fed in parallel at the center, and tapered in length can produce a reasonable impedance match over a very wide range of frequencies. The dipoles are arranged in a fan pattern to minimize adjacent dipole coupling. Over a 3:1 frequency range, the radiation pattern is similar to the simple dipole radiator but varying in direction depending on the frequency being used. As the range of frequencies is extended the pattern develops lobes that can be good or bad depending on the application.

Bob Larkin, W7PUA, has been active in amateur radio since he was first licensed in 1951 as WN7PUA. He received a BSEE from the University of Washington and an MSEE from New York University. He is retired from a career in electronic circuit and system design along with work in analog and digital signal processing. Amateur radio interests include radio design and microwave operation. Other activities include boat building, sailing, hiking and camping.

Notes

[1] J. Pawlowski, SP3L, "Collection of Broadband HF Antenna Designs, Parts 1 and 2." *QEX* Mar./Apr. 2020, pp. 16-22 and May/June 2020 pp. 28-35.

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[7] F. Weyerhaeuser, W9YPQ, "Three-Band Automatic Antenna, " *QST*, June 1937, p 54. This seems to be the origin of the multi-ham-band fan dipole; uses 3 dipoles to cover the 40, 20 and 10 m bands.

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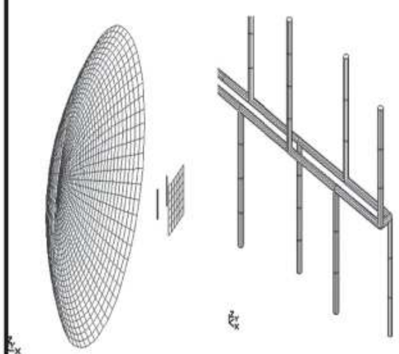

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Self-Paced Essays — #14 Quality Factor Q

Q is the ratio of stored energy to energy lost in one RF cycle.

In Essay 9, Angular Frequency, (*QEX* Jan./Feb. 2022), we gave a brief introduction to *dimensionless* or *unit-less* numbers. Q is one such number, and is of great importance in radio engineering. It also has significant meaning in mechanical systems, as well, though sometimes designated by different terms, such as *damping factor,* which is actually the reciprocal of Q.

Conceptually, quality factor Q is the ratio of energy stored to energy dissipated in a system. In an electrical system, especially one in which resonance is in play, it is a measure of how long a system will oscillate. The higher the Q is, the longer a system will "ring," whether that system consists of an actual bell, or a radio "tank circuit."

We will modify the 5.033 MHz resonant circuit that has served us so well (using all the same component values as in Essays 12 and 13), and shock excite it with a short pulse (**Figure 1**) and see how it responds.

We have used a value of 1 Ω for R1, which will give us a value of 31.6 for Q.

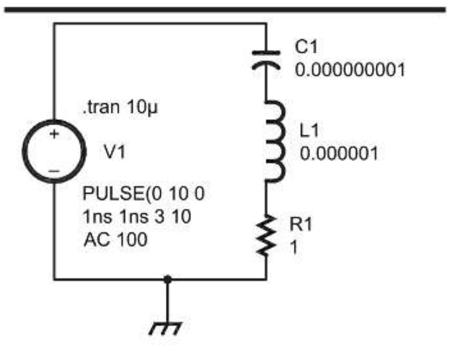

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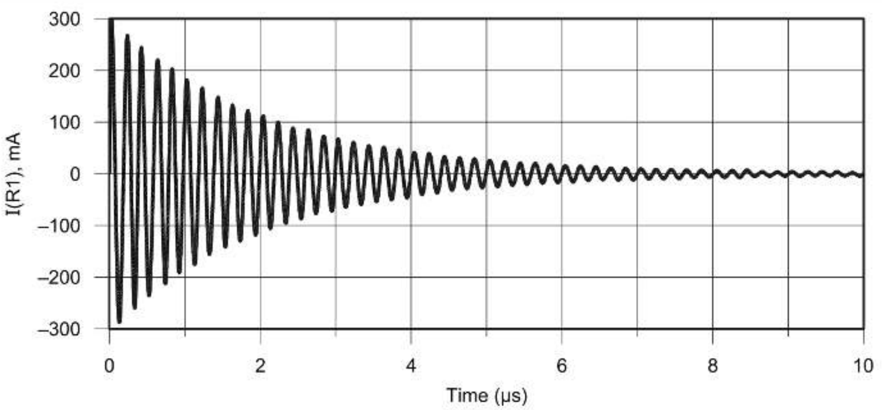
Figure 1 — Pulse excited series resonant circuit with a Q of 10.



QX2211-Nichols02

Figure 2 — Response of the pulse excited series resonant circuit having a Q of 10.

Let's see what the response looks like (**Figure 2**).

Here we have a very nice *damped oscillation.* Notice that the *envelope* of this curve is a perfect exponential curve, just like an *RC* time constant. Also, like an *RC* time constant, the wave never decays to zero, but continues on indefinitely. The *frequency* of the wave is exactly the same as it would be if it were continually excited. You can count 5 complete cycles every microsecond, giving a period of 0.2 microseconds.

We can change the time the circuit "rings" by changing the Q, and we can do that by two methods. Either change the value of R1, or change the ratio of L1 to C1. Let's increase L1 by a factor of ten and decrease C1 by a factor of ten (**Figure 3**).

We now have a Q of 316, which is about the greatest Q you can achieve using practical air core inductors and capacitors at HF, see **Figure 4**. As you can see, the oscillation has taken a lot longer to decay. Although the decay looks almost linear in this display, it indeed follows the same exponential curve as in the previous example.

There's another subtle point that we can make here, which we will explore in more detail as we move into Fourier Analysis. A damped oscillation, as smooth as the oscillation envelope seems to be, does *not* consist of an absolutely pure sine wave. This damped oscillation is the product of a decaying exponential and a pure sine wave. It is impossible to change the amplitude of a sine wave without changing its *shape,* no matter how subtly. This wave actually has sidebands that are very weak and close in, but they are there nevertheless. Only a wave of absolutely constant amplitude can produce a pure sine wave, or more succinctly; only a sine wave of absolute constant amplitude is a sine wave! As we look at modulation later on, we will analyze

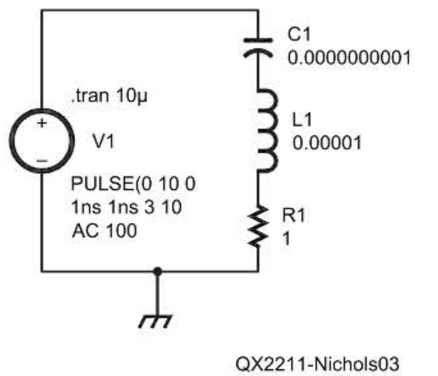


Figure 3 — Pulse excited series resonant circuit with a Q of 316.

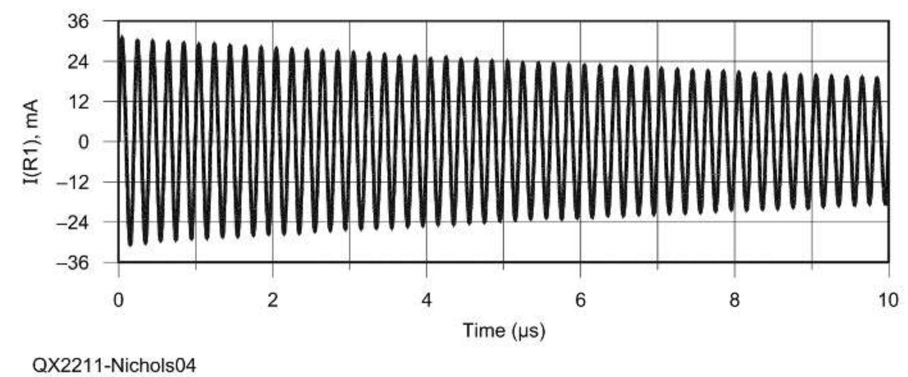


Figure 4 — Response of the pulse excited series resonant circuit having a Q of 316.

this in detail.

The Q of a tuned circuit not only has a tremendous effect on the efficiency of a tank circuit, such as in a transmitter output stage, but is of primary importance in frequency selectivity in receivers and other devices, where efficiency is of little consequence

.

Q and Bandwidth

One of the chief tasks a radio receiver has to perform is to select a desired frequency (or narrow band of frequencies) and reject all others. While modern digital signal processing allows modern receivers to achieve effective selectivity that is far less at the mercy of the Q of tuned circuits, highly selective high-Q tuned circuits are not going to go away any time in the foreseeable future.

The relationship between Q and bandwidth of a tuned circuit is simple:

$$BW = \frac{f}{Q_U}$$

where *BW* is the *half-power* bandwidth of the circuit, *f* is the resonant frequency, and Q_U is the *unloaded* Q of the circuit, which is the "raw" Q of the circuit not connected to anything. The unloaded Q is not dependent on any external circuitry that the tuned circuit may be powering.

If one tunes a circuit precisely to resonance, and then moves off the resonant frequency in either direction, there will be a frequency at which the power consumed by the resistive component (R1 in our previous circuits) drops 3 dB down, or to one half. The upper and lower half power points f_U and f_L are symmetrically located above and below the resonant frequency when Q is plotted against frequency on a logarithmic scale. The resonant frequency is the square root of the product $f_U f_L$. At lower Qs, the response can appear to be quite asymmetrical when plotted on a linear frequency scale. Since voltage is much easier to measure than power, it is standard practice to measure the voltage across the resistance, where a decrease by 0.707 of the center frequency value represents half power. Remember we said that number 0.707 (the square root of a half) will keep popping up. Here is just one example.

By the way, R1 might or might not be an actual physical resistor; it can be any *real* resistive component, such as the radiation resistance of an antenna. Much more on this later.

Useful Losses

We have perhaps unjustly made the assumption, based on the term "quality factor," that circuits with higher Q are inherently "better" than circuits with lower Q. We shouldn't forget that real work is only produced in a "real" component, not a reactive one. A tuned circuit with infinite Q is incapable of producing any useful work. Any *useful* circuit must have a Q of something less than infinity — sometimes a *lot* less. The entire purpose of an antenna, for instance, is to "lose" all the applied RF energy into free space.

Reciprocating

There's a numerical quantity that isn't much used in radio work, but is quite prominent in power generating circles: Dissipation Factor (*D*). *D* is simply the reciprocal of Q. Old school capacitor testers measured dissipation factor; more modern instruments are more likely to characterize a capacitor by its *ESR*, or Equivalent Series Resistance. This is essentially the same as Dissipation factor and equals ESR/X_C, if you include the actual capacitive reactance value. When it comes to capacitors, a high Q (or low *D*) *is* always a desirable thing. You only want a capacitor to store energy, not to get hot. One of the famous capacitor trademarks of past ages was "Vitamin-Q" for obvious reasons. See **https://en.wikipedia.org/wiki/Dissipation_factor**.

Some Notes on Inductors

From a purely practical standpoint and at radio frequencies, the Q of most tuned circuits is primarily limited by the inductor. Inductors are made of wire, sometimes lots of it, and wire always has some resistance. Fortunately, the resistance of a coil of wire in an inductor increases linearly with length, while the inductance of a uniform inductor increases as the *square* of the number of turns, or total length of wire, and in turn, its reactance. What this means is that, up to a limit, higher inductance inductors will have higher Q than lower inductance ones, all other things being equal.

If one looks at the entire history of radio, one will find an extreme amount of experimentation and craftsmanship dedicated the construction of high Q inductors. One of my mentors once wryly stated that radio is nothing more than winding coils. Looking over the past hundred years of amateur radio experimentation, one can see the truth in this statement. Great effort was expended to devise coils that reduced as much as possible the inter-winding capacitance, which is always present in any practical inductor. The ancient and venerable *RCA Radiotron Designers's Manual* dedicates over one hundred pages to the construction of coils. So it must be important.

In the next installment, we will consider filters of all kinds. Filter design was one of the most tedious exercises in all of electronics; now with *SPICE* modeling (and *SPICE*-based derivatives like *ELSIE*), even the most complicated filter design can be carried out by mere amateurs.— *Stay tuned! Eric.*

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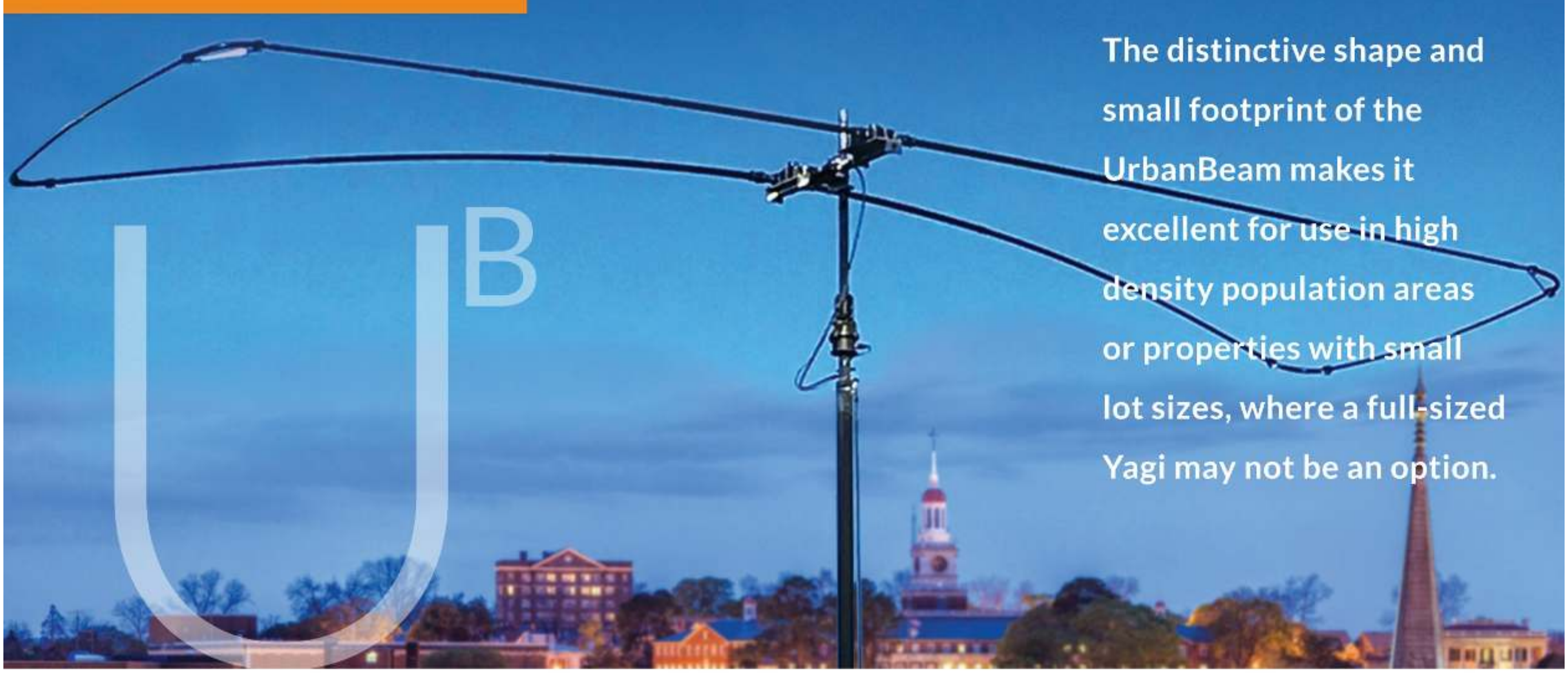

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