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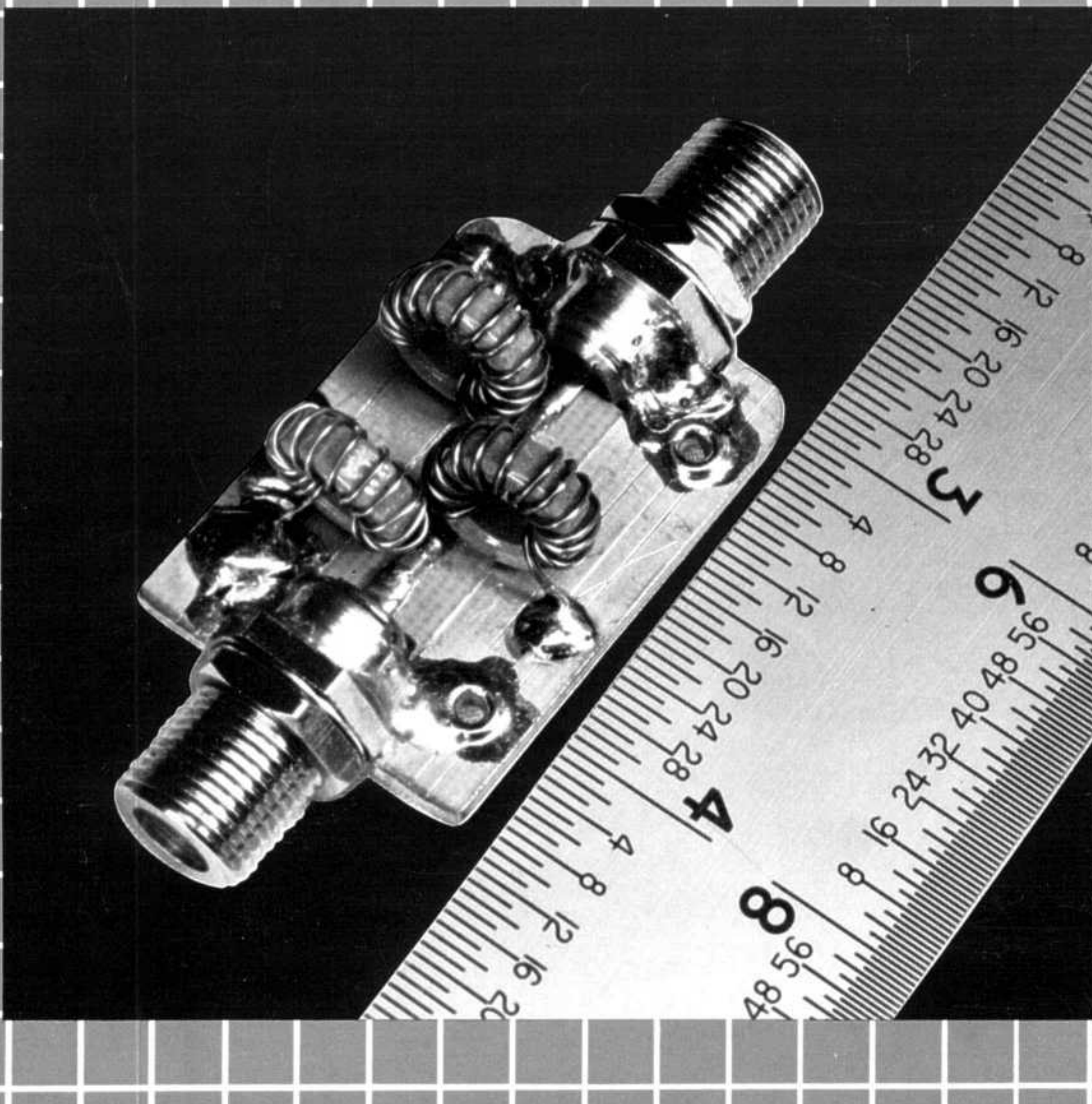

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ABOUT THE COVER

This high-pass filter was constructed using microstrip techniques and hand tools. The finished product fits into a small 1.1 × 1.4 inch enclosure. Ceramic-chip capacitors were employed in its design. For a look at the filter's performance characteristics, turn to page 7.

Purposes of *QEX*:

1) provide a medium for the exchange of ideas and information between Amateur Radio experimenters
2) document advanced technical work in the Amateur Radio Field
3) support efforts to advance the state of the Amateur Radio art.

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Both theoretical and practical technical articles are welcomed. Manuscripts should be typed and double spaced. Please use the standard ARRL abbreviations found in the 1985 and 1986 ARRL Handbooks and in the January 1984 issue of *QST*. Photos should be glossy, black-and-white positive prints of good definition and contrast, and should be the same size or larger than the size that is to appear in *QEX*.

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Empirically Speaking. . .

Sunspot Cycle 22

In 1848, a method was introduced for the daily measurement of sunspot numbers. That method, which is still used today, was devised by the Swiss astronomer, Johann Rudolph Wolf. The observer counts the total number of spots visible on the face of the sun and the number of groups into which they are clustered, because neither quantity alone provides a satisfactory measure of sunspot activity. The observer's sunspot number for that day is computed by multiplying the number of groups he sees by ten, and then adding to this value the number of individual spots.

As can readily be understood, results from one observer to another can vary greatly. Measurement depends on the capability of the equipment in use, on the stability of the earth's atmosphere at the time of observation, and on the experience of the observer. A number of observatories around the world cooperate in measuring solar activity. A weighted average of the data is used to determine the International Sunspot Number or ISN for each day. (Amateur astronomers can approximate the determination of ISN values by multiplying their values by correction factors determined empirically.)

Sunspot records were formerly kept in Zurich, Switzerland, and the values were known as Zurich sunspot numbers. They were also known as Wolf sunspot numbers. International records are now compiled and kept at the Sunspot Index Data Center, 3 avenue Circulaire, B 1180 Bruxelles, Belgium.

In more recent years an additional method of determining solar activity has been put to use—the measurement of solar radio flux. The quiet sun emits radio energy across a broad frequency spectrum, with a slowly varying intensity. Solar flux is a measure of energy received per unit time, per unit area, per unit frequency interval. These radio fluxes, which originate from atmospheric layers high in the sun's chromosphere and low in its corona, change gradually from day to day, in response to the number and size of spot groups on the solar disk. Thus, there is a close correlation between solar flux values and sunspot numbers. One solar flux unit equals the factor 10 to the power -22 Joules per second per square meter per hertz. Solar flux values are measured daily at 2800 MHz (10.7 cm) at the Algonquin Radio Observatory in Ottawa, Ontario, where daily data are available since February, 1947. Measurements are also made at other observatories around the world, at several frequencies. With some variation, the measured flux values increase with increasing frequency of measurement, at least up to 15.4 GHz. Sunspot numbers on a given day do not relate directly to maximum usable frequency. Solar flux values are more reliable for this purpose.

In the United States, solar data records are maintained at the National Geophysical Data Center, Solar-Terrestrial Physics Division (E/GC2), 325 Broadway, Boulder, CO 80303. Monthly solar indices bulletins are available on a subscription basis ($20 annually), containing data on both sunspot numbers and solar flux. (Much of the above historical information was extracted from these bulletins.) The bulletins also contain predictions of sunspot numbers for the coming 12 months.

The article, "Spots Before Your Eyes," appearing on pages 34 and 35 of May 1986 *QST*, gives a brief history of sunspot cycles and indicates that we'll likely be entering Sunspot Cycle 22 by mid-1987. However, it is possible that we have *already* entered Cycle 22. Traditionally, a new cycle begins when the 12-month smoothed ISN has "bottomed out" and begins to increase.

From current data, the lowest smoothed monthly values since Cycle 21 began in 1976 were 13 for both February and March, 1986. April and May were higher, with a value of 14 for each month. The results for later months are not yet in, because smoothed monthly sunspot numbers must include data for 6 months following the month being considered. A lengthy period of solar calm in November, December and into January may cause smoothed monthly values subsequent to May to dip again. Even so, W1AW's Propagation Forecast Bulletin 51 of December 22, 1986, written by Ed Tilton, W1HDQ, states that new cycle activity has appeared several times recently. Quoting Ed from another bulletin, "We are at a point in the solar cycle where new activity can break out at almost any time with little or no warning." Cycle 22, here we come!—Gerald L. Hall, K1TD, Deputy Manager ARRL Technical Dept

Analyzing Simple Matching Networks

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Impedance matching networks are among the most useful and most written about ac circuits known. Many systems which generate or process RF energy use matching networks like the pi or T to obtain efficient power transfer from one link of the circuit chain to the next. The criterion for maximum power transfer is simple: The input impedance of one circuit link *must* match (represent the complex conjugate of) the load impedance of the *upstream* link, from which it is drawing power. If the two links aren't matched, a simple impedance transformer or matching circuit can be inserted between the links and *tuned* to transform the input impedance of the downstream link to the complex conjugate of the load impedance of the link feeding it. Fig 1 illustrates the matching process. Details can be found in any book on basic ac circuits.

If so much has been written about matching networks, why this article? Most of the amateur literature offers graphical solutions of matching networks or mathematical details concerning specific matching networks. A fundamental property of these networks—their ability to attenuate out-of-passband energy—is often passed over or only lightly brushed. Straightforward ac circuit analysis can open the door to the characterization of both the input impedance and the attenuation of a matching network at any desired frequency.

The necessary tools for this characterization are complex numbers and the concept of series and parallel equivalent circuits. Both tools have recently been described in *QST* articles and are recommended background reading.[1,2]

Equivalent Circuits

The concept of equivalent circuits states that any circuit branch containing resistance and reactance in series can be replaced by an electrically equivalent branch containing conductance and susceptance in parallel. The complex admittance of the parallel-equivalent branch, Y, is the reciprocal of the impedance, Z, of the series branch. That is, $Y = Z^{-1}$. The converse (replacing a parallel branch with a series-equivalent branch) is equally valid.

[1,2]Notes appear on page 13.

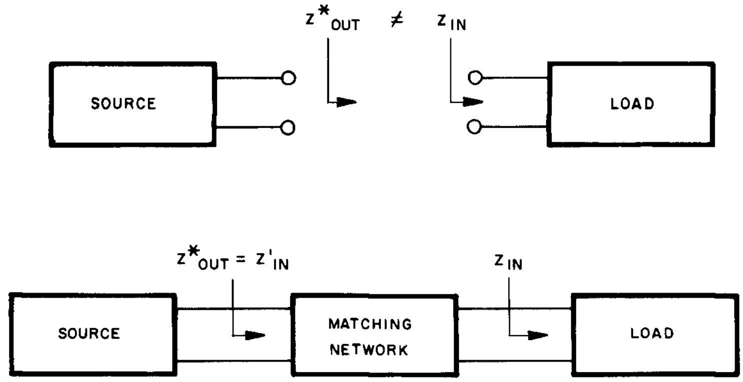


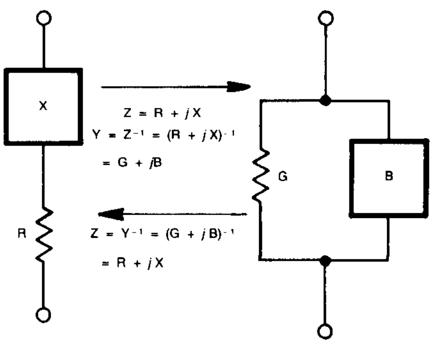
Fig 1—A matching network impedance matches a source of RF energy to a load. Z_{in} and Z'_{in} are the natural and transformed input impedances of the load and Z^*_{out} is the complex conjugate of the output impedance of the source.

Let's illustrate these principles with the example in Fig 2. There are several ways to accomplish the transformation. I prefer the coordinate conversion approach outlined by Schetgen.[3] If we use actual components—resistors and coils in this example—to construct the two circuits, the equivalence will hold only at one frequency. One (or both) of the equivalent circuits would have to be *retuned* with different values of resistance and inductance to reestablish electrical equivalence at a new frequency. A final note on equivalent circuits: If the reactive element in the original circuit is inductive, the reactance in the equivalent circuit will also be inductive. The same holds true for capacitive elements.

Admittance, conductance and susceptance convey a certain fear in most amateurs. The familiar *product over the sums* approach described by Napurano is used most often in circuit analysis.[3] Once you become familiar with the concept of admittance, however, you'll abandon the other, more cumbersome method!

A Pi Network Circuit

Now that we've reviewed the fundamentals, let's tackle a practical example such as the pi-network tank of a class-AB tetrode amplifier with a plate-load resistance of 2700 ohms (Fig 3). From a circuit analysis viewpoint, the tank circuit's function is to step up the 50-ohm antenna resistance to the 2700-ohm plate-load resistance of the tube. This impedance transformation allows for maximum undistorted power transfer to the antenna within the ratings of the tube.



If R = 50 ohms, X = +110 ohms (inductive)

then $Z = 50 + j110$ ohms

$Y = Z^{-1} = (50 + j110)^{-1} = \frac{1}{50 + j110}$ S

Converting from rectangular to polar form,

$Z = \sqrt{50^2 + 110^2} = 120.83$ ohms

$\theta = \tan^{-1} 110/50 = 65.556°$

$R + jX = Z\angle\theta = 120.8\angle 65.556°$ S

$Y = \frac{1\angle 0}{120.8\angle 65.556°} = 8.276 \times 10^{-3}\angle -65.556°$ S

Converting from polar to rectangular form,

$\cos(-65.56°) = 0.4138$

$\sin(-65.56°) = -0.9104$

$G = Y\cos\theta = 3.425 \times 10^{-3}$ S

$B = Y\sin\theta = -7.534 \times 10^{-3}$ S

$Y = G + jB = 3.425 \times 10^{-3} - j7.534 \times 10^{-3}$ S

Fig 2—Equivalence of a series R-X and a parallel G-B network branch.

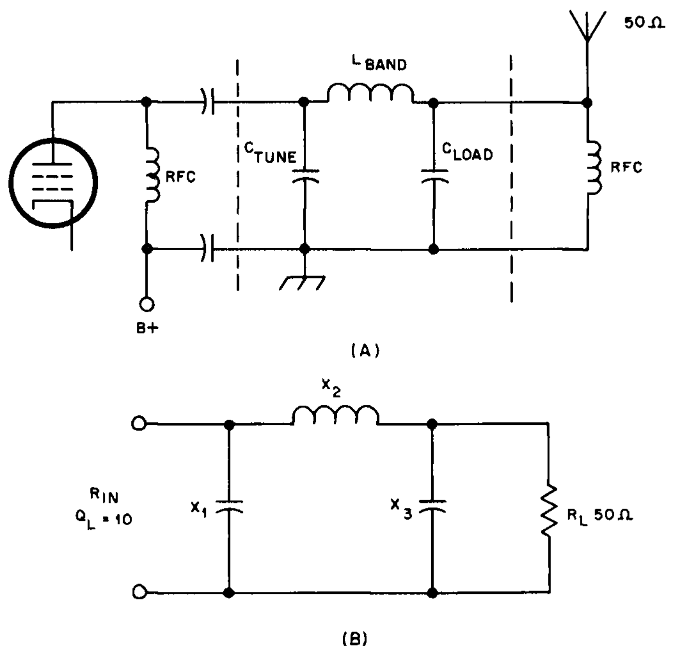


Fig 3—A shows the output circuit of a class-AB amplifier and B depicts a pi network isolated from the rest of the circuit.

A typical loaded or circuit Q for this application is 10.

Let's calculate the second harmonic attenuation of the plate tank circuit in Fig 3. The design formulas, adapted from Hayward and DeMaw's *Solid State Design*[5], will help us determine the reactances at the operating frequency.

$$X_1 = \frac{R_{in}}{Q} = \frac{2700}{10} = 270.0 \text{ or}$$

$$-270.0 \text{ ohms (capacitive)} \qquad \text{(Eq 1)}$$

$$X_3 = R_L \left[\frac{R_{in}/R_L}{Q_2 + 1 - R_{in}/R_L} \right]^{1/2}$$

$$= 50 \left[\frac{2700/50}{10^2 + 1 - 2700/50} \right]^{1/2}$$

$$= 53.59 \text{ or } -53.59 \text{ ohms (capacitive)} \qquad \text{(Eq 2)}$$

$$X_2 = \frac{Q \times R_{in} + R_L \times R_{in}/X_3}{Q^2 + 1}$$

$$= \frac{(10)(2700) + (50)(2700)/(53.59)}{10^2 + 1}$$

$$= 292.3 \text{ ohms (inductive)} \qquad \text{(Eq 3)}$$

The next step is to find the input impedance of the pi network at the second harmonic. (It isn't 2700 ohms!) Here is where network reduction comes in. Circuit elements are combined to reduce, stepwise, the total number of elements. Reactances or resistances in series can be combined by addition to yield a single net reactance or resistance. The same addition rule holds for conductances and susceptances in parallel.

The steps in the reduction process are shown in Figs 4A-G. The goal of the first equivalence transformation (Fig 4C) is to place the equivalent reactance X_a in series with X_2. The network is then simplified by the addition of X_a and X_2 to give the net reactance X_b (Fig 4D). The second equivalent conversion places the susceptance B_c in parallel with B_1. These two susceptances add to give the simplest parallel equivalent circuit in Fig 4F. A final transformation converts this circuit to the simplest series equivalent circuit, Fig 4G. This equivalent circuit has the same input impedance and phase shift at a given frequency as the original pi network.

Our ultimate objective is to arrive at the value of the second harmonic attenuation—the attenuation at twice the operating frequency. Since capacitive reactance is inversely proportional to frequency, X_1 and X_3 of our pi network are halved at the second harmonic. X_2 is directly proportional to frequency and is, therefore, doubled (Fig 4A). The subsequent arithmetic in Figs 4B-G is guided by the network reduction to give us a second harmonic input impedance of 1.106 – *j*177.5 ohms, practically a short circuit to the second harmonic current flowing in the amplifier.

Attenuation

How do we go from input impedance to attenuation? We use the following formula (derived in the Appendix):

$$A(\text{dB}) = 10 \log (4i^2 \times R_{in} \times R_s) \qquad \text{(Eq 4)}$$

where
- $A(\text{dB})$ = total attenuation in decibels
- i = magnitude of the complex current flowing into the network
- R_{in} = input resistance of the network (the real part of Z_{in})
- R_s = source resistance

An explanation of the decibel has recently appeared in *QST*.[6]

Z_{in} and Z_s are sometimes confused in the amateur literature. Z_{in} is the input impedance of the network—1.106 – *j*177.5 ohms at the second harmonic in our example. Z_s, on the other hand, is the internal impedance the network *sees* looking back into the source (Fig 4H). If $Z_{in} = Z^*_s$ at the fundamental or operating frequency, the network is said to be *doubly loaded* or *doubly terminated*. Both the input and the output terminals see the complex conjugates of the design impedances. In practice, however, matching networks are seldom doubly terminated.[7] In our case, a class-AB tube amplifier has a very high source impedance. Since it is a nearly *open* source, we'll estimate $Z_s = 1 \times 10^9 + j0$ ohms for our attenuation calculation. A more accurate estimate of Z_s would divert us into the theory of nonlinear devices and greatly complicate the calculation. Fortunately, the attenuation of the plate tank circuit doesn't depend strongly on the source impedance. A ballpark figure will do nicely.

We now have the two quantities, R_s and R_{in}, we need for our attenuation calculation. What about the third quantity, the current? From Ohm's Law, the complex current, ic, flowing into the network is

$$i_c = \frac{V_s}{Z_s + Z_{in}} \qquad \text{(Eq 5)}$$

In a (relative) attenuation calculation, the voltage source can be set for any desired voltage. Setting V_s = 1 V simplifies Eq 5 to

$$i_c = \frac{1}{Z_s + Z_{in}} \qquad \text{(Eq 6)}$$

The magnitude of the current is the quantity Eq 4 calls for. Since $i_c = i_{re} + ji_{im}$,

$$i = \sqrt{i^2_{re} + i^2_{im}} \qquad \text{(Eq 7)}$$

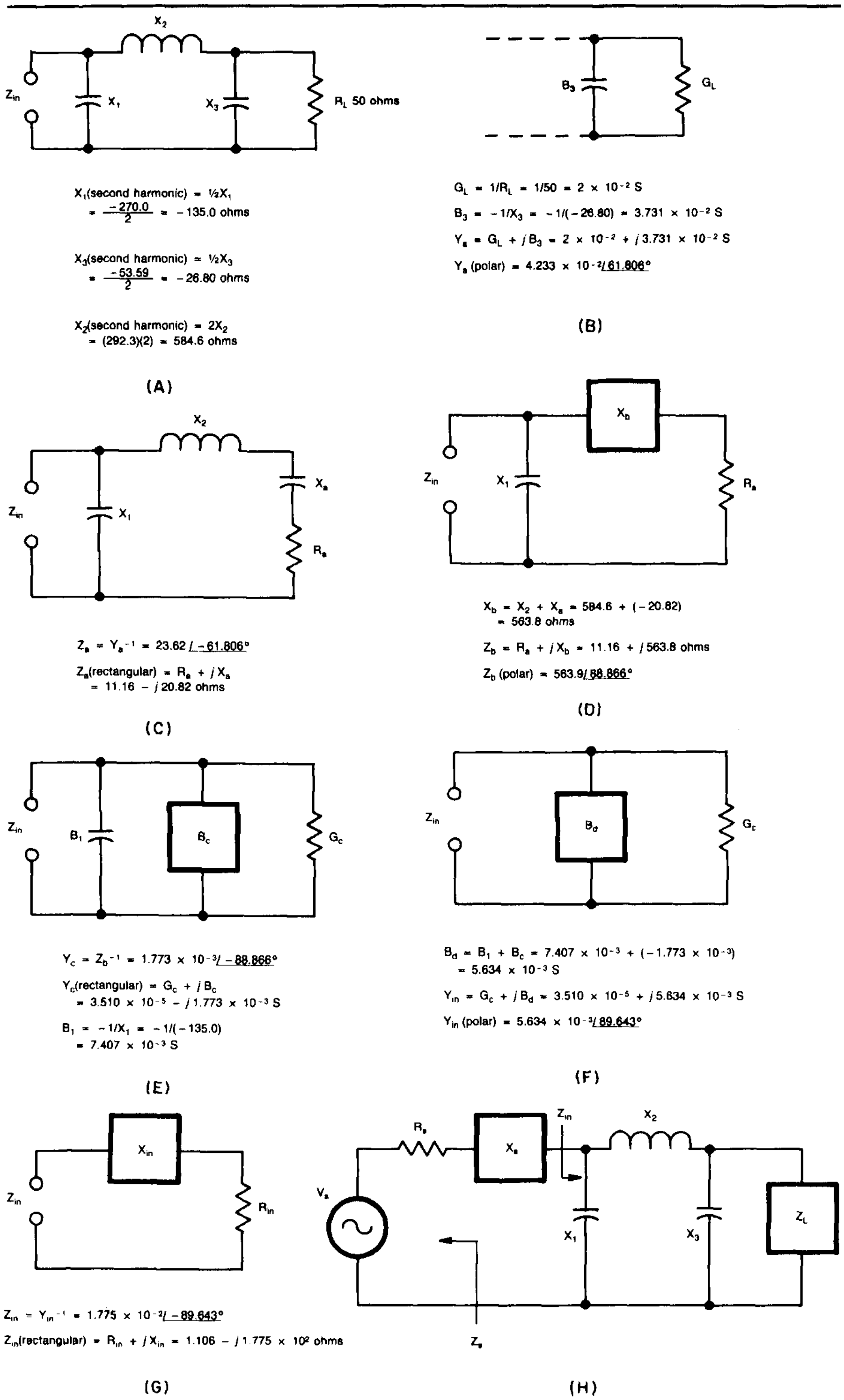


Fig 4—Analysis of the 2700 to 50-ohm pi network. Figs A through G show the stepwise network reduction with calculation of the input impedance Z_{in} at the second harmonic. The relation between the input impedance and the source impedance Z_s is shown in H.

where

i_{re} = the real part of i_c
i_{im} = the imaginary part of i_c

Now let's calculate the second harmonic attenuation of our output tank. From Eq 6,

$$i_c = \frac{1}{(1 \times 10^9 + j0) + (1.106 - j177.5)}$$

$$= \frac{1}{1.000 \times 10^9 - j177.5}$$

$$= 1.000 \times 10^{-9}\ \text{A} \qquad \text{(Eq 8)}$$

The reactive component of the second harmonic current is too small to matter in this case. Therefore, $i = 1.000 \times 10^{-9}$ A. From Eq 4

$$A(\text{dB}) = [10 \log (4)(1.000 \times 10^{-9})^2 (1.106)(1 \times 10^9)]$$

$$= -83.54\ \text{dB} \qquad \text{(Eq 9)}$$

In case −83.54 dB looks like an unrealistically huge second harmonic attenuation, you're right! This is the total attenuation relative to the power a matched source ($Z_s = Z_{in}$ = 2700 ohms) would deliver at the fundamental (operating) frequency. In other words, A = −83.54 dB includes an apparent *attenuation* because of the difference at the operating frequency between the inpuut impedance of the network, 2700 + *j*0 ohms, and the assumed source impedance, 1 × 10⁹ + *j*0 ohms. The disparity between these two impedances is not a true mismatch from the standpoint of the source. The tube is delivering rated power into a 2700-ohm load corresponding to the plate-load resistance. The actual second-harmonic attenuation is the difference between the total attenuation determined above and the apparent attenuation at the operating frequency owing to the apparent mismatch between Z_s and Z_{in}.

We find the mismatch attenuation at the operating frequency using the same route we used previously. The current is

$$i_{fund} = \frac{1}{Z_{in} + Z_{in}}$$

$$= \frac{1}{(1 \times 10^9 + j0) + (2.7 \times 10^3 + j0)}$$

$$= 1.000 \times 10^{-9}\ \text{A} \qquad \text{(Eq 10)}$$

from Eq 6. Substituting into Eq 4,

$$A(\text{dB}) = 10 \log [(4)(1 \times 10^{-9})^2 (2.7 \times 10^3)(1 \times 10^9)]$$

$$= -49.67\ \text{dB} \qquad \text{(Eq 11)}$$

The net second harmonic attenuation, then, is

$$A_{net} = -83.54\ \text{dB} - (-49.67\ \text{dB})$$
$$= -33.87 \text{ or } -33.9\ \text{dB}. \qquad \text{(Eq 12)}$$

This is approximately the second-harmonic attenuation expected from the pi network output circuit of a linear amplifier with an operating Q of 10.

A Shortcut

Notice that the magnitude of the current flowing into the network was 1×10^{-9} A at both the operating frequency and the second harmonic. Since the source impedance is also constant at $1 \times 10^9 + j0$ ohms, Eq 4 simplifies to

$$A(\text{dB}) - A_{fund}(\text{dB})$$

$$= 10 \log (4i^2 \times R_{in} \times R_s)$$

$$= 10 \log (4i^2 \times R_{in(fund)} \times R_s)$$

$$= 10 \log\left[\frac{4i^2 \times R_{in} \times R_s}{4i^2 \times R_{in(fund)} \times R_s}\right]$$

or

$$A_{net} = 10 \log (R_{in}/R_{in(fund)}) \qquad \text{(Eq 13)}$$

A_{net} is the net or true harmonic attenuation as defined above, and R_{in} and $R_{in(fund)}$ are the network input resistances at the harmonic and operating (fundamental) frequencies. We have made use of a fundamental property of logarithms, log A − log B = log (A/B). Applying Eq 13 to our pi network example

$$A_{net} = 10 \log(1.105/2700) = -33.9 \text{ dB}. \qquad \text{(Eq 14)}$$

This is the same attenuation we calculated using the longer approach.

Eq 13 can only be used when the input current to the network is virtually constant (independent of frequency). The network attenuation under these conditions is sometimes called the *constant current* attenuation. Constant current is a consequence of a practically infinite source impedance. However, when Z_s is comparable to the input impedance of the network, we can resort to the longer calculation demonstrated above.

Now you may realize why the "how to" of network analysis hasn't received more attention in amateur circles. Calculating an impedance step by step as we did involves a great deal of number crunching! If you're serious about network calculations, you'll want to write a computer program. A PC or a powerful programmable calculator can easily handle these computations.[8] A network calculation that once ate up a computation time of 8-10 minutes and had to be checked for errors now takes 35 seconds on a TI-58 calculator and is practically error free.

Response Curves

By calculating the attenuation at a range of frequencies above and below the bandpass of the network, the frequency response can be profiled. Under the constant-current conditions described, the network attenuation at various frequencies is called the *current response.* On the other hand, if the source impedance is zero, the voltage across the input terminals is constant. The attenuation v frequency characteristic under these conditions is known as the *voltage response*.

Fig 5 shows the current and voltage response of our 2700 to 50-ohm pi network. Included for comparison is the doubly terminated response ($Z_s = Z_{in} = 2700 + j0$ ohms). The voltage response of this network hardly resembles the familiar peaked curve of most tuned circuits. The current response reflects the nearly infinite source impedance of a class-AB tube amplifier. A class-AB amplifier has intrinsic harmonics perhaps 20 dB below the fundamental. This low harmonic level, combined with a current response curve, adds up to a total second-harmonic attenuation of about −54 dB and a net third-harmonic response of approximately −65 dB. These figures reflect the measured harmonics of a tube-type linear with a pi-network output circuit.

Zero source impedance and the consequent voltage response of the output network approximates the operating conditions of a class-C (or D) amplifier. As Fig 5 shows, a pi network does a poorer job (in decibels) of filtering class-C harmonics. The shunt input circuit element (X_1 in Figs 3 and 4) is effectively shorted out by the low impedance source. The relatively high response of the network to class-C harmonics, combined with the large harmonic content of the amplifier, suggests additional filtering may be required to avoid "getting out" on more than one frequency. Another approach might be to find an alternate tank circuit configuration which offers 15-20 dB additional harmonic attenuation with a constant voltage source.

Conclusions

The analysis of a matching network is a little less straightforward than many of the textbook networks one often sees. The results, however, have direct, practical applications. Calculated attenuations can be used to compare different networks in a specific application, to evaluate a single network with various source or load impedances, or to see how a change in circuit Q affects the attenuation.

Networks other than the low-pass pi can be reduced to the simplest series equivalent circuit using the same route we discussed. The attenuation of other networks is also calculated the same way. Eq 4 is usable with any source impedance—the source need not be restricted to an open or short circuit. Even a reactive source impedance works just as well, as long as it is in series-equivalent form. I hope this excursion whets your appetite to explore what matching network analysis can do for you!

(Continued on page 13)

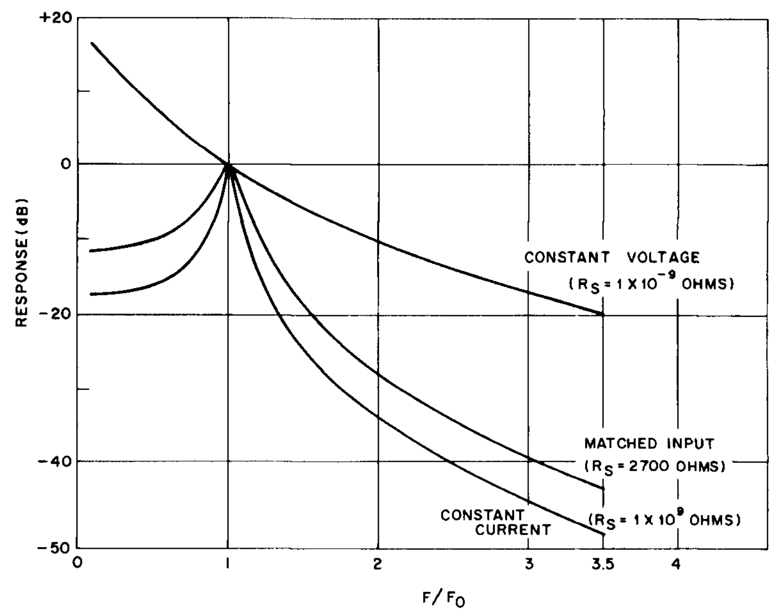


Fig 5—Frequency response of a 2700 to 50-ohm pi network, with an operating Q of 10. Reduced frequency f/f_o is relative to the resonant or operating frequency, f_o.

High-Pass Filters to Combat TVI

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The Problem

While casting about to solve TVI problems associated with a VCR installation, I consulted the interference chapter in *The 1987 ARRL Handbook*[1]. I found several high-pass filters that appeared to be relatively easy to construct and instantly recognized the filters to be the handiwork of Ed Wetherhold, W3NQN. Having done other filter evaluations with Ed, I knew that the selection of components could have some very subtle effects on the performance of these filters.

The Solution

My selection of capacitors was limited. I did not have the recommended ceramic or monolythic types, but I did find "small" epoxy-coated, silver-mica capacitors and usable ceramic-chip capacitors. Following the instructions for constructing a handmade microstrip, the size of the silver-mica capacitors drove the total filter size to require a 2.1 × 1.6 × 2.8 inch BUD Mini-box enclosure. Using chip capacitors in the filter design allowed for an intrinsically neat assembly job and resulted in a filter able to fit into a small 1.1 × 1.4 inch case (Figs 1 and 2).

With the investment of my time and materials in the two filters, I was curious as to the actual performance differences. Did the chip capacitors really enhance the filter performance? The *Handbook* did not indicate what kind of stopband characteristics were available from the design, although Fig 40-21 presents the passband results. This is where W3NQN gets into the act. Ed was kind enough to honor my request for spectral data on both of the filters that I sent to him. I'll let Ed continue the story.

Testing the Solution

"Insertion loss evaluations of your two high-pass filters have been completed. Because the test equipment required was not as freely available as I would have preferred, I did not have enough time to perform additional testing that might have answered some questions I have regarding the results.

"For example, the filter response curve of my high-pass filter shown in Fig 40-21 appears to be flatter (and much better) than I recall. An attempt to repeat the passband response shown in the figure using my filter in the same test setup used to test your filters was unsuccessful. Instead, the passband response of my filter now looks similar to that shown for the chip-capacitor filter."

Ed went on to thoroughly explain the problems of evaluating 75-ohm filters in 50-ohm equipment; not a trivial problem when a precise answer is desired. He notes, "Consequently, because of my present doubts regarding the validity of the absolute insertion loss results, I recommend you do not put too much trust in the filters. I think, however, you can have faith in the relative responses between the chip capacitor and the mica-capacitor filter responses.

"Comparing the relative responses of the chip v mica-capacitor filters, (see Figs 3 and 4), it seems like the 50- to 250-MHz passband of the chip-capacitor filter is slightly better than the mica type.

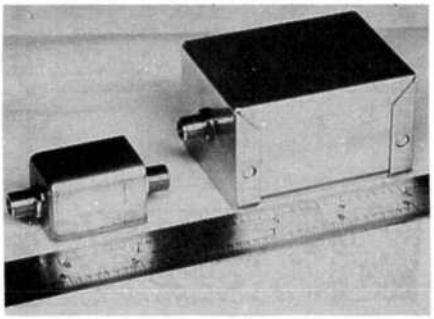
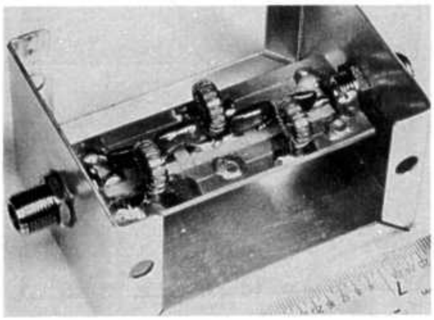
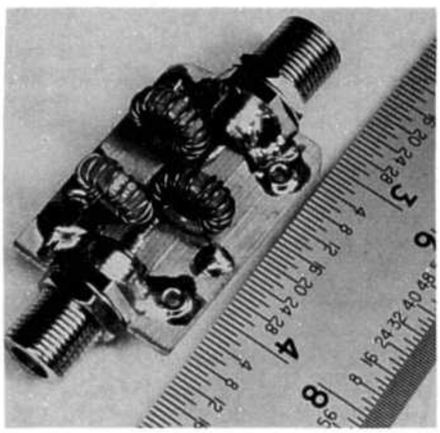


Fig 1—High-pass TVI filters. The unit to the left uses ceramic-chip capacitors, whereas the unit to the right uses silver-mica capacitors.

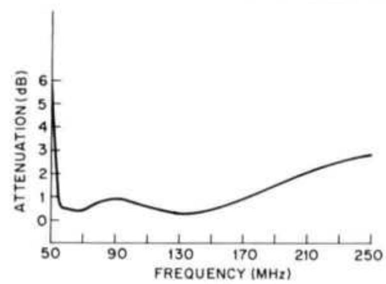


(A)

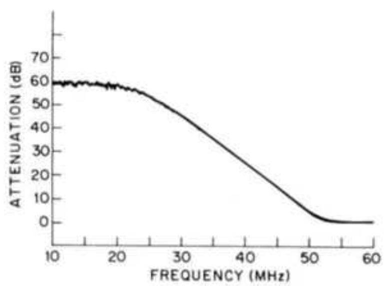


(B)

Fig 2—Internal details of *The 1987 ARRL Handbook* filters. A shows the silver-mica capacitor version and B is the filter using ceramic-chip capacitors.



(A)



(B)

Fig 3—Mica-capacitor high-pass filter performance curves. The curve in A shows the passband curve of the filter; B shows its stopband performance.

[1]Notes appear on page 13.

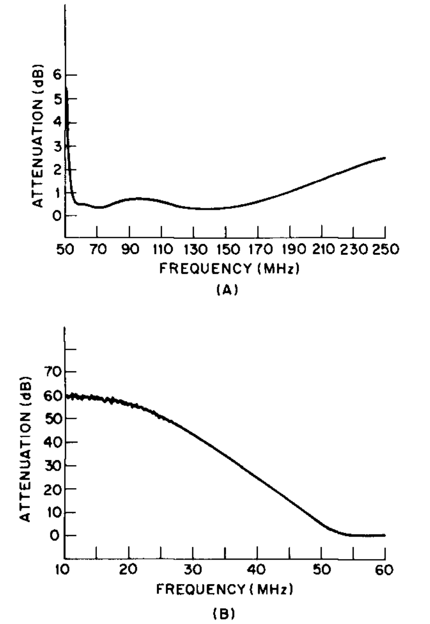


Fig 4—Chip-capacitor high-pass filter performance curves. A shows the passband curve and B shows its stopband performance.

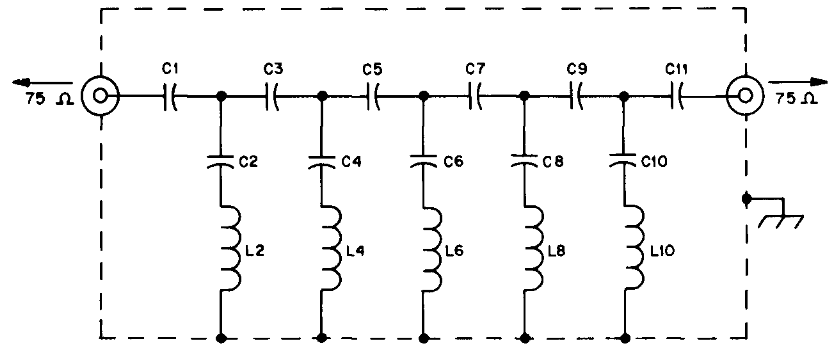


Fig 5—The J. W. Miller TVI high-pass filter, 75/75 ohms (part no. C-513-T2).

The stopband responses are essentially identical and I don't see any significant differences below 50 MHz. Although the chip-capacitor filter has slightly better passband performance, it isn't so much better that it absolutely requires chips to be used."

So much for trying to extend UHF technology to TVI high-pass filters. Personally, the stopband performance was disappointing for my 50-MHz VHF interests and the protection of the 43-MHz TV IF frequencies. I considered the experiment worthwhile enough to get some data on. Ed, provided the clincher when he sent me data on a commercial filter.

The Cure

"Incidentally, if you are looking for an excellent high-pass filter, I recommend the J. W. Miller 75-ohm TV filter (part no. C-513-T2) shown in Fig 5. This filter costs $22.50 plus $3 for UPS shipping. Table 1 is the parts list for the filter. Fig 6 shows an exceptionally good stopband for the C-513-T2; it should eliminate any TVI problems stemming from an input overload. I don't believe it is practical for a ham to duplicate the response of the J. W. Miller filter at a price less than $25."[2]

Ed has kindly presented all of the spectral data as well as the information on the approximate values contained in the J. W. Miller C-513-T2 filter. For my part, I have been unable to locally acquire the J. W. Miller filters and am going to write for the address of my nearest supplier. Send inquiries to Bell Industries, J. W. Miller Div, 19070 Reyes Ave, Box 5825, Compton, CA 90224, tel 213-537-5200.[3]

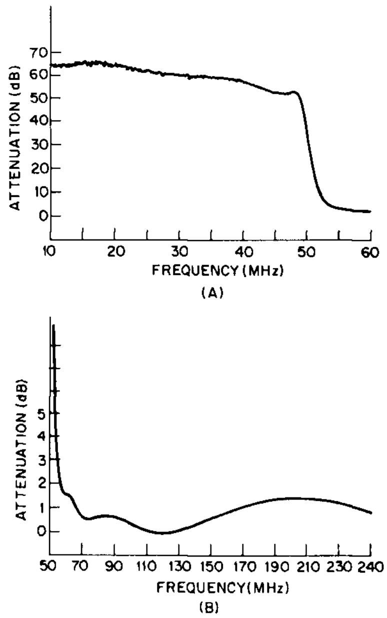


Fig 6—Performance curves of the C-513-T2 filter. A shows the filter's stopband performance and B displays its passband performance.

Conclusion

As for my TVI-solving abilities, I have had successes and failures. The particular problem that initiated my research of high-pass filters was to help a friend. We completely rewired his TV antenna feed from twin-lead to coaxial cable, added a coaxial-cable, common-mode ferrite choke at the VCR, a high-pass filter (that needs to be upgraded to a J. W. Miller filter), and finally changed the VCR/TV IF from Channel 4 to Channel 3. With this setup, the interference from his near-field HF emissions were almost completely eliminated.

At my own QTH, problems still exist on 20 and 10 m. Six meter emissions do not affect my own TV set. My TV antenna is at an altitude of 25 ft and some 30 ft directly beneath the 6-m beam. A nearby neighbor experiences severe audio detection of my 6-m SSB signals on every channel, but FCC spectral examination of these signals show them to be very "clean." The problem appears to be

(Continued on page 13)

Table 1

Component Values For J. W. Miller's TVI High-Pass Filter

Capacitor No.	*Values (pF)*	*Inductance ID No.*	*(µH)**	*Color Mark*	*Q @ 25 MHz*	*Suggested Standard Value*
C1, C8, C9	33	L2	0.178	Silver	50	0.18 µH
C2	390	L4	0.272	Black	50	0.27 µH
C3,C6	27	L6	0.400	Yellow	50	0.39 µH
C4	56	L8	0.348	Green	50	0.33 µH
C5	39	L10	0.227	White	50	0.22 µH
C7,C11	47					
C10	100					

*Inductance measured with H-P Q-Meter, Model 4342A.

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(All dates 1986 unless otherwise noted.)

Bits

RF Design Contest No. 2

This is the time of year for engineers to put on their thinking cap, take pencil in hand and design a circuit with an obvious RF function and one that operates in the below 2-GHz frequency range. If you can meet these requirements, then you will be interested in the second annual *RF Design* contest, sponsored by *RF Design Magazine*. Entry designs are required to be RF circuits containing no more than 6 single active devices (tubes or transistors), or 4 ICs, or be passive circuits of comparable complexity. The circuit must be the original work of the entrant. If it was developed as part of the entrant's employment, entries must have the employer's approval for submission. Components used must be generally available, not obsolete or proprietary. The submission of an entry implies permission for *RF Design* to publish the material. If the design is published, the entrant will receive the normal author's honorarium.

Judges for the entries are Gary Breed, K9AY, editor of *RF Design*, Andy Przedpelski, VP of Development, ARF Products, and Dan Baker, WA7KRN, 1986 winner and engineer, Tektronix. Each entry will be evaluated in three categories, each having equal weight: 1) originality, 2) engineering and 3) documentation. Prizes are valued at over $18,000 and include a Hewlett-Packard 8590A spectrum analyzer, the Touchstone/RF circuit design program, a Bird Electronic 4421 RF Power Meter and the SuperStar program, which simulates a wide variety of RF and microwave circuits.

All entries must be submitted to *RF Design*, 6530 S Yosemite St, Englewood, CO 80111 by March 31, 1987. Good luck!—*Gary Breed, K9AY, RF Design Magazine*

By Bill Olson, W3HQT
Box 1655A, RFD 1
Burnham, ME 04922

Solid-State Construction Practices, Part 2

Last month I discussed various RF components and their use at VHF and higher frequencies. This month I will discuss how to stick all these components down and connect them together, as well as how to enclose them and get the RF in and out of the enclosure.

There are a few basic rules for VHF and UHF construction that, when understood, make everything else fall into place.

1) Parasitic inductances and capacitances must be kept to a minimum.

2) Transmission lines (usually 50-ohm lines between stages) and their associated connections in and out of the box must not contain appreciable discontinuities.

3) The enclosure must not look like a waveguide at the frequency of use. Rather, it should look like a waveguide beyond cutoff so that the box itself doesn't become a feedback path for the circuit within the box.

4) With tightly packed, no-lead construction, care must be taken that there are no mechanical stresses on *any* of the components within the box.

The following discussion is meant to impart a general understanding of what is going on, not to specify, for example, how big a box should be for a particular frequency. Since this is an "experimenters" magazine, *experiment* a little. That's the best way to learn. I will be glad, however, to answer any specific questions via the mail. Please include an SASE.

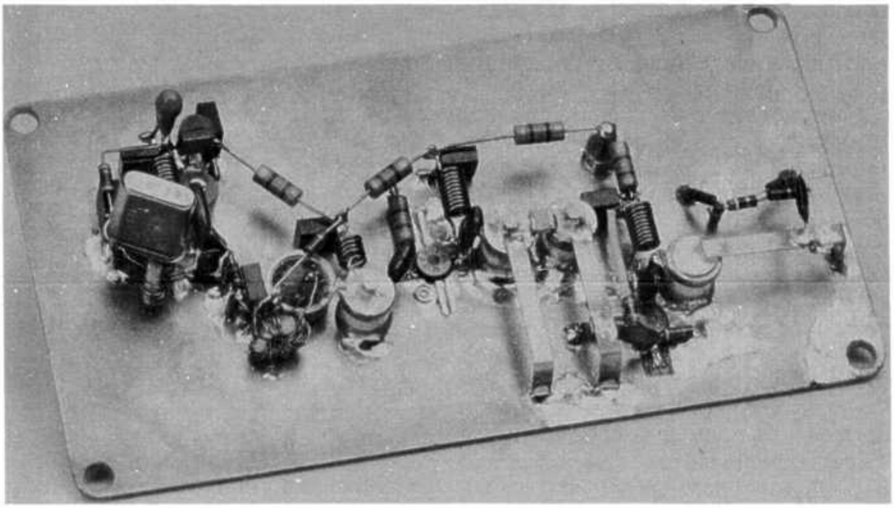

Fig 1—Dead roach construction techniques were used on this 404-MHz local oscillator built by Paul Drexler, WB3JYO. Components are supported by the connector, feedthrough capacitor and the piston trimmer capacitors. Remember that with this construction technique, the ground for RF is on the component side of the board.

Circuit Layout

A circuit should be laid out so that lead lengths are kept to a minimum and so that the transition to the outside world is through connections that maintain a constant impedance. With most VHF and UHF circuits, the actual layout looks very much like the schematic diagram. This is the best place to start.

All ground connections should be kept as short as possible. This often means "no leads" and using leadless components (like sandwich mica capacitors and so on) where applicable. Remember that in microstrip circuitry, the ground return is the back side of the circuit board (the unetched side), so all component grounds should be returned to this plane. More on this later.

There are two basic circuit building techniques. The first is where the components are supported and soldered together by their own leads in a compact mess. See Fig 1. The main supports for the assembled mess are those components that actually mount to the enclosure: capacitors, connectors, feedthroughs, stand-off insulators and the like. This type of construction is commonly known as "dead roach" construction (or dead bug construction for those of us not blessed). Visualize a TO39 transistor with its metal cap soldered to the floor of the box and its legs sticking up in the air. . . that's dead roach construction. While commonly used below 500 MHz, I have "dead bugged" power amplifier circuits at 1296 MHz with good success. Many low-noise preamp circuits are constructed this way, even up through 2304 MHz.

The circuit is built on a conductive ground plane—most often a piece of unetched PC-board material because of its availability and easy workability. Sheet brass or copper works as well. The ground return is on the component side of the board so assembly is quite easy. It quickly gets tedious, though, if there are a lot of components or if you are building more than one of the same circuit. When building a circuit dead roach style, the components that go to ground, as well as the connectors and feedthrough capacitors, are usually mounted first. It helps to draw the whole thing to scale first so you know where to drill the holes. Then the rest of the components are soldered in. If everything is kept as close as possible, parasitic capacitances and inductances will be acceptably low, and there won't be any components flapping around.

RF connectors can either mount through the ground plane itself or in the final enclosure. Make connections with short lengths of coaxial cable or *very* short *low-inductance* leads.

Tin-can type transistors that have an isolated can or a grounded can (emitter) can be soldered down to the ground plane. (Don't try to solder the collector-connected can of an NPN transistor to ground. You will have a great heatsink, but the 0-volt collector-to-emitter potential will result in rather low stage gain. . .) Transistors with studs or flanges are usually mounted through the ground plane to a heatsink. The leads can be bent up on the inside of the box for convenient attachment points.

If you just want to try a circuit concept out in the VHF or low UHF range, dead roach construction may be the way to go. It's quick and dirty, and with the proper precautions, it works!

The second common construction technique is the microstrip method. In this technique, many of the inductances, capacitances and transmission lines are etched on the PC board. The advantages here are ease of duplication and ease of assembly, as well as ease of use with surface mount and microstrip components. At microwave frequencies, where leads on components cannot be tolerated, microstrip construction is the only way to connect everything together. Disadvantages are large size at lower frequencies and indeterminate coupling through the dielectric material.

From a mechanical standpoint, a disadvantage is that all grounds must be returned to the ground-plane side of the board. With components that have leads, this means drilling a hole through the board and soldering the component lead to the ground plane. With surface-mount or stripline-type components, this means soldering the component to the top side of the board but connecting that top ground to the bottom ground plane. This is usually accomplished by wrapping the edges of the board with copper foil and soldering both sides, or by "riveting" through the board at the component location either with small copper or brass rivets or with pieces of wire. The rivets (or wire) are soldered top and bottom. The idea here is to get the RF current flowing through the component to ground by the shortest path possible. A long return path represents inductance and can cause low gain and instability in a circuit. See Fig 2.

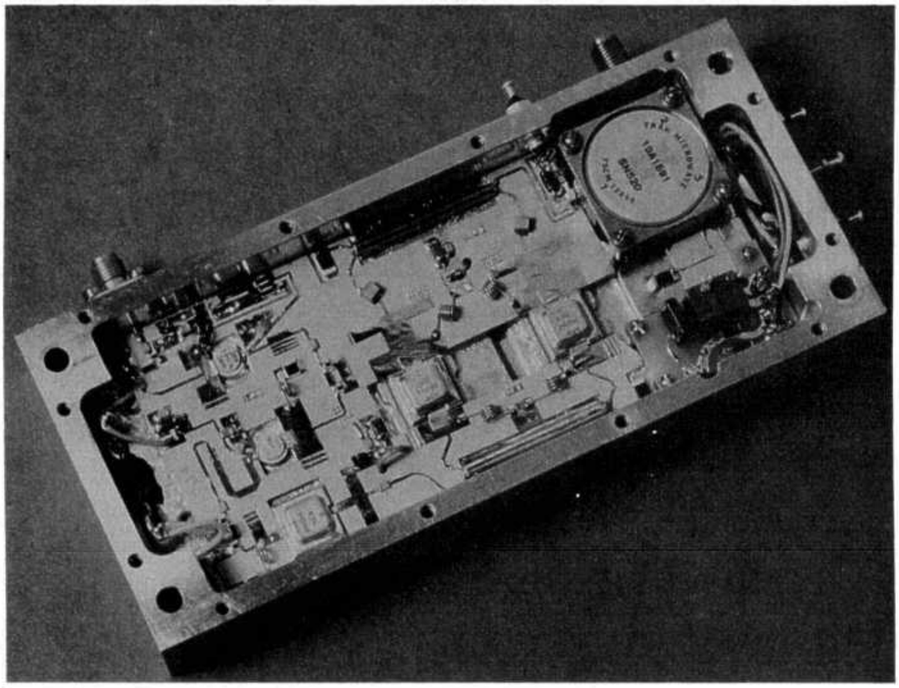

Fig 2—At the other end of the spectrum from Fig 1, note the microstrip construction techniques in this commercially built 1.7-GHz telecommunications amplifier. All ground returns on the component side are connected to the groundplane on the back of the board by plated through holes. The RF input connector uses the end launcher method (see Fig 3). The circuit board and transistors are soldered down to the case in this design for the best possible ground connection! (*photo courtesy Microwave Semiconductor Corp*)

RF Connectors

RF connectors must also connect to the ground plane with low inductance (or more properly, with constant impedance). See Fig 3. A good method is to bolt or solder the connector to the ground-plane side of the circuit board. In microstrip construction, the center pin goes through a hole in the board and attaches to the PC trace on the etched side. (Be sure to remove a little copper around the hole where the pin goes through the ground-plane!) Another good method of attaching the connector to a microstrip board is to use the connector as an *end launcher*. In this method, the PC-board 50-ohm trace is brought to the edge of the board, and the center pin lays on the top of the trace. The flange of the connector can be soldered or attached by some other mechanical means to the ground-plane side of the board. While this method presents some problems when putting the board into a box, it is an excellent technique that is usable to 10 GHz or higher with the proper connectors.

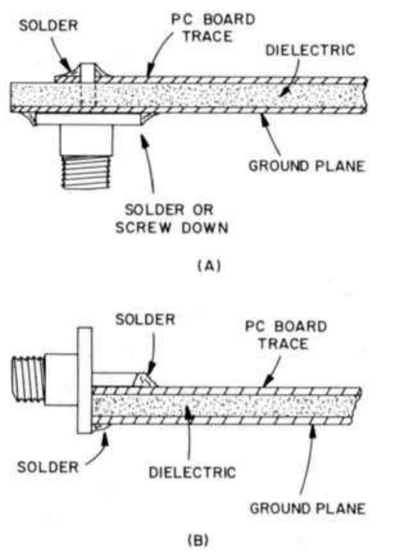

Fig 3—At A, the RF connector is attached through the board. At B, details of the end launcher method are shown.

Small diameter coaxial cable, of course, can be used to connect a board to the outside world. Keep the pigtails as short as possible. Teflon® cable works best because you don't have to be as careful when soldering it. Chances are this connection will be soldered and unsoldered a few times. Make sure the shield returns to the ground-plane side of the board.

Enclosures

The simplest enclosure to build is a box made from PC board material. Build sides right around the circuit board and use the ground-plane side of the circuit board itself as the bottom box. Then solder everything together and tack-solder on a lid. Another method is to bolt the circuit board into the bottom or cover of a ready made box (die cast aluminum boxes are nice here). Bring the dc and RF connections out through both the PC board and theenclosure in a manner consistent with the techniques described above.

The *biggest* problem with enclosures, especially at 900 MHz and above, usually shows up when you put the cover on and the circuit either oscillates or stops working. What is happening here is that the box becomes a waveguide and creates an excellent feedback path between the input and output. This is especially troublesome when there is a lot of gain in the box (multiple stages). Partitions help, but the best cure is a smaller box, or a long skinny box with each stage end to end inside it. Pack those components together and put less stages in each enclosure. You may end up using a few more connectors, but the thing will work!

Mechanical Considerations

When we build circuits without leads, another purely mechanical problem can arise: components crack and break when the circuit is placed under any mechanical stress because there is no place for flexing to occur. For this reason, pay attention to the components that attach to the connectors and feedthrough capacitors. What will happen when the connector is flexed? If a ceramic chip capacitor is soldered right to the connector pin, or even close to the pin on a microstrip circuit, it may crack in two. You will not see the crack, so all manner of frustration will ensue.

Another way to avoid unexplained failures is to *always* screw the PC board, connectors and transistor flanges down to the box *before* soldering in the rest of the components. If you solder the components down first and then bolt the PC board to a box, the stress on the board may cause some components to break. Finally, don't solder a power transistor to the board and then tighten the nut on the stud or screw down the flange ears. That little "crack" you hear is the cost of a new transistor flushing down the drain! Remember: Ceramic is brittle.

Summary

Basically the only trick to VHF and UHF circuitry is to keep it small and keep the leads short. Most components for the higher frequencies that we use these days are quite small, and they don't mind being put close together. Our fingers are really the only problem, and eyesight too I guess. Enclosure size only really becomes a consideration above 500 MHz, since below that frequency it would take a *very* big box to propagate a signal. Believe me, there's nothing as discouraging as building up a whole project and then finding out it can't be made to work except by leaving it out of the box! The rest is all common sense. . .

Next month I will discuss thermal considerations and transistor handling and talk a little about microstrip theory and "home workshop" techniques. I hope you're getting all this down. There's going to be a quiz at the end!

High Pass Filters to Combat TVI

(Continued from page 8)

internal to the TV set. The local cable TV company and I are working together to find a solution. Neither a power-line choke, a common-mode filter, nor a high-pass filter, in any combination, solved the problem. The cable company have acknowledged leaks in the area and are mapping them for correction. Regarding the TVI problems that plague electronic devices, there are no "pat" answers, only a need for a plethora of problem-solving information.

Notes

[1]M. Wilson, ed., *The 1987 ARRL Handbook* (Newington: ARRL, 1985), p 40-13.

[2]Finding a dealer who stocked the J. W. Miller filter was more than a casual exercise. I finally found a distributor: Walder Electronic Distributors, 1801 NE 2nd Ave, Miami, FL 33132, tel 305-339-4794, their Florida WATS no. is 1-800-492-5337 The unit price per filter in Aug 1986 was $27, plus $3 shipping and handling. Florida residents add 5% sales tax.

[3]See note 2.

Figs 1 and 2 were photographed by WD4FAB. The camera was a Nikon F3HP; Lens was a Micro-Nikkor 105mm f/2.8. Kodak technical pan film 2415 (TP135-36) was used and the developer was Kodak LC. The exposure time was 8 seconds at f/32. The prints were developed on Kodak Polyprint RC paper.

Analyzing Simple Matching Networks

(Continued from page 6)

Notes

[1]F. Napurano, "How to Perform AC-Circuit Analysis," *QST*, May 1985, pp 19-22.

[2]R. Schetgen, "Simple Conversion of Complex Networks," *QST*, Nov 1985, pp 41-43.

[3]Ref 2, p 43.

[4]Ref 1, p 21.

[5]W. Hayward and D. DeMaw, *Solid State Design for the Radio Amateur*, (ARRL: Newington, CT), 1977, p 53.

[6]Paul Shuch, "Gaining on the Decibel: Part 1," *QST*, Feb 1986, pp 20-22.

[7]Ref 5, p 52.

[8]For $1 and an SASE, the author will supply TI-58/59 calculator programs for the common matching networks.

Appendix

Let us justify Eq 4 from Eq 5 by

$$i_c = \frac{V_s}{Z_s + Z_{in}} \qquad \textit{(Eq A1)}$$

The power drawn by the network at frequency f, and transferred to the load, is

$$P_f = i^2 R_{in} \qquad \textit{(Eq A2)}$$

where
R_{in} = *input resistance at f*
i = *magnitude of the current (Eq 7)*

For a constant voltage source, V_s, maximum current i flows and maximum power P_{max} is consumed when the source and the network input are matched, $Z^*_s = Z_{in}$ in Fig 4H. From Eq A1,

$$i_{max} = \frac{V_s}{Z_s + Z^*_s} = \frac{V_s}{2R_s} \qquad \textit{(Eq A3)}$$

The denominator was simplified by using the definition of complex conjugate: $Z_s + Z^*_s = R_s + jX_s + R_s - jX_s$. The two reactive terms cancel.

By definition,

$$P_{max} = i_{max}^2 R_{in} = i_{max}^2 R_s \qquad \textit{(Eq A4)}$$

since $R_s = R_{in}$ with source and input matched. Substituting Eq A3 into Eq A4,

$$P_{max} = \left[\frac{V_s}{2R_s}\right]^2 R_s = \frac{V_s^2}{4R_s} \qquad \textit{(Eq A5)}$$

The attenuation in decibels is

$$A(dB) = 10 \log\left[\frac{P_f}{P_{max}}\right] \qquad \textit{(Eq A6)}$$

Substituting Eqs A2 and A5 into Eq A6,

$$A(dB) = 10 \log\left[i^2 R_{in}\left(\frac{4R_s}{V_s^2}\right)\right] \qquad \textit{(Eq A7)}$$

Setting V_s = 1 V and simplifying yields the attenuation expression, Eq 4:

$$A(dB) = 10 \log (4i^2 \times R_{in} \times R_s) \qquad \textit{(Eq A8)}$$

Intellectual Properties

By Bill Conwell, K2PO
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Portland, OR 97204

Patents Of Interest to Amateur Radio Operators

This month's column features excerpts from recently issued patents in the fields of antennas and diversity receiving systems. Among the antenna patents, Jim Fisher, W8JF, was granted patent 4,595,930 for his design of a *Planar Log-Periodic Quad Array.*[1] Fig 1 shows this antenna as consisting of several concentric coplanar loops mounted about a balanced transmission line feed point. Two twisted balanced lines extend diametrically from the feed point and drive each of the loops at two opposite corners.

The inventor terms this configuration as an improvement over the more conventional three dimensional log-periodic quad array. In this design, the current is distributed symmetrically in a single plane, producing a more uniform feed-point impedance over wide bandwidths. The planar LP array seems to have use in several applications, either as a standalone multiband antenna or as a driven element in a multiband parasitic array.

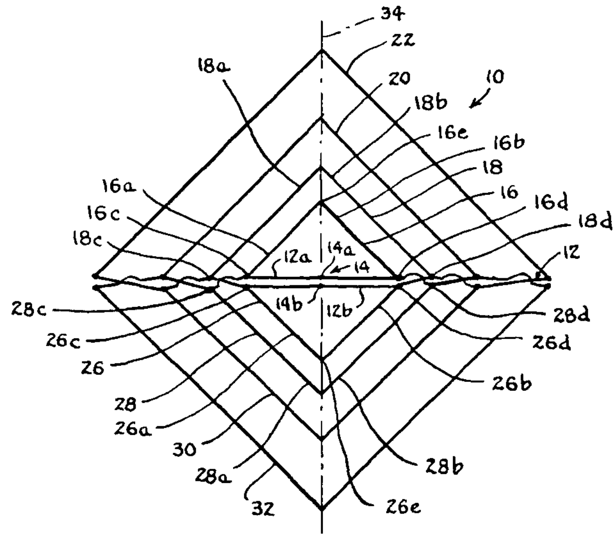

Fig 1—W8JF's log-periodic antenna consists of several concentric coplanar loops mounted about a balanced transmission line feed point.

Telex Communications has patented a *Parasitic Array With A Driven Sleeve Element* (patent 4,604,628). The driven element is cut for 20 m and includes traps for 15 m. Operation on 10 m is provided by "open-sleeve" elements 30 and 32 (Fig 2) which are parasitically driven from the driven element. Element 30 is resonant in the 10-m band. Element 32 is shorter than element 30 by the distance between the elements. In addition to providing driven-element operation on 10 m without additional traps, the sleeve elements are claimed to reduce the SWR on the other bands.

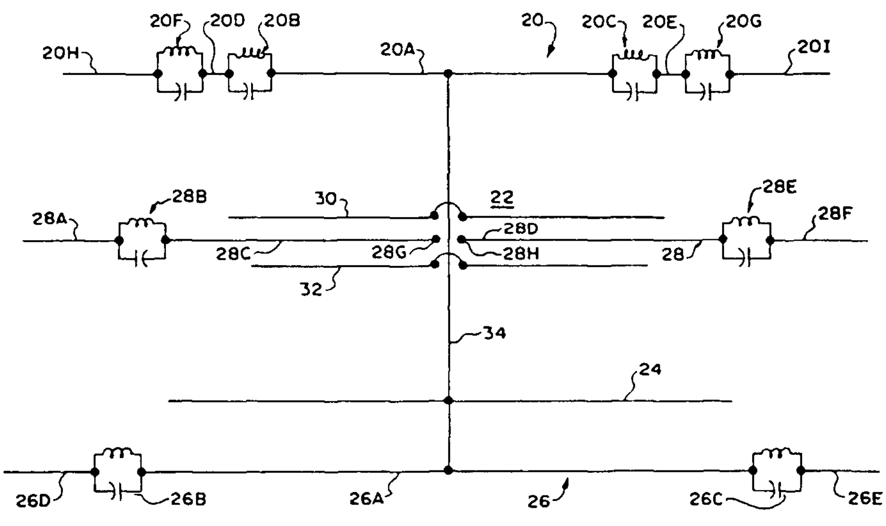

Fig 2—Telex Communications' Parasitic Array With A Driven Sleeve Element. Elements 30 and 32 allow operation on the 10-m band.

Another multiband antenna is described in patent 4,593,289, owned by Butternut Electronics and entitled *Multi-Band Dipole Antenna With Matching Stubs.* The antenna, shown in Fig 3, uses a shortened bow tie-shaped dipole to resonate on 20, 15 and 10 m by gamma rods driven from a centrally extending conductor. Gamma rods 122 and 124 resonate the antenna on 15 m, while rods 90 and 92 resonate the antenna on both 20 m and its harmonic, 10 m.

The "wingspan" of the 20- and 10-m version is approximately 12 feet, about one third the size of a full size 20-m dipole. The bandwidth of the antenna on each band is said to be about 600 kHz. Parasitic elements of similar design can be added to form a compact unidirectional array.

In the *Obvious-in-Hindsight*[2] class of inventions is Honeywell's patent 4,584,716. Called the *Automatic Dual Diversity Receiver*, it is a simple diversity system for receiving signals concurrently broadcast on two frequencies, such as

[1]Notes appear on page 15.

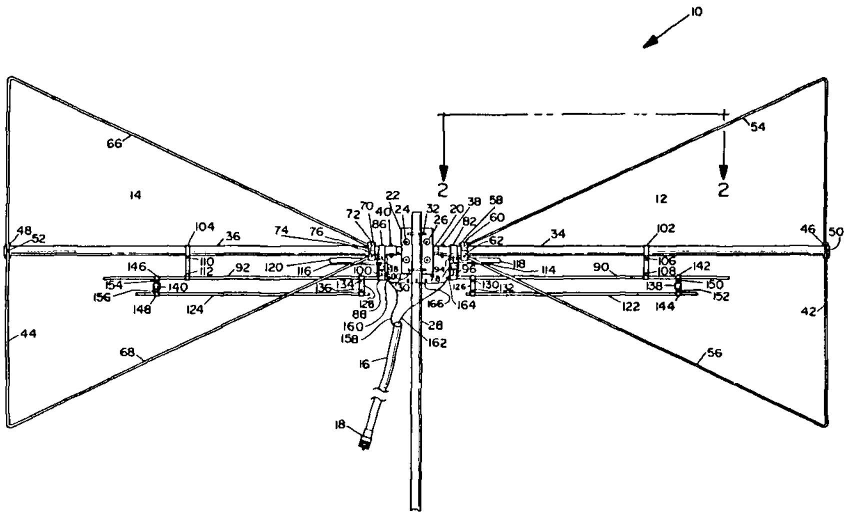

Fig 3—The Multi-Band Dipole Antenna With Matching Stubs. Its theory of operation is explained in the text.

NBS broadcasts on 5 and 10 MHz. No switching circuitry is involved. Instead, a broadband front end provides both the 5- and 10-MHz signals to a mixer. The local oscillator frequency is selected midway between the two receive frequencies (ie, 7.5 MHz) thereby producing an IF signal consisting of both received signals at 2.5 MHz. The remainder of the receiver circuitry is conventional and serves to demodulate the IF signal so as to recover the audio from both the 5- and 10-MHz broadcasts.

Preposterous Patents

On the lighter side, the Patent Office does not pass judgment on an invention's technical merits. It only checks to confirm that the invention is new, useful and nonobvious. Sometimes, technically questionable inventions slip by. One example is patent 4,584,586, a VHF-UHF TV antenna shown in Fig 4. A brief excerpt from the patent illustrates this point:

"It is believed that RF signals are magnetically drawn along patterns that conform to the magnetic field lines of the earth. Therefore, the configuration of the present array, which itself bears a general resemblance to the magnetic flux lines of the earth, contributes to its effectiveness. That is, it is believed that the shape of the present array contains spherical curvatures which conform to the curvature of the magnetosphere of the earth. . ."

The patent goes on to note that the hollow space inside the 0.325-inch tubing, from which the antenna is made, acts as a waveguide within which the signals captured by the array are carried to the twin-lead feed line.

Other recent patents of potential interest to the amateur community are:

4,578,819, Sony's *Space Diversity Receiver.*

4,580,092, *Pocket Size Antenna Match Indicator For VSWR Measurement*, owned by USA (through the Secretary of the Air Force).

4,591,867, *Attenuation Apparatus For An Aerial Array* (isolates transmit and receive antennas mounted on the same tower), owned by Telefonaktiebolaget LM Ericsson.

4,608,572, *Broad-Band Antenna Structure Having Frequency- Independent Low-Loss Ground Plane* (a variant of the discone antenna), owned by Boeing.

4,608,574, *Backfire Helix Antenna*, owned by USA (through Secretary of the Air Force).

Complete details on each of these inventions can be found in the associated patent. Copies of US patents are available for $1.50 each from The Commissioner of Patents and Trademarks, Washington DC 20231 (expect about a four-week delivery time).

Caveat: The US Patent laws provide that "Whoever without authority makes, uses or sells any patented invention, within the United States during the term of the patent therefor, infringes the patent."

Information on these patented inventions is printed here in accord with *QEX*'s stated purposes of (1) providing a medium for the exchange of ideas and information; (2) documenting advanced works in the Amateur Radio field; and (3) supporting efforts to advance the state of the Amateur Radio art. These inventions may not be practiced, without permission of the patent owners, until seventeen years after the patents' issue dates. Nonetheless, it is believed that publicity of such advances in the state of the art will broaden the perspective of amateur experimenters and promote future technical advances within the amateur community.

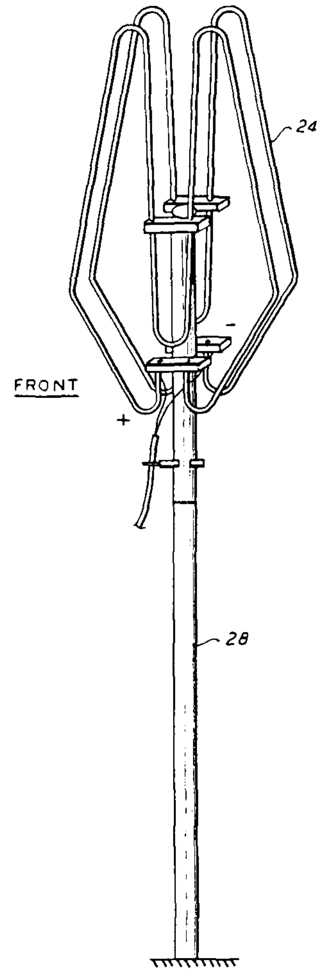

Fig 4—Inventions that pass through the Patent Office are not judged on their technical merits. This VHF/UHF TV antenna claims that because of its shape, its operation conforms to that of the Earth's magnetosphere.

Notes

[1]J. Fisher, "Development of the W8JF Waveram: A Planar Log-Periodic Quad Array," *Antenna Compendium*, p 50.

[2]B. Conwell, "Patents," *QEX*, Apr 1986, p 16.