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Issue No. 269

VK1BG describes the steps he took to quiet the RF noise generated by an inexpensive switch mode power supply. He reports no detectable noise produced by the SMPS from the bottom end of the AM broadcast band through the 10 meter amateur band.

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About the Cover

Ian Cowan, VK1BG, bought an inexpensive switch mode power supply to use with a broadcast band receiver and his Yaesu FT-897 transceiver while he was traveling. As he feared when he bought the supply, it was a source of birdies all across the MF and HF spectrum. Here, he describes the steps he took to quiet that RF noise. He reports no detectable noise produced by the SMPS from the bottom end of the AM broadcast band through the 10 meter amateur band.

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Larry Wolfgang, WR1B

Empirical Outlook

On the weekend of September 15 through 19, I attended the 30th ARRL and TAPR Digital Communications Conference. What a fun and exciting weekend this turned out to be! If you attended, you will know what I mean. If not, you missed an excellent conference.

I will give you a few highlights from the conference, in the hope that you might start making plans to attend next year's DCC. Plans are just in the beginning stages, and neither a date nor exact location are known at this time. DCC has usually been held in late September — generally the 4th weekend of the month — and the 2012 Conference is being planned for Atlanta. We will announce the details in *QEX* as soon as they become available.

The DCC includes full days of technical presentations on Friday and Saturday. This year we started with Scotty Cowling, WA2DFI, giving us an update on the High Performance Software Defined Radio (HPSDR) project. Scotty had several exciting developments to report on, including a new ½ W exciter and plans for a new 15 W power amplifier, as well as work on a 100 W PA. If you are one of the many TAPR members and Conference attendees who have been purchasing boards and assembling radios for several years. you may have something that even *looks* like a "real radio" now. If you have not been involved with building an HPSDR, several of the new developments will make the project much more attractive. It is now possible to purchase all of the pieces to build a radio either directly from TAPR (**www.tapr.org/hpsdr_index.html**) or one of several other suppliers. Boards no longer available from TAPR are now sourced through iQuad-Labs (**http://iquadlabs.com**). The HPSDR project is beginning to move beyond the purely experimental stage, although it is still intended for hams who want to experiment with the hardware and software. For more information about the HPSDR project, see the website at **http://openhpsdr.org/**.

George Heron, N2APB, and Dave Collins, AD7JT, gave a presentation about adding CW to the NUE-PSK modem. (The NUE-PSK modem was described in the Mar/Apr 2008 issue of QEX, after a presentation at the 2007 DCC.) At last year's DCC Sunday Seminar, George heard about a DSP filter type called a Goertzel filter, and thought it might have application on this project. It turns out that this filter gives perhaps the best CW decoding of any stand-alone code reader. I have been reading a lot of comments about this new feature on the NUE-PSK Yahoo group, and there is a lot of praise for its ability to copy CW.

In a talk that certainly generated a lot of interest and enthusiasm, David Rowe, VK5DGR, presented "Codec2: Open Source Speech Coding at 2400 bits/s and below." This talk presented work by David to produce a Codec that can replace the AMBE Codec now used in D-Star and other digital voice systems.

In addition to the detailed technical presentations, many attendees look forward to the introductory sessions presented separately on Saturday. This year there were presentations on digital data modes, HF digital voice, D-Star digital voice and data, and an introduction to APRS. These sessions always present an opportunity to learn the basics from real experts in the field. For example, if you want to learn about APRS, who better to teach you than Bob Bruninga, WB4APR?

A highlight for many attendees each year is the Sunday morning seminar. This is a 4 hour in-depth presentation, and this year we learned about "Universal Ham Radio Connectivity by Callsign" from Bob, WB4APR, and several other presenters. The basic concept is that we all have a unique identifier in our call sign, and there are a variety of ways that we can be contacted using our call sign. For example, the D-Star system can find a registered user just about anywhere, and establish a link to that station from any other caller. The latest reported location of APRS users can be found at **www.aprs.fi** and a few other websites, and the APRS system can forward and deliver text messages to other users. Many hams use the WinLink system to send and receive radio e-mail as well. The question is, why can't these systems talk to each other? Bob and his panel of presenters described efforts to create just such connectivity, and invited everyone to join the efforts.

Of course the DCC isn't just about technical presentations. If it were, you could simply purchase the set of DVDs produced by Gary Pearce, KN4AQ, of Amateur Radio Video News, so you could watch the presentations at your leisure (still a good idea). The opportunities to socialize with the "movers and shakers" of the digital communications world are priceless. If you have a question (or idea) after seeing a presentation, you can talk with the author and learn even more.

Everyone is encouraged to bring some project to share in the Demo Room. There are often manufacturers there showing off their latest digital communications devices. This year you could try out one of George (N2APB) Heron's SDR Cube transceivers, examine the RPC Electronics (Jason Rausch, KE4NYV) new R-Trak Mini APRS tracker and talk to Jason about his YagTracker APRS tracker and terminal. Turn just about any radio into a complete APRS terminal and tracker with this modem and display unit. There were several other projects on display as well.

There are several other excellent technical conferences, such as VHF/UHF conferences, Microwave Update and technical presentations at hamfests and conventions that are all worth attending. Start planning now for which ones you will attend in 2012!

Tom Apel, K5TRA

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A New Horizontal Polarized High Gain Omni-Directional Antenna

Tom presents a helical antenna design that produces significantly horizontally polarized gain.

Background

High gain omni-directional antennas are more difficult to realize with horizontal polarization. Vertical radiating elements stacked along a vertical line provide a natural means of achieving high gain with vertical polarization. Radiating elements are often $\lambda/2$ or near $\lambda/2$. When these elements are rotated to a horizontal mode, the familiar bidirectional dipole azimuth pattern is seen. Turnstile arrays were early answers to the challenge of omni-directional performance with horizontal polarization.[1,2] Dipoles also have been wrapped into circular (Halo) or square (Squalo) shapes to mitigate the pattern; however, gain is reduced.[3] Perhaps the best implementation of circularly wrapped dipoles is the Big Wheel where three dipoles form a circular array.[4] An excellent printed board implementation has been done by Kent Britain, WA5VJB. Basic performance of the Halo and Big Wheel structures has been extended by use of folded dipole elements. The folded dipole Halo has been done by Delbert Fletcher, K5DDD, and the folded dipole Super Wheel is credited to Tom Haddon, K5VH. Slots in cylinders or in rectangular wave guides have offered another approach to horizontal polarization at higher frequencies where $\lambda/2$ elements become quite small. Figure 1 shows photos of some of these horizontally polarized omni-directional antennas. Cebik and Cerreto should also be mentioned for the three dipole array that yields a far field radia-

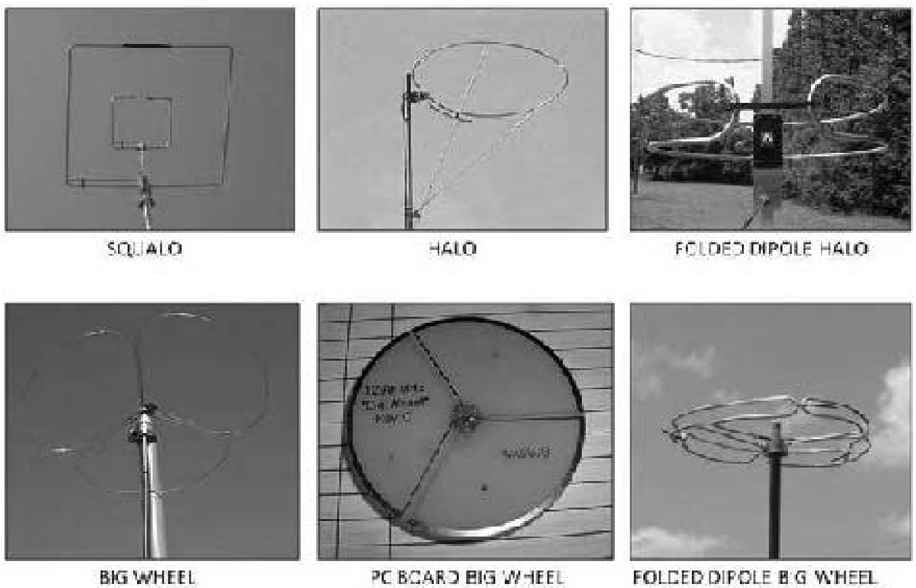

Figure 1– Photos of popular horizontal polarized omni-directional antennas

[1]Notes appear on page 9.

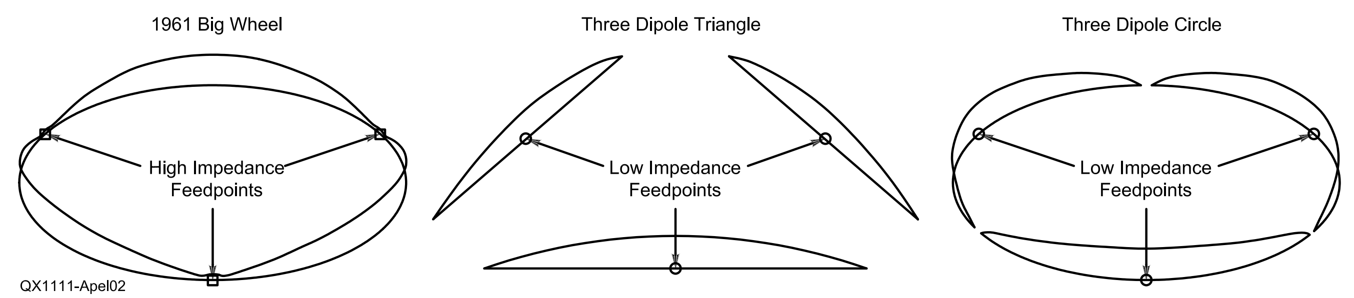

Figure 2 – Current distributions on several arrays of three half wave elements (from Cebik)

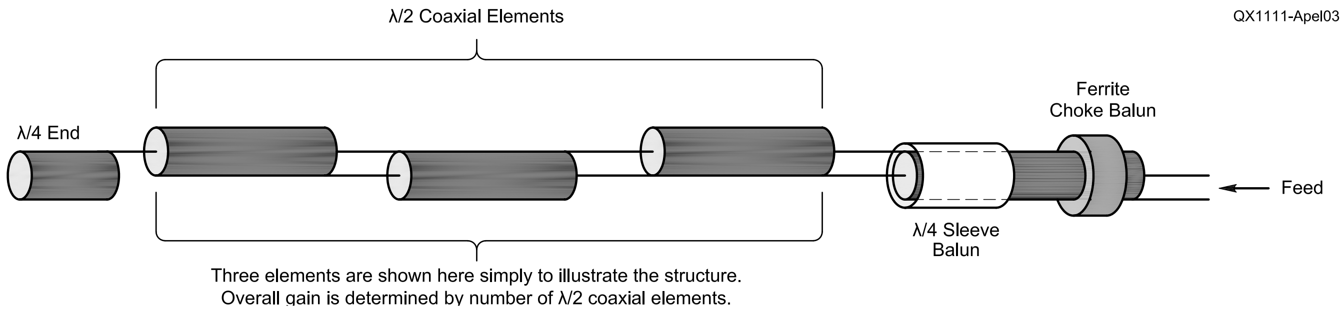

Figure 3 – Coaxial collinear structure showing the electrical connections

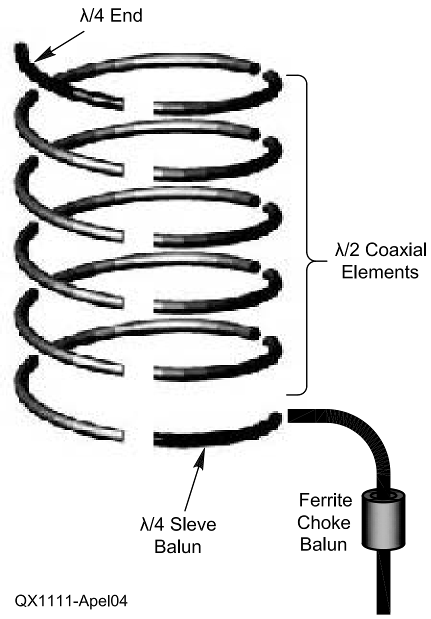

Figure 4 – Coaxial collinear structure showing the physical configuration

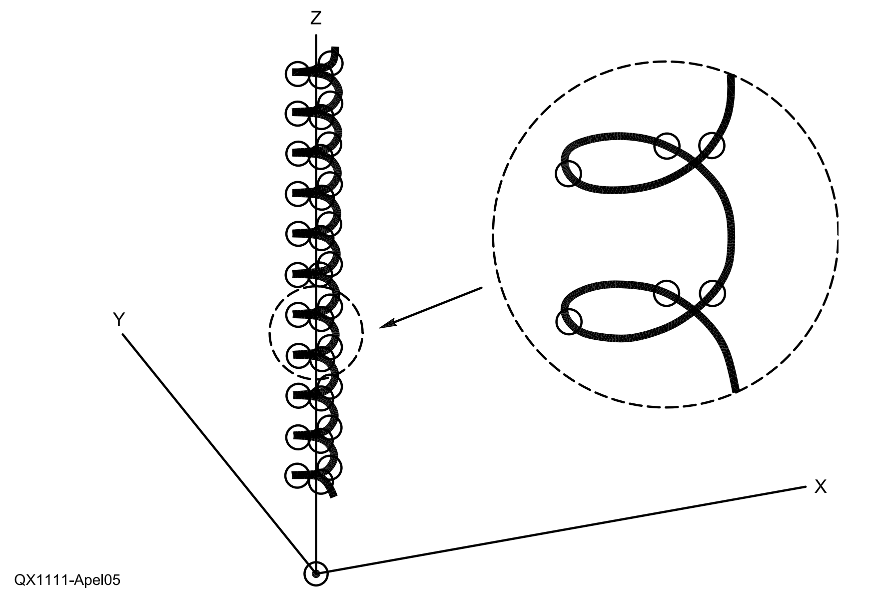

Figure 5 – EZNEC model of 11 turn helical collinear

tion pattern nearly identical to that of the Big Wheel.[5] This is a result of the similarity in current distribution on the three dipoles to that of arrays of three dipoles around a circle as illustrated in Figure 2.

High gain requires stacking an array of horizontally polarized unit structures. The feed complexity associated with ten or twelve stacked elements is not trivial. At this point, it is noteworthy to point out the relative ease in feeding many elements in vertical collinear arrays.

Table 1
Helical Collinear Calculations 902 MHz

	(Inches)	(mm)		
Length λ/2	4.60	116 (0.35λ)	Elements/Turn	3
Diameter	4.10	104 (0.31λ)	Turns	11
Pitch	4.95	126 (0.38 λ)	Total Segments	264
Linear Total	153.78	3906 (11.65 λ)	N λ/2	32
Helix Length	54.48	1384 (4.13 λ)	Segments/Turn	24
Bottom	5.91	150		
Top	60.39	1534		

The Idea

Consider the coaxial collinear structure shown in Figure 3. With the exception of the end elements, all radiating elements are comprised of λ/2 segments of coaxial cable. Each section is end fed by the previous section. Current on the shield of each coaxial section produces the desired radiation. The delay through each section must be 180° in order to properly feed the next section in the array. Hence, the length should be cut to λ/2 in the coax medium. If the λ/2 elements are wrapped around a vertical axis into a helix with three elements per turn, the resulting structure approximates the circular array comprised in stacked unit wheel structures. Figure 4 illustrates the helical collinear structure. This approach allows many elements to be fed simply from a common point. The turn-to-turn pitch is an important design parameter. It trades off horizontal polarization "purity" with gain. The large pitch limit is, of course, the vertical collinear with no horizontal component. One expects good gain in the horizontal mode when pitch approaches λ/2.

Simulation

EZNEC was used to simulate the performance of the helical collinear.[6] The first case considered consisted of 11 turns with a turn-turn pitch of 0.38 λ. This model is illustrated in Figure 5. A sample model-prep calculation sheet is shown in Tables 1 and 2.

With these preliminary calculations, a helical model with periodic current sources can be constructed. The current magnitudes can also be tapered to allow for attenuation on the coaxial line segments. Azimuth and elevation simulation results are shown in Figures 6 and 7. A good omni-directional pattern is achieved with +10.6 dBi gain. This is a bit more than +8 dB over a dipole. The azimuth pattern has approximately ±0.4 dB ripple. A larger turn pitch will reduce this but also increase the vertical polarization content.

The results from a 22 turn simulation are shown in Figures 8, 9, and 10. The turn pitch for this case was approximately λ/2. This structure yields +12.2 dBi (+10 dBd) gain.

Sensitivity

The practical matter of errors in the length of elements must be emphasized. With half wave dimensions of 4.6” in RG-316, a 2° error in the element length is represented by 50 mils. Careful measurement with a caliper can keep element length errors within an acceptable level. The worst case would be

Table 2
902 MHz *EZNEC* Port Definition

Element		*Linear (Inches)*	*Linear (mm)*	*Z-coordinate (mm)*	*Segments*	*Bottom Segment*
1	λ/4	150.48	3822	1504	4	260
2	λ/2	145.88	3705	1462	8	252
3	λ/2	141.28	3588	1421	8	244
4	λ/2	136.68	3471	1379	8	236
5	λ/2	132.08	3354	1338	8	228
6	λ/2	127.48	3237	1297	8	220
7	λ/2	122.88	3121	1255	8	212
8	λ/2	118.28	3004	1214	8	204
9	λ/2	113.68	2887	1173	8	196
10	λ/2	109.08	2770	1131	8	188
11	λ/2	104.48	2653	1090	8	180
12	λ/2	99.88	2536	1048	8	172
13	λ/2	95.28	2420	1007	8	164
14	λ/2	90.68	2303	966	8	156
15	λ/2	86.08	2186	924	8	148
16	λ/2	81.48	2069	883	8	140
17	λ/2	76.88	1952	841	8	132
18	λ/2	72.28	1835	800	8	124
19	λ/2	67.68	1719	759	8	116
20	λ/2	63.08	1602	717	8	108
21	λ/2	58.48	1485	676	8	100
22	λ/2	53.88	1368	634	8	92
23	λ/2	49.28	1251	593	8	84
24	λ/2	44.68	1134	552	8	76
25	λ/2	40.08	1017	510	8	68
26	λ/2	35.48	901	469	8	60
27	λ/2	30.88	784	427	8	52
28	λ/2	26.28	667	386	8	44
29	λ/2	21.68	550	345	8	36
30	λ/2	17.08	433	303	8	28
31	λ/2	12.48	316	262	8	20
32	λ/2	7.88	200	220	8	12
33	λ/2	3.28	83	179	8	4
34	λ/4	0.00	0	150	4	0

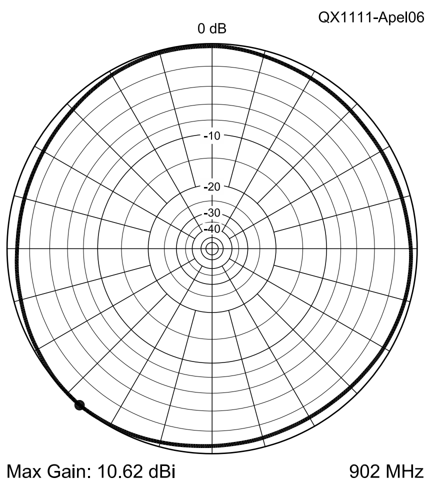

Figure 6 – *EZNEC* azimuth plot of 11 turn helical collinear

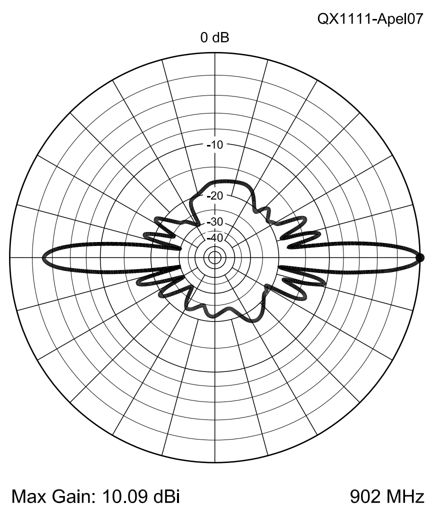

Figure 7 – *EZNEC* elevation plot of 11 turn helical collinear (pitch = 0.5 λ)

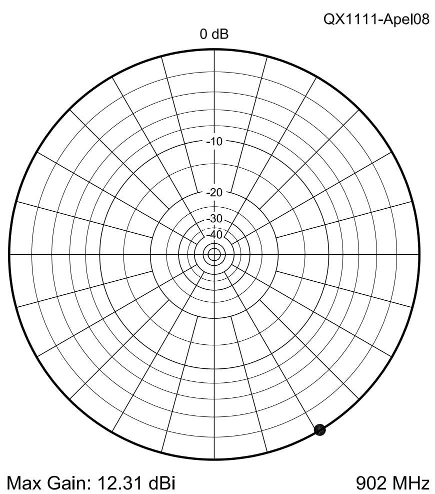

Figure 8 – *EZNEC* azimuth plot of 22 turn helical collinear (pitch = 0.5 λ)

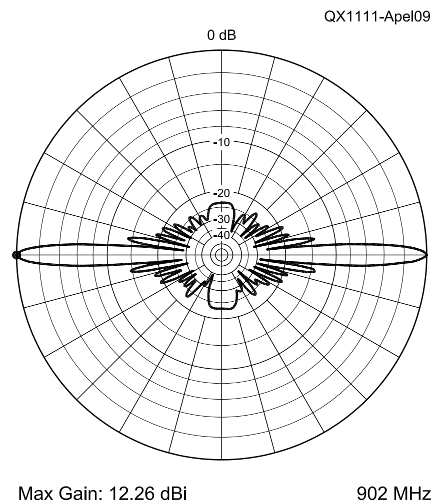

Figure 9 – *EZNEC* elevation plot of 22 turn helical collinear (pitch = 0.5 λ)

if ***all*** elements were short or ***all*** were long. This would propagate a cumulative error throughout the array. While this is unlikely, it is a worst case worthy of analysis. Figure 11 shows a net up or down tilt in the main lobe elevation by approximately 2°, from an analysis of the 11 turn helix with a 2° error in the length of each element. Similarly, Figure 12 shows the same result from an analysis of the 22 turn helix with a 2° error in the length of each element. Practically, one would expect the elements to be constructed to the correct length in the mean, with some distribution in length errors (both short and long). This will spread the main lobe for some net loss in gain rather than a net tilt up or down.

Design Considerations

Each turn contains three λ/2 elements as in "big wheel" structures. I briefly looked at four elements per turn. My thinking was that opposing elements would be antiphased with separation near λ/2, so gain might be good. While this type of structure yields gain, it is not as good as the three elements per turn case and it has a larger diameter. It was concluded that the three elements per turn case was the best.

The most important design parameter is turn pitch. This is the vertical distance between the beginning and end of each turn in the helix. When the pitch is smaller, the larger number of turns emulates "big wheel" structures, but the stacking distance is closer. Horizontal polarization dominates and gain is poor as a result of the effective closer stacking. The other extreme is the vertical collinear when pitch approaches 1.5 λ. This yields very good gain in vertical polarization only. The best stacking distance without grating lobes is λ/2. The obvious question is: How does pitch trade-off the fraction of radiated energy in the horizontal polarization? Figure 13 plots this trade-off for an 11 turn helix. These analysis results are qualitatively representative of other cases with differing number of turns. Several important observations can be made:

1. While total gain continues to increase with pitch, best gain in the horizontal polarization is achieved with pitch ≥0.38 λ.
2. Vertical polarization is –13 dB down from total at pitch of 0.38 λ. This degrades to –8 dB as the pitch is increased to λ/2.
3. Based on the above observations, the optimum pitch is 0.38 λ to 0.4 λ.

Driving point impedance depends on the number of elements. As the number of elements is increased, the impedance lowers. I have built prototypes for 902 MHz and 1296 MHz with 11 turns and 15 turns respectively. Both have yielded good VSWR to

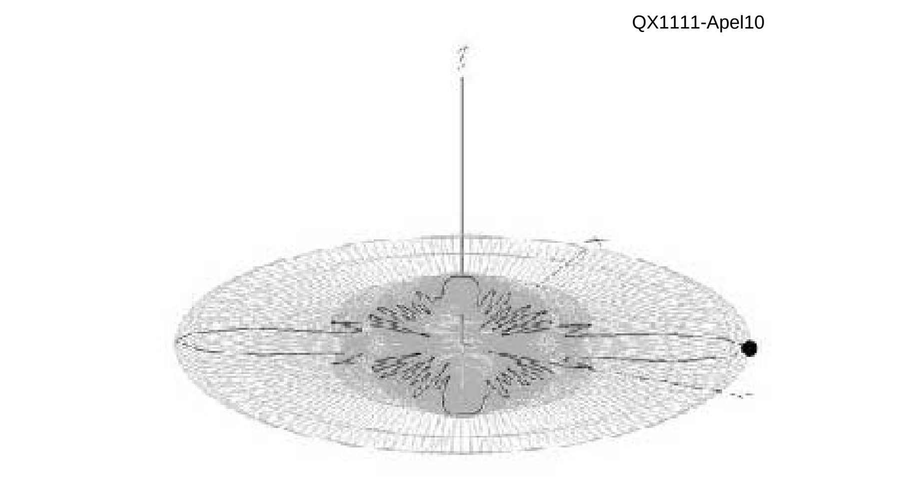

Figure 10 – *EZNEC* 3D plot of 22 turn helical collinear (pitch = 0.5 λ)

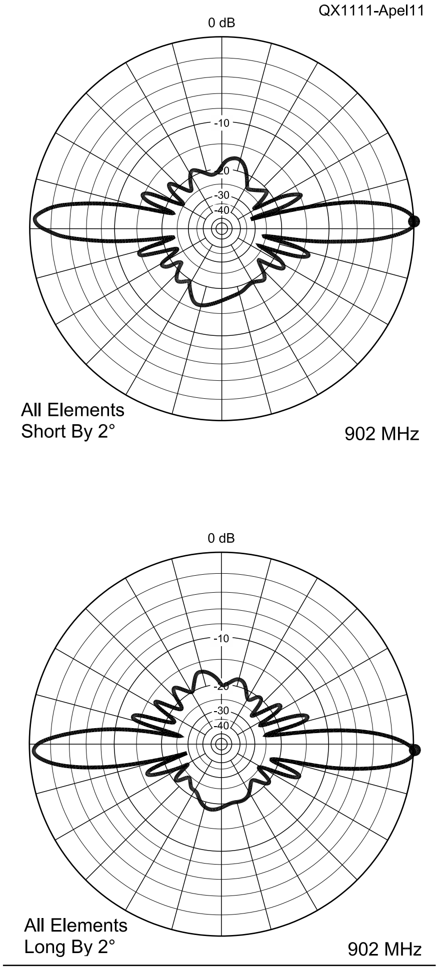

Figure 11– *EZNEC* simulation of 11 turn helix with ± 2° length error on ALL elements (pitch = 0.38 λ). Note the upward tilt for all short and downward tilt for all long.

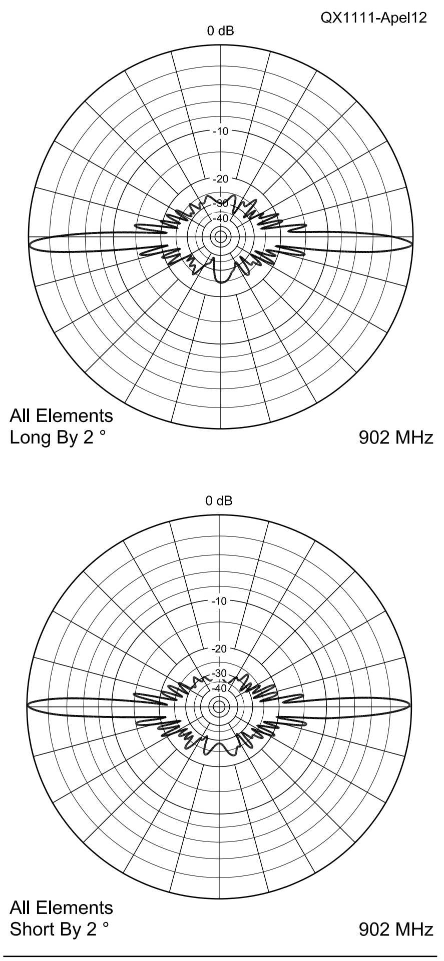

Figure 12 – *EZNEC* simulation of 22 turn helix with ± 2° length error on ALL elements (pitch = 0.5 λ)

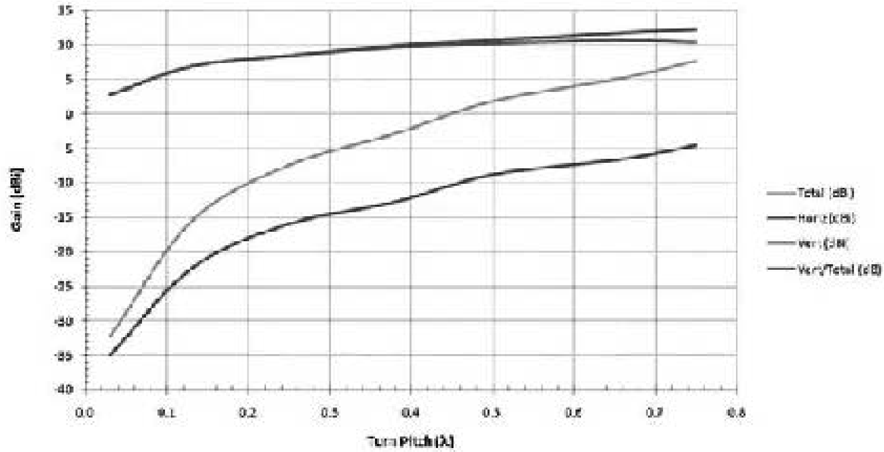

Figure 13 – Gain and polarization trade-offs with turn pitch (from 11 turn helical collinear)

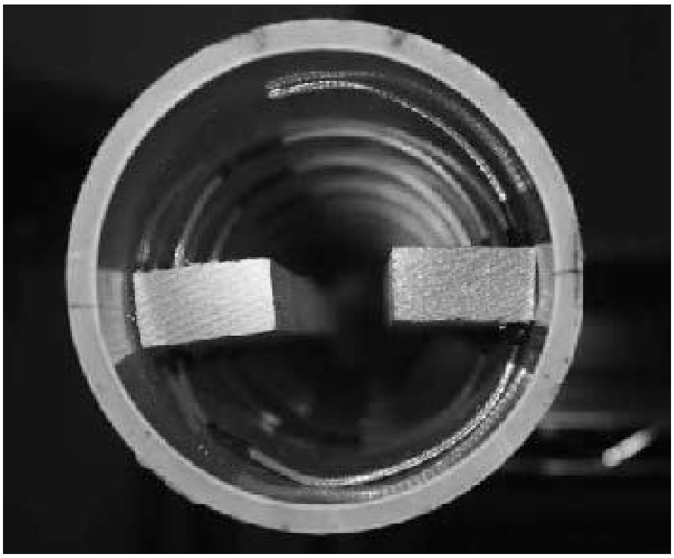

Figure 14 – Interior view of PVC with wood supports for RG316 elements (11 turn helical collinear)

Figure 15 – Completed 902 MHz helical collinear prototype

50Ω. On the other hand, I constructed a single turn "big wheel" (plus ¼ λ end section) with RG-8 coax for 6 meters. This required a 4:1 transformer for good VSWR.

Construction

A PVC radome can be used to enclose high frequency realizations of this antenna. For 902 MHz, 4" tubing works well. At 1296 MHz, a 3" PVC pipe yields good results. The coaxial elements are radiating due to currents on the shield. Since they are each cut to λ/2 in the coax medium, each will be less than λ/2 as a radiating element in free space. The dielectric loading effect of the PVC actually helps mitigate this.

Wood slats were inserted into the PVC tube and screwed to opposite side walls. These wooden slats form supports to attach the coaxial elements. The interior view of the 902 MHz prototype can be seen in Figure 14. The overall completed 902 MHz antenna can be seen in Figure 15.

Table 3
Helical Collinear Calculations 1296 MHz

	(Inches)	*(mm)*			
Length λ/2	4.60	116	(0.35λ)	Elements/Turn	3
Diameter	4.10	104	(0.31λ)	Turns	11
Pitch	4.95	126	(0.38 λ)	Total Segments	264
Linear Total	153.78	3906	(11.65 λ)	N λ/2	32
Helix Length	54.48	1384	(4.13 λ)	Segments/Turn	24
Bottom	5.91	150			
Top	60.39	1534			

Teflon dielectric coax such as RG400 and RG316 should be used because it can withstand soldering temperatures without shorting. The velocity factor is also a bit higher.

Assembly of the 1296 MHz prototype array onto the wooden supports can be seen in Figure 16. The supports were pre-drilled prior to assembly. During assembly, the wooden supports were 'zip-tied' together as shown in the figure. After all elements are arrayed along the support, the ties can be removed and the assembly can be inserted into the PVC tube. Heat shrink tubing was also used to cover and reinforce each junction. For additional mechanical support, short lengths of Tygon or Excelon fuel line tubing can be placed over the heat shrink tubing. For RG-316, I have used tubing with 3/16" OD and 3/32" ID. At 902 and 1296 MHz, it is critical to keep the element to element transitions extremely short. Parasitic inductance can have a significant cumulative effect on performance. Proper element length is also critical. As discussed previously, a

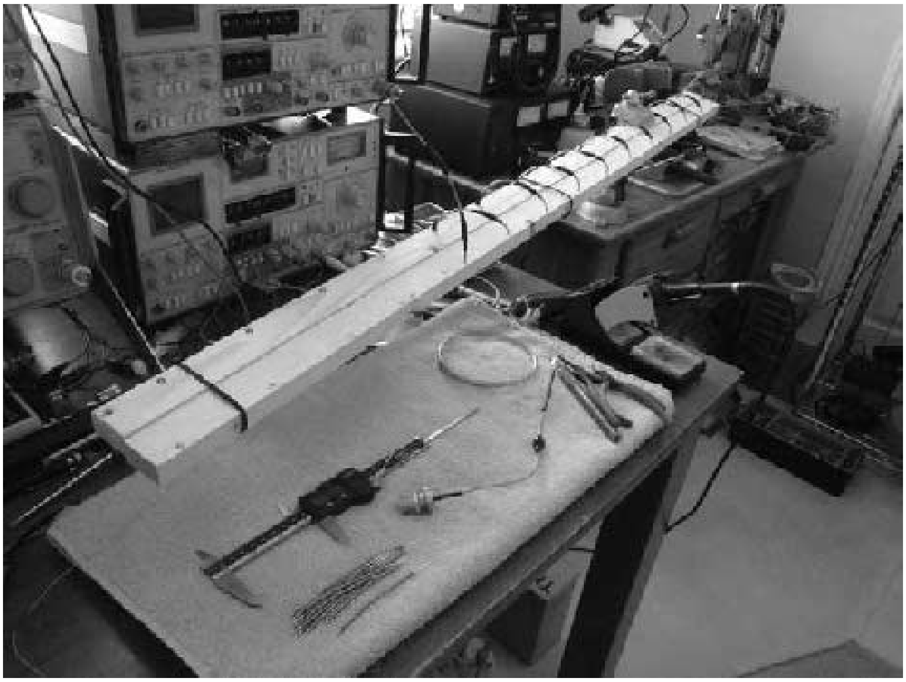

Figure 16 – Assembly of 1296 MHz elements on wood supports

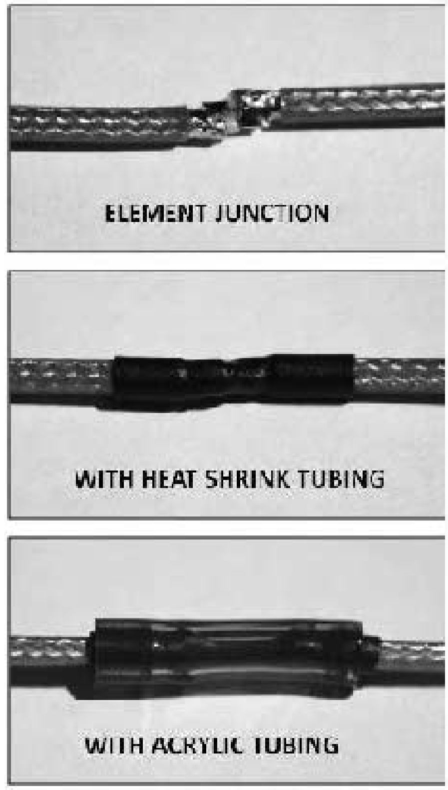

Figure 17 – Construction of element junctions

Figure 18 – Completed 1296 MHz helical collinear prototype

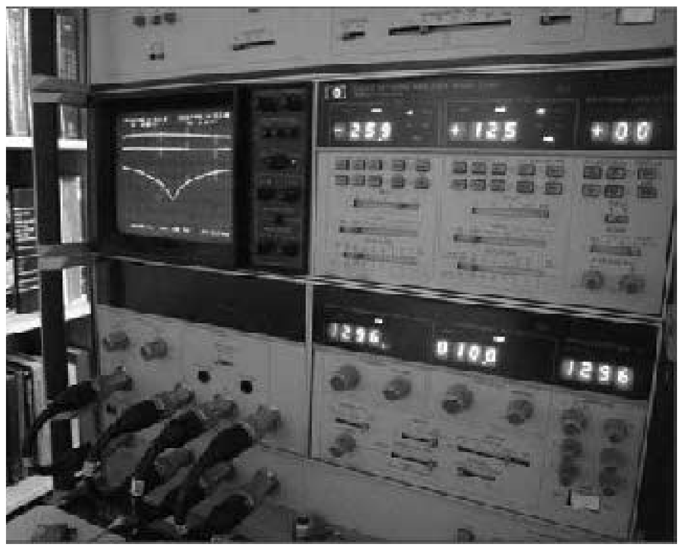

Figure 19 – Measured comparison with 1296 MHz "big wheel"

Table 4
1296 MHz *EZNEC* Port Definition

Element		Linear (Inches)	Linear (mm)	Z-coordinate (mm)	Segments	Bottom Segment
1	$\lambda/4$	143.10	3635	1449	2	178
2	$\lambda/2$	139.90	3553	1420	4	174
3	$\lambda/2$	136.70	3472	1391	4	170
4	$\lambda/2$	133.50	3391	1362	4	166
5	$\lambda/2$	130.30	3310	1333	4	162
6	$\lambda/2$	127.10	3228	1304	4	158
7	$\lambda/2$	123.90	3147	1275	4	154
8	$\lambda/2$	120.70	3066	1245	4	150
9	$\lambda/2$	117.50	2984	1216	4	146
10	$\lambda/2$	114.30	2903	1187	4	142
11	$\lambda/2$	111.10	2822	1158	4	138
12	$\lambda/2$	107.90	2741	1129	4	134
13	$\lambda/2$	104.70	2659	1100	4	130
14	$\lambda/2$	101.50	2578	1071	4	126
15	$\lambda/2$	98.30	2497	1042	4	122
16	$\lambda/2$	95.10	2415	1013	4	118
17	$\lambda/2$	91.90	2334	984	4	114
18	$\lambda/2$	88.70	2253	955	4	110
19	$\lambda/2$	85.50	2172	926	4	106
20	$\lambda/2$	82.30	2090	897	4	102
21	$\lambda/2$	79.10	2009	868	4	98
22	$\lambda/2$	75.90	1928	839	4	94
23	$\lambda/2$	72.70	1846	810	4	90
24	$\lambda/2$	69.50	1765	781	4	86
25	$\lambda/2$	66.30	1684	752	4	82
26	$\lambda/2$	63.10	1603	723	4	78
27	$\lambda/2$	59.90	1521	694	4	74
28	$\lambda/2$	56.70	1440	665	4	70
29	$\lambda/2$	53.50	1359	636	4	66
30	$\lambda/2$	50.30	1278	607	4	62
31	$\lambda/2$	47.10	1196	577	4	58
32	$\lambda/2$	43.90	1115	548	4	54
33	$\lambda/2$	40.70	1034	519	4	50
34	$\lambda/2$	37.50	952	490	4	46
35	$\lambda/2$	34.30	871	461	4	42
36	$\lambda/2$	31.10	790	432	4	38
37	$\lambda/2$	27.90	709	403	4	34
38	$\lambda/2$	24.70	627	374	4	30
39	$\lambda/2$	21.50	546	345	4	26
40	$\lambda/2$	18.30	465	316	4	22
41	$\lambda/2$	15.10	383	287	4	18
42	$\lambda/2$	11.90	302	258	4	14
43	$\lambda/2$	8.70	221	229	4	10
44	$\lambda/2$	5.50	140	200	4	6
45	$\lambda/2$	2.30	58	171	4	2
46	$\lambda/4$	0.00	0	150	2	0

50 mil error in element length can introduce a 2° error in the element length. Elements are best pre-cut using dial or digital calipers. A detailed view of the construction of a junction of elements is shown in Figure 17. For lower frequencies where PVC tubing is not practical in the necessary dimensions, a turnstile support framework is suggested.

The completed 15 turn 1296 MHz antenna is shown in Figure 18. See Tables 3 and 4 for a sample model calculation sheet.

Conclusions

To date, prototype antennas of this type have been constructed and tested for 1296 MHz, 902 MHz, and 50 MHz, although the 50 MHz case was only a single turn. Good results have been obtained in each case. In an attempt to make a measurement of the gain relative to a Big Wheel, a reference path was established between two Big Wheels and then one was replaced by the 15 turn 1296 MHz helical collinear. The network analyzer display of this measurement can be seen in Figure 19. The two horizontal traces on the CRT display the pair of |S21| responses (10 dB/division). I must say that this was not performed on a good antenna range. I am sure reflections were causing errors, so the +12.5 dB gain over a single Big Wheel is likely optimistic. On the other hand, it is safe to say that the omni-directional gain offered from this type of structure is quite good. The ease of feeding the array from a single point is also a very significant advantage.

Tom Apel is an electrical engineer. He retired in 2010 from Triquint Semiconductor as Senior Engineering Fellow where he managed advanced component development. Mr. Apel has 33 years in microwave and RF component design at VHF through Ka band. He developed the first 6-18 GHz 2W power amplifier MMIC to achieve volume production. More recently, his work has resulted in many power amplifier products for handset applications. During his career he was responsible for 34 US Patents. He earned a BS Physics and BS Mathematics from Loras College, and MSEE from University of Wisconsin, Madison. Tom was first licensed in 1963 and has been home brewing since then.

Notes

[1]G.H Brown, "The Turnstile Antenna" *Electronics*, April 1936.
[2]D.W. Masters, "The Super-turnstile Antenna" *Broadcast News*, January 1946.
[3]F.H.Stites,W1MUX/3, "A Halo for Six Meters" *QST*, October 1947.
[4]R.H. Mellen, W1UD,and C.T. Milner,W1FVY, "The Big Wheel on Two" *QST*, September 1961.
[5]L.B. Cebik, W4RNL, and R. Cerreto, WA1FXT, "A New Spin on the Big Wheel" *QST*, March 2008.
[6]*EZNEC* software available from developer Roy Lewallen, W7EL at **www.eznec.com**.

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Remote Rig Operation

A creative approach to remote transceiver control.

My ham radio career started in 1970. My shack was in the corner of the basement. Ever since, my shack has always ended up in the basement if we had a basement to put it in.

We moved into our present home in 2002. As always, my shack went into the basement for lack of anywhere else to put it. Granted, I ended up with a very nicely done corner of the basement, but it was a basement, after all. I wanted to get out of the basement without having to actually move the shack; there would be too many large holes to drill.

It occurred to me, "Why not operate my rig remotely?" My Kenwood TS-2000 had that capability. It talked to an RS-232 port. One small hole between the shack and the bedroom to fish an RS-232 cable and I'd be set -- except I still didn't have audio. It is all well and good to be able to operate the radio remotely, but does little good if I cannot hear it! Okay, so I run speaker and microphone cables, too.

I didn't want to run these analog signals any distance in an RF environment. I have a wired and wireless network in my home. So, why couldn't I use the existing Ethernet network to transmit control and audio?

The 3CX Phone System Management Console showing the settings for the shack (GrandsStream HT286) extension.

Controlling the Rig Remotely

The control side was quite easy. Lantronix makes RS-232 to Ethernet adapters. These come with software to make your computer "think" it is talking to a COM port (virtual COM port, RS-232). So, the software that controls the rig sees a COM port and is happy. The control signals happily zip back and forth over the Ethernet and the rig sees an RS-232 port talking to it. We're half way there.

Audio to and from the Rig, the Biggest Challenge

I had the same problem, though… no audio. My original plan was to build a full duplex audio interface that talks RS-232 and utilizes the second port of my Lantronix UDS2100 RS-232-to-Ethernet adapter. This would require software expertise on the computer end to talk RS-232 to send and receive audio from the sound card to another virtual COM port. I didn't have that expertise. The project stalled….at least until I discovered *Voice Over IP* (VOIP) technology.

This technology has been around a long time. Magic Jack and Vonage are two common carriers who use VOIP to provide telephone service. But, how do I use this technology in my application?

Getting Audio to and from the Ethernet

GrandStream (among others, I'm sure) have Analog Telephone Adapters (ATA) that interface between standard VOIP network and any analog telephone. I purchased a GrandStream HT 286 for under $30 off the Internet. This requires that I have a VOIP network for it to talk to. How do I get a VOIP network in my house? This was *way* easier than I thought.

Setting Up a VOIP Network

3CX has free, downloadable software to do the job (**www.3cx.com/phone-system/index.html**). One application is the VOIP system that resides on a computer on your network.

This was a little daunting to me as I was sailing into unknown waters. I had help from Matthew Orr of InfoSys Consulting, Inc. (**www.InfoSysHelp.com**) in getting this set up. My personal thanks to him; he was a great resource. Later in this article I'll describe how to go about configuring the GrandStream HT286 with the VOIP system.

Turning Your Computer Into a Phone

The second needed software that you can download free from 3CX is the *Softphone* (**www.3cx.com/VOIP/voip-phone.html**). It is an application that sets up a VOIP telephone on your computer. You use the *SoftPhone* to call the GrandStream ATA down in the shack. With this *SoftPhone* loaded on every computer in the house, you can make calls between computers (each computer becomes an extension with video phone capabilities) and you can connect to your rig's audio from any computer! I'll describe later how to set up *SoftPhone*.

Interfacing the ATA with the Rig's Audio + More Control

We're mostly there, but not quite. I have the audio in the shack now in the form of an analog telephone signal. How do I get it into and out of the rig? Most of you are probably ahead of me on this. If you are thinking "phone patch," you are 100% correct. But, this phone patch has to have certain functionality to make it convenient for this application. As a minimum it needs to:

• Answer the phone and connect to the rig automatically … like an answering machine does after a set number of rings

• Allow you to hang up the extension remotely

For even more convenience, I'd like it to allow me to turn on the power to the rig's power supply and connect the antennas to the rig.

If you are thinking "ring detector and counter" to take care of the first requirement, you got that right. If you are thinking "touch tone decoder" for the second requirement and the added extra, you got that right, too.

So, that is what I did. I built a phone patch with a ring detector and counter so that it would automatically answer the phone after four rings. I added a touch tone/DTMF decoder to provide control. The DTMF decoder I chose gives me the potential of controlling up to 16 devices with a single key press for each. Of course, one of those is taken up with the hang up functionality and another for the power up functionality. You are still left with 14 more things you can do remotely. These lie fallow in my design, for the moment.

The downside of the DTMF decoder is that it doesn't work when there is any amount of audio on the telephone line. So, it has to be quiet on the line to use this. Switching antennas with the din of a pile-up going on just won't happen. Mute rig, switch antennas, un-mute rig.

One More Degree of Freedom

Now, to give myself yet one more degree of freedom, I purchased a Bluetooth hands-free headset and a USB Bluetooth interface. Now I do not need a wired microphone. The *SoftPhone* can be configured to use the PC speakers for its speakers and the Bluetooth headset for its microphone. The whole system looks something like the diagram shown in Figure 1.

The Phone Patch Details

Back in the early 1970s I built a rudimentary phone patch. Believe it or not, I kept the guts of the patch all these years, stuffed away on a box full of Styrofoam noodles. When I started looking for more technical information for my new patch, I was amazed at who didn't have it. I did find what I was looking for, but it wasn't where I expected it to be. In the end, it wasn't rocket science and I didn't expect it to be. After all, I was just 16 when I built my first one, but it didn't have the needed bells and whistles that this one needed to have.

Let's walk through the schematic in Figure 2. The basic phone patch portion of the schematic is shaded in gray. The relay contacts that are connected across C15 may be replaced with a SPST switch. The purpose for the diodes included in the design across the speaker and microphone is to prevent large voltage swings in the event of the ring signal. They are there for protection of the transceiver. Please note that the pin numbers that appear on the transformer *may* not be the pin numbers that apply to the transformer you choose to use. The transformer is a 600 Ω primary connected to the telephone line. The speaker winding is a 600 Ω secondary. The microphone winding is a 150 Ω secondary. The primary/secondary that is used for the ring detector is a 150 Ω winding.

The Ring Detector

The ring signal is an ac signal. There are a number of ways I could have done this, but

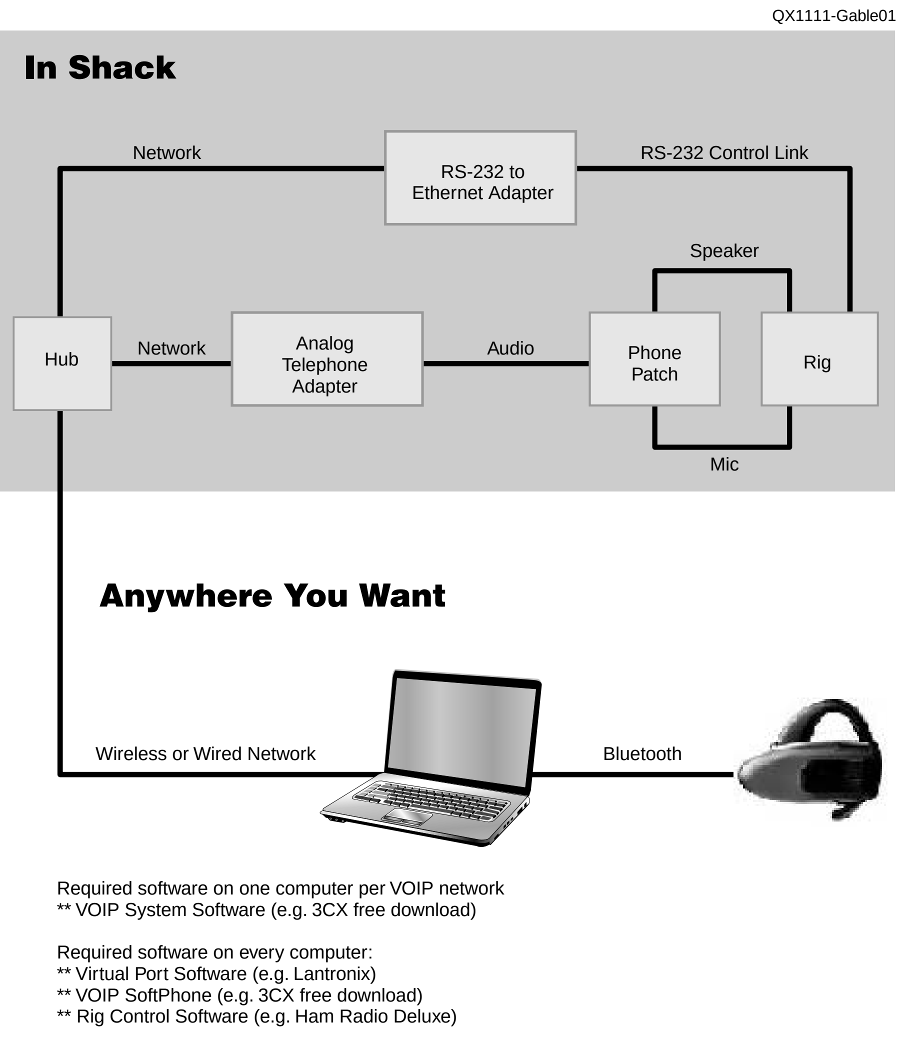


Figure 1 – System diagram.

QX1111-Gable02

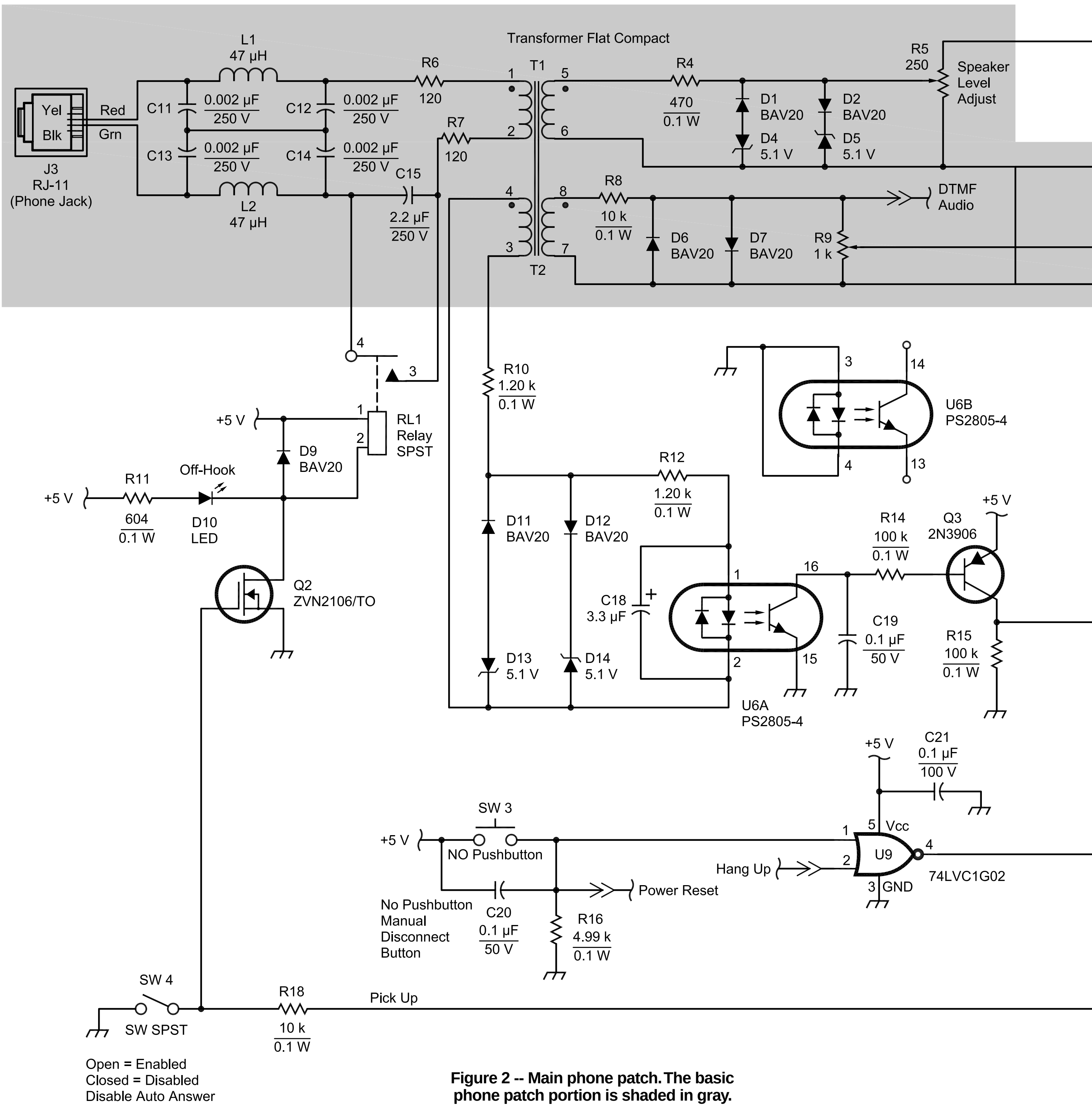
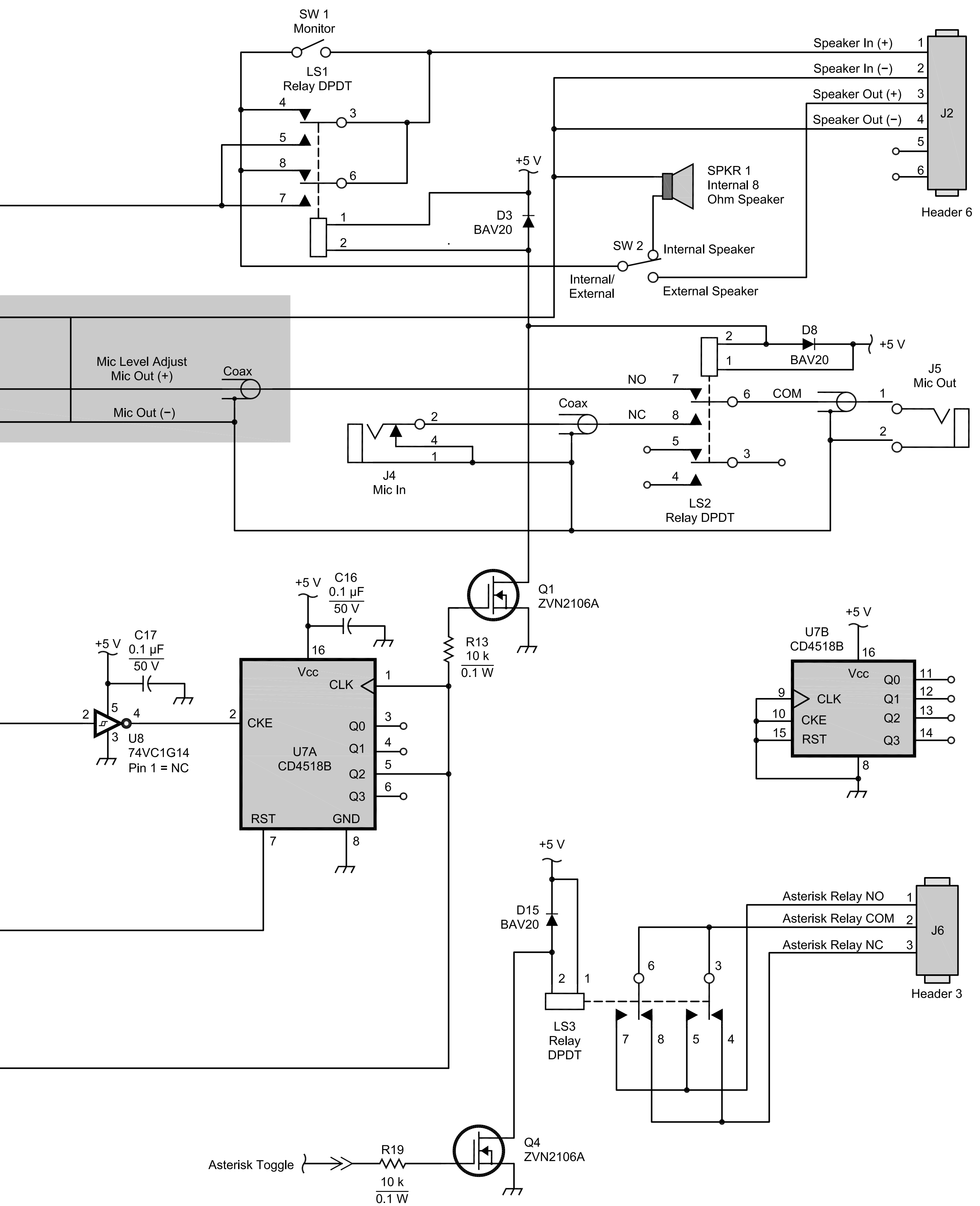


Figure 2 -- Main phone patch. The basic phone patch portion is shaded in gray.


SW 1
Monitor
LS1
Relay DPDT
Speaker In (+)
Speaker In (−)
Speaker Out (+)
Speaker Out (−)
J2
Header 6
+5 V
SPKR 1
Internal 8
Ohm Speaker
D3
BAV20
SW 2
Internal Speaker
Internal/
External
External Speaker
D8
BAV20
J5
Mic Out
Mic Level Adjust
Mic Out (+)
Coax
Mic Out (−)
NO
NC
COM
J4
Mic In
LS2
Relay DPDT
Q1
ZVN2106A
C16
0.1 μF
50 V
C17
0.1 μF
50 V
R13
10 k
0.1 W
U7B
CD4518B
Vcc
CLK
CKE
RST
Q0
Q1
Q2
Q3
U8
74VC1G14
Pin 1 = NC
U7A
CD4518B
GND
D15
BAV20
Asterisk Relay NO
Asterisk Relay COM
Asterisk Relay NC
J6
Header 3
LS3
Relay
DPDT
Asterisk Toggle
R19
10 k
0.1 W
Q4
ZVN2106A

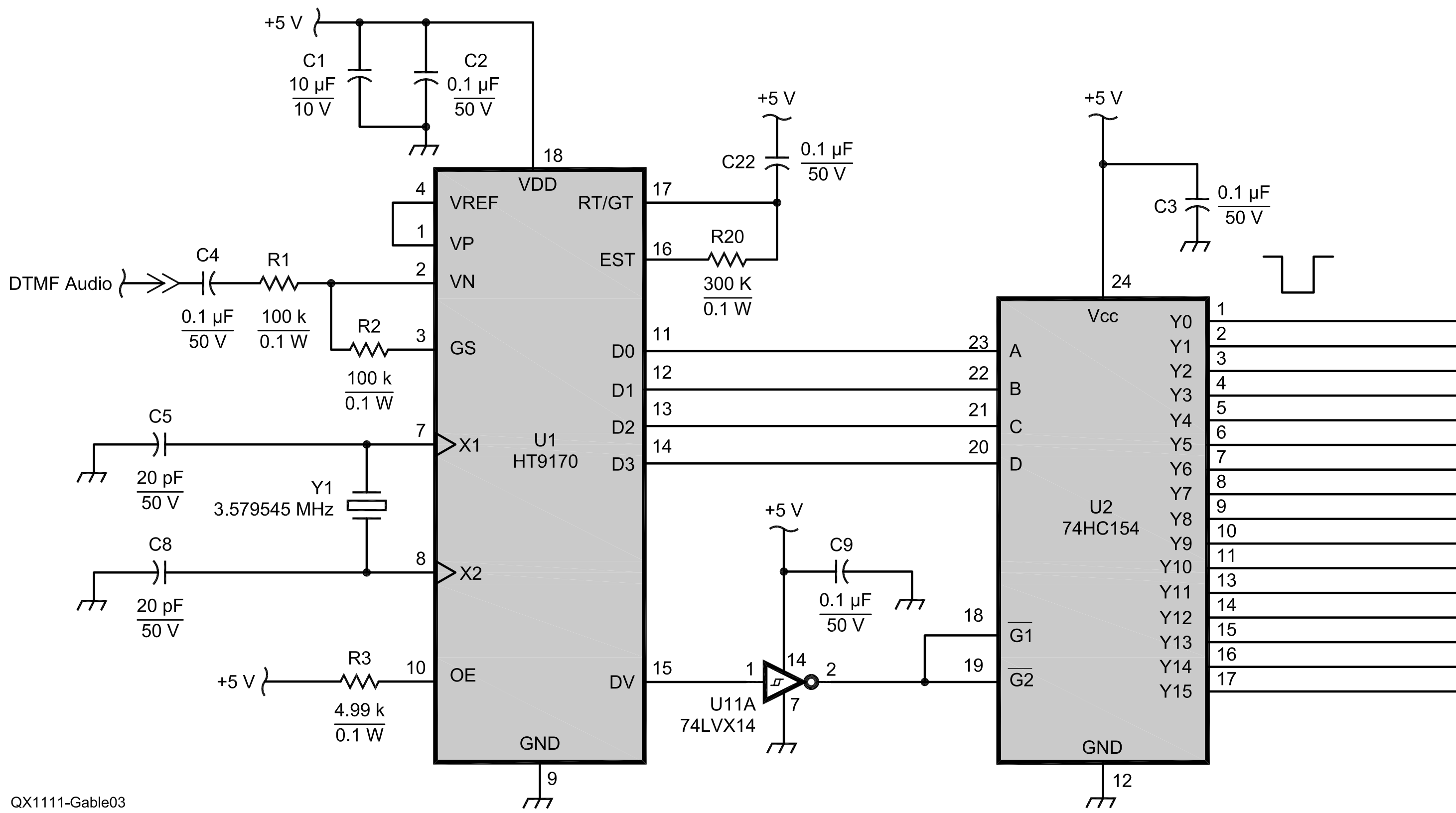
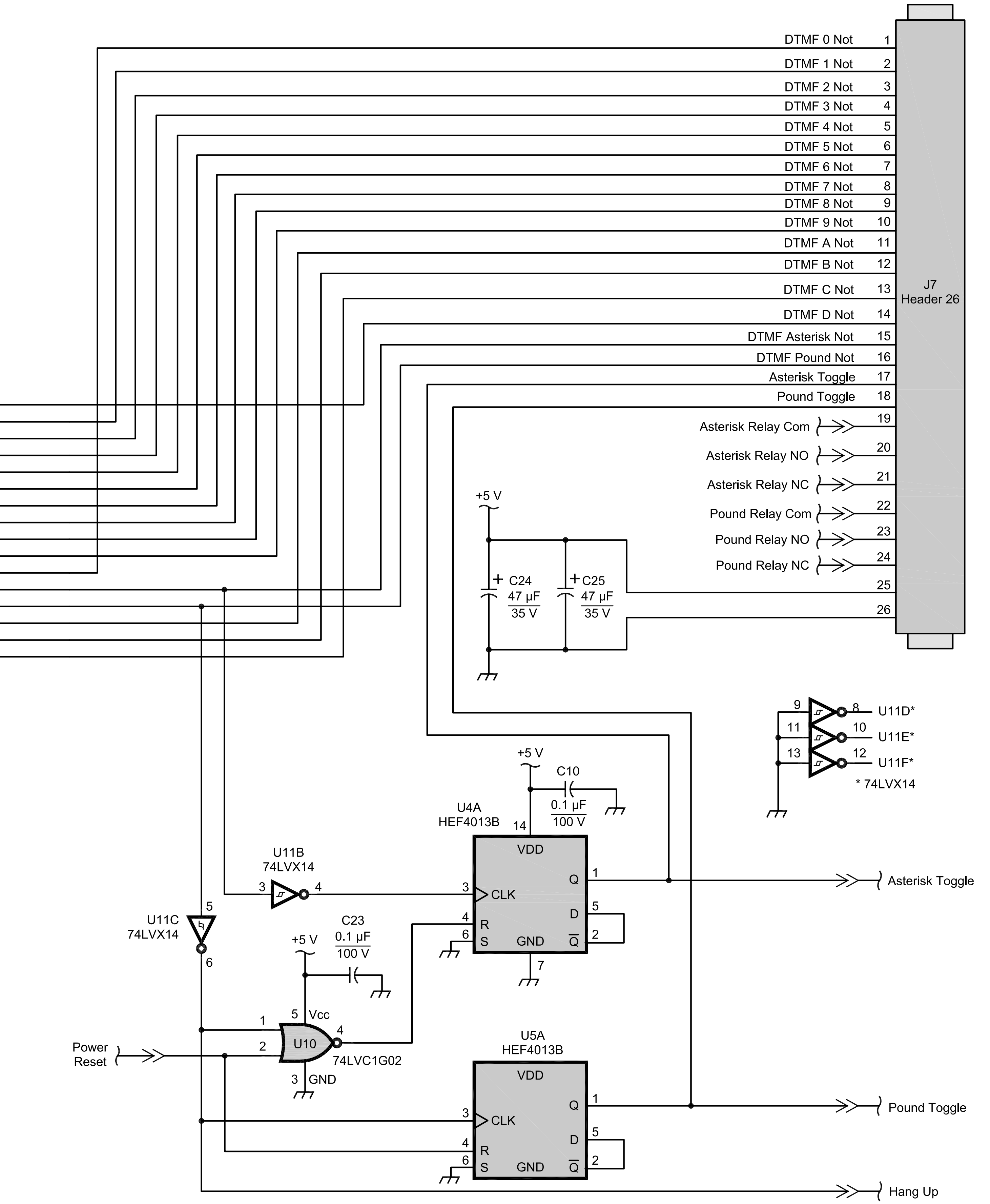


Figure 3 — DTMF Decoder and Control.


DTMF 0 Not 1
DTMF 1 Not 2
DTMF 2 Not 3
DTMF 3 Not 4
DTMF 4 Not 5
DTMF 5 Not 6
DTMF 6 Not 7
DTMF 7 Not 8
DTMF 8 Not 9
DTMF 9 Not 10
DTMF A Not 11
DTMF B Not 12
DTMF C Not 13
DTMF D Not 14
DTMF Asterisk Not 15
DTMF Pound Not 16
Asterisk Toggle 17
Pound Toggle 18
Asterisk Relay Com 19
Asterisk Relay NO 20
Asterisk Relay NC 21
Pound Relay Com 22
Pound Relay NO 23
Pound Relay NC 24
25
26
J7
Header 26
+5 V
C24
47 μF
35 V
C25
47 μF
35 V
U11D*
U11E*
U11F*
* 74LVX14
+5 V
C10
0.1 μF
100 V
U4A
HEF4013B
VDD
CLK
R
S
GND
Q
D
Q
U11B
74LVX14
U11C
74LVX14
C23
0.1 μF
100 V
+5 V
Vcc
U10
74LVC1G02
GND
Power
Reset
U5A
HEF4013B
VDD
CLK
R
S
GND
Q
D
Q
Asterisk Toggle
Pound Toggle
Hang Up

I settled on an ac input opto-isolator. The ac ring signal coming from the transformer goes through a current limiting resistor to a diode circuit (clipper) designed to limit the voltage to be applied to the opto-isolator circuit to no more than the zener voltage plus the forward voltage drop of the other diode that is in series. This is still an ac signal.

The next step is a current limiting resistor to the input of the opto-isolator (PS2805-4). The selection of the limiting resistor is based on the forward drop of the input of the opto-isolator and the current you want to push through it. I added a 3.3 µF capacitor to turn this same circuit into a low pass filter. This way stray clicks and the like that may appear on the telephone line do not get counted as a bona fide ring signal.

The output side of the opto-isolator is an open collector with a following common emitter amplifier (2N3906). The object is to make the rise and fall time of the incoming signal as quick as we can. Digital gates are not real happy with slow slew rate signals. To finish the processing, I stuck a Schmitt trigger inverter between the opto-isolator output and the counter input; this finishes squaring the edges up. Now we have a nice, truly digital signal to apply to the counter.

To follow the signal a little bit just for clarity of understanding…

• A ring signal occurs

• The input side of the opto-isolator conducts

• The output side of the opto-isolator conducts and its collector goes low

• This pulls the base of the 2N3906 low causing it to conduct, connecting the collector to Vcc ($V_{COLLECTOR}$ = Vcc – ≈0.2 V = ≈4.8 V). The collector goes high.

• When the ring signal stops, the collector of the 2N3906 drops low again.

The Ring Counter

Admittedly, the counter circuit *does* look a bit strange. Thinking this through, I wanted the counter to stop counting after it reached a count of four. That means that bit Q2 would go high. I had to come up with a way to do this. The CD4518 has a Clock Enable input. If the Clock Enable is high, then positive transitions on the clock cause it to increment. On the other hand, if the clock input is low, then negative transitions on the Enable input cause it to increment. If the clock is held high, then transitions on the Enable do nothing. That is *just* what the doctor ordered.

I tied the counter's Q2 (bit 2 of 0 through 3) output to the CLK input and the output of the opto circuit to the Enable input of the counter. Now the counter increments a count at the cessation of each telephone ring. Once it reaches a count of four, then it quits counting. The only thing that will allow it to count again is a reset.

The counter is reset either by the PWR_RESET signal or the "#" output (HANG_UP) of the DTMF decoder. To reset the counter is to hang up the phone.

The output of the counter goes high when the phone patch is supposed to answer the phone. This drives three relay driver inputs. The first (RL1) connects the phone patch to the phone line, answering the phone. The second (LS1) connects the rig's speaker output to the audio input of the patch. The third (LS2) connects the audio output of the patch to the microphone input of the rig.

Power Up Reset

The power up reset is nothing more than an RC network whose output goes high at power up and then drifts low as the capacitor charges. The pushbutton across the capacitor is the reset button for manual reset. It pulls the PWR_RESET signal high and discharges the capacitor. Once released, the PWR_RESET signal drifts low again just as if the power were just turned on.

DTMF Decoder

The DTMF, or touch-tone, decoder is an off the shelf DTMF receiver, an HT9170 which is available through Newark Electronics. See Figure 3. There is nothing special about this circuit. I took the circuit right out of Application 1 on page 8 of the datasheet.

One of the outputs of the DTMF receiver is the DV output which goes high when a valid DTMF tone pair is detected. More on this later.

The HT9170 presents the decoded DTMF tones as four bit binary numbers. To make this truly useful, this has to be further decoded. I chose a 4:16 decoder/demultiplexer, the 74HC154. It is a simple matter of connecting the four bits coming out of the DTMF receiver to the four bits of the 4:16 decoder. The decoder also has enable inputs. The output of this decoder will not be asserted unless these enable inputs are both low. So, we invert the DV signal from the DTMF receiver to give us an active low for these inputs. This prevents the output of the decoder from being asserted unless the DV signal is asserted (valid DTMF tone pair received by the receiver).

Table 1
DTMF Controls

Output	*DTMF*	*Output*	*DTMF*
Y0	D	Y8	8
Y1	1	Y9	9
Y2	2	Y10	0
Y3	3	Y11	*
Y4	4	Y12	#
Y5	5	Y13	A
Y6	6	Y14	B
Y7	7	Y15	C

We now have 16 fully decoded controls to be used. They are defined in Table 1.

Control Circuitry

The output of the 74HC154 is negatively asserted. That means that the output goes low when it is true. I chose to use the "#" to remotely hang up the phone patch and the "*" to toggle auxiliary relay contacts on and off, which I plan on using to control the ac power to the power supply that powers the rig.

The first step of this control is to invert the output of the decoder that I have chosen to use. The "DTMF_Pound_not" becomes "Hang_Up," which is positively asserted when the user wants to disconnect the phone patch from the ATA by pressing the "#" key on their *Softphone*.

The "DTMF_Asterisk_not" becomes the positively asserted "DTMF_Asterisk." This is used to drive the clock input of a D Flip Flop. The flip-flop has its Q_not output connected to its D input so that the Q output toggles at each clock edge it sees.

To make sure that it comes up in the right state, the PWR_Reset signal is used to reset it to the Q = low state at power up. Furthermore, we want it to reset whenever the phone patch is not connected to the ATA. So, I use an OR gate with the Pwr_Reset signal on one input and the "Hang_Up" signal on the other. The OR gate's output becomes the "Asterisk_Reset."

The Q output of the flip-flop serves as the input of the relay driver. When Q is high, then the relay is engaged. When it is low, it is disengaged.

All of the other outputs from the 4:16 decoder just sit there waiting to feel useful. And, someday, I may just help them to feel that way. For the time being, they will just have to wait.

Station Control

I chose to use the Asterisk relay contacts to control a relay that turns on the primary power to the rig's power supply. The output of the rig's power supply (12 Vdc) supplies power to my Kenwood TS-2000. It also activates the antenna relays. These relays connect the antennas to ground when not activated and to the rig when activated. These relays are 10 kW RF relays I bought from Vector Solutions (**www.arraysolutions.com/Products/rf_relays.htm**). I mounted them in an EMI shielded aluminum box from Bud (AN-1322). To be a little more cautious, I also added additional lightning arrestors to the antenna lead in cables.

How to Set Up the VOIP System and It's Peripherals

The first step is making your PC a "static IP" box. To do this, follow the following steps for *Windows XP*:

• Open a "Command Prompt" window.

• Type "ipconfig /all" and press enter.

• Note the following entries:

• IP Address: __________.__________.__
____

• Subnet Mask: __________.__________
____

• Default Gateway: __________.________
____

• DNS Servers:
__________.__________.__________.

__________.__________.__________.

• Click "Start" and then "Control Panel"

• Double click on "Network Connections"

• Right click on the "Local Area Connection" icon that represents your LAN connection and then click on the "Properties" pop-up menu item.

• Click on the "Internet Protocol (TCP/IP)" entry in the "This connection uses the following items:" area and then click on the [Properties] button.

• The "Internet Protocol (TCP/IP) Properties" dialogue box opens. Click on the "General" tab. You will most likely see the "Obtain an IP address automatically" radio button selected.

• Select the "Use the following IP Address" radio button.

In the IP address, enter the IP address that you got from the Command Prompt window *except* the last number. Choose a new number for this that is at least 20 higher than the number you got from the Command Prompt window. This makes sure that other computers/devices that connect to your home network that "obtain an IP address automatically" will not be assigned the IP address of this computer if they do so when your computer is off at the time they connect. Otherwise, you will get an IP address collision on your network and this is not pretty.

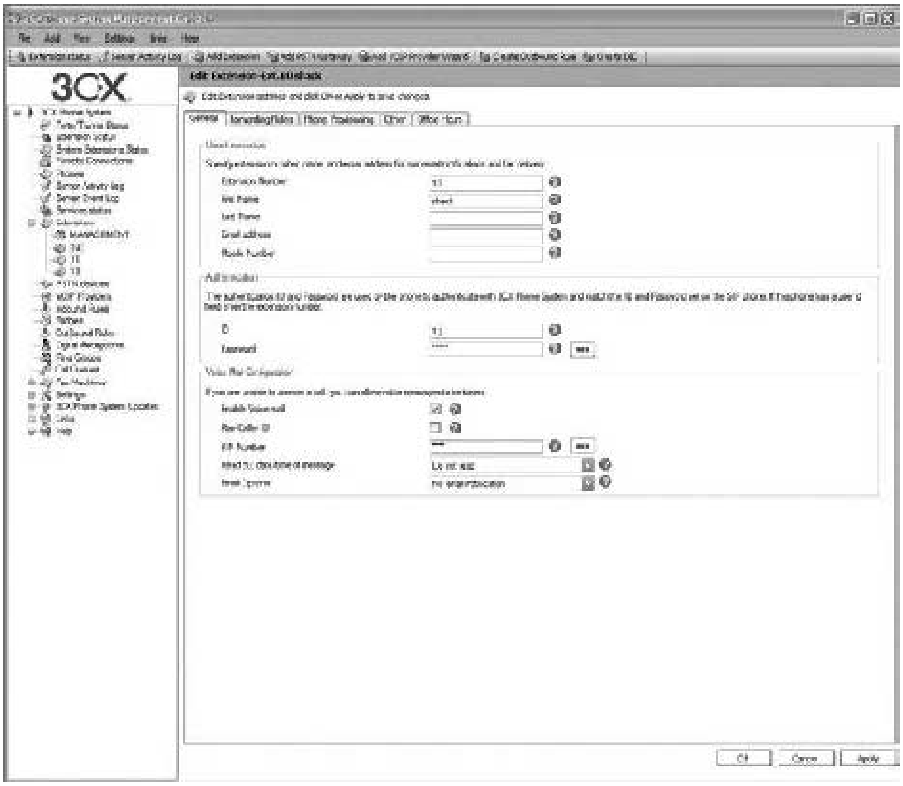
• Click on [OK] and [OK] and restart your computer.

• Install the 3CX PBX Phone System from **www.3cx.com/phone-system/download-phone-system.html**

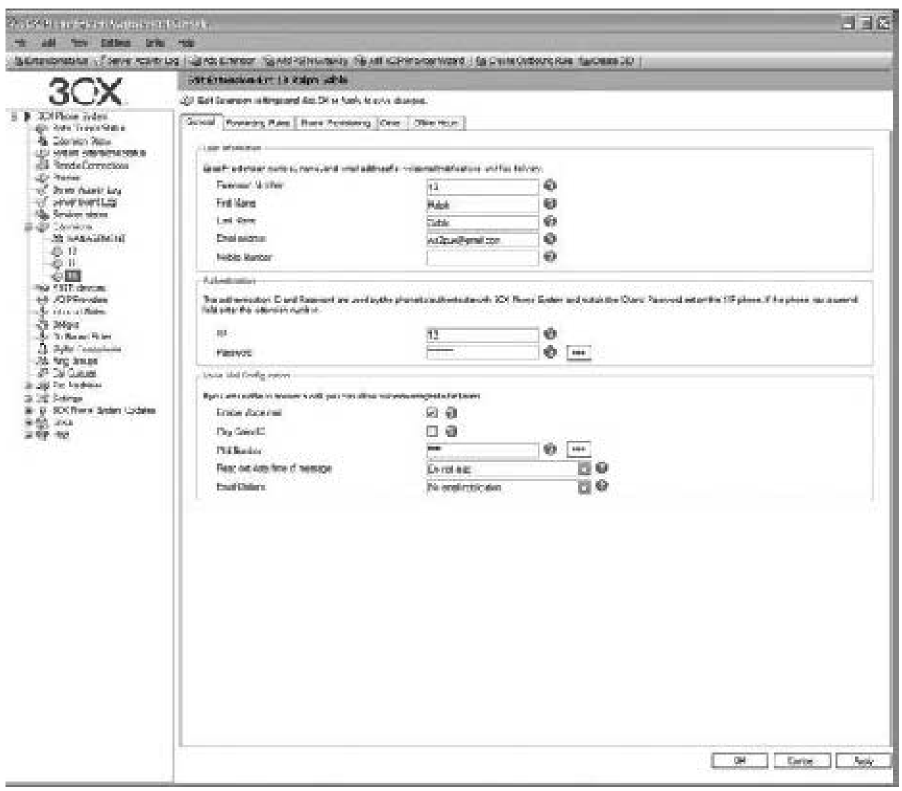
• When asked for the number of digits in the extension, I put two.

• They will eventually ask you for adding extension information. I used the following:

Extension Number: 10
First Name: shack
Last Name: ---left blank---
Email address: ---left blank---
Mobile Number: ---left blank---
ID: 10
Password: 10



The 3CX Phone System Management Console showing the settings for the computer (3CX Phone) extension.



The author's remote control system package.

Add another extension for the 3CX phone for the computer. I used the following:

Extension Number: 13
First Name: Ralph
Last Name: Gable
Email address: ---left blank---
Mobile Number: ---left blank---
ID: 13
Password: 13

• Complete the installation of the 3CX PBX Phone System.

• Install the 3CX Phone for the computer that you downloaded from: **www.3cx.com/downloads/3CXPhone6.msi**

• Create a new account on the 3CX *SoftPhone*.

Account Name: ---anything you want to call it----
Caller ID: 13
Extension: 13
ID: 13
Password: 13
My Location:
I am in the office
Local IP IP address chosen above of PBX

• Click on [OK] and then [OK] again.

Not connected….reading configuration…on hook

• Now for the Grandstream HT286

• Locate and record the MAC address of the HT286 (usually on the label on the bottom)

• Connect the HT286 to an analog telephone.

• Connect the HT286 to power with its adapter.

• The following steps come directly from an application note found on the 3CX site (**www.3cx.com/sip-phones/GrandStream-HandyTone286.html).** "Configuring GrandStream HandyTone 486(487),286(287), ATA for 3CX Phone System"

• Reset it to factory defaults.

• Pick up the headset and press "****". This will start up a Voice Prompt Menu. Now press "99". Dial the MAC of the device, where …

a. A=22
b. B=222
c. C=2222
d. D=33
e. E=333
f. F=3333

For example, if the MAC address is 000b8200e395, it should be encoded as "0002228200333395"

• Setup the HT286 to work with the 3CX PBX system

• Connect the LAN to the Wan port of the HandyTone

• Connect an analog phone to the HandyTone phone port. Pick up the headset and press "****". This will start up a Voice Prompt Menu. Now press "02" to listen to the IP address that was assigned to the device by the DHCP server

• Enable the wan side Web access, dial "****" and then "129"

• Launch a browser and go to the IP Address determined or configured earlier. The default username is "admin"

• Select the "Advanced Settings 1" tab.

• Set the "SIP Server" field to the IP Address or FQDN of the server on which 3CX Phone System is installed – in this example 10.172.0.2

• Set the "Outbound Proxy" field to the same value as in step 13 above.

• The "SIP User ID" field should match the "Extension Number" field of the extension created for this phone in the 3CX Phone System Management Console

• In the "Authenticate ID" and "Authenticate Password" fields enter the ID and Password that you entered for the extension in the 3CX Phone System Management Console. These fields must match the Authentication ID and Password set for that extension in the 3CX Phone System Management Console

• The "Name" field is optional. A suitable value would be the name of the user using this phone

• Set the "Preferred Vocoder" to choice 1: PCMU, choice 2: PCMA. The settings for choices 3 to 7 will not come in use since they will not be used by 3CX Phone System

• Set "User ID is phone number" to "Yes"

• Set "Sip Registration" to "Yes"

• Set "Unregister on Reboot" to "Yes"

• Set the "Register Expiration" field to a suitable value. For testing purposes you may want to use 60 seconds, but once configuration is tested a larger value would be more appropriate to limit unnecessary network traffic. A good setting for general use could be 3600 seconds (1 hour).

• Set "Allow outgoing call without Registration" to yes.

• Set "Enable Call Features" to "Yes". This setting enables features like transfer, on hold, etc from the analog phone connected to the FXS port of the device.

• Set "Send DTMF" to "in audio" or "Via RTP (RFC 2833)" or both. SIP info is not recommended although 3CX Phone System can support it.

• Scroll to the bottom of the page and click the "Update" button. You will be prompted to reboot the device. Click the "Reboot" button.

• After the HandyTone has restarted, switch to the 3CX Phone System Management Console, and click on the Phone System -> "Line Status" (the default page). In the section "Extensions", your new extension connected to the PBX should be listed with a green status light.

At this point you should be good to go.

Ralph Gable is an ARRL member originally licensed in 1970 as WN2PUX, a Novice class operator. The following year he upgraded to General class and received the WA2PUX call sign. Ralph has been active to varying degrees over the years and has lived in a wide variety of locations. He received his Canadian reciprocal license (VE4PUX) when he lived in Canada for a number of years. During that time, Ralph was published in The Canadian Amateur *magazine and was a volunteer examiner for the Canadian version of the FCC. He upgraded to an Amateur Extra license since moving back to the US. Ralph is presently working as a Senior Electrical Engineer for TomoTherapy, a wholly owned subsidiary of Accuray Incorporated (newly acquired), in Madison, WI. TomoTherapy designs and manufactures state of the art Radiation Oncology machines.*

Ian Cowan, VK1BG

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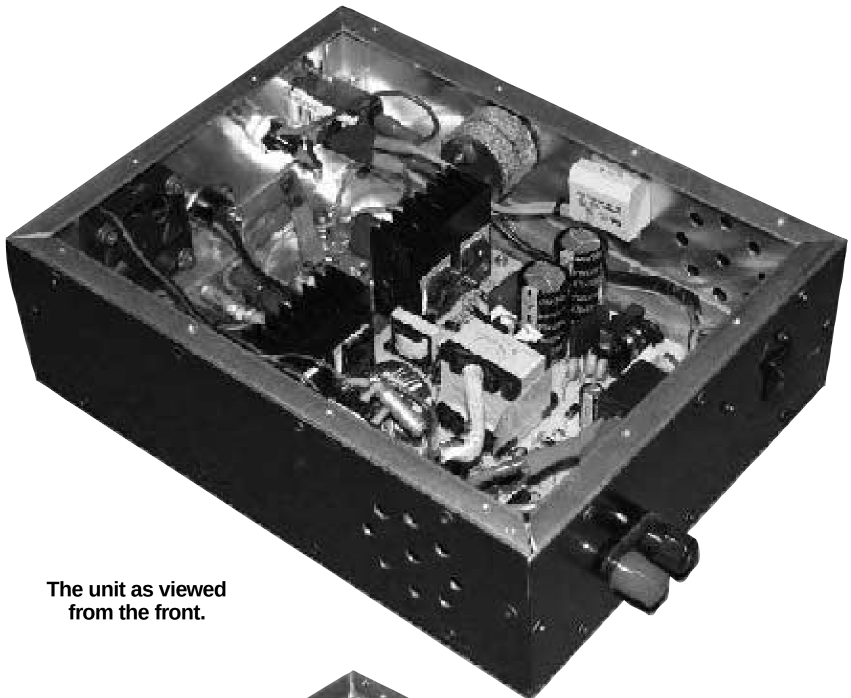
Taming the SMPS Beast

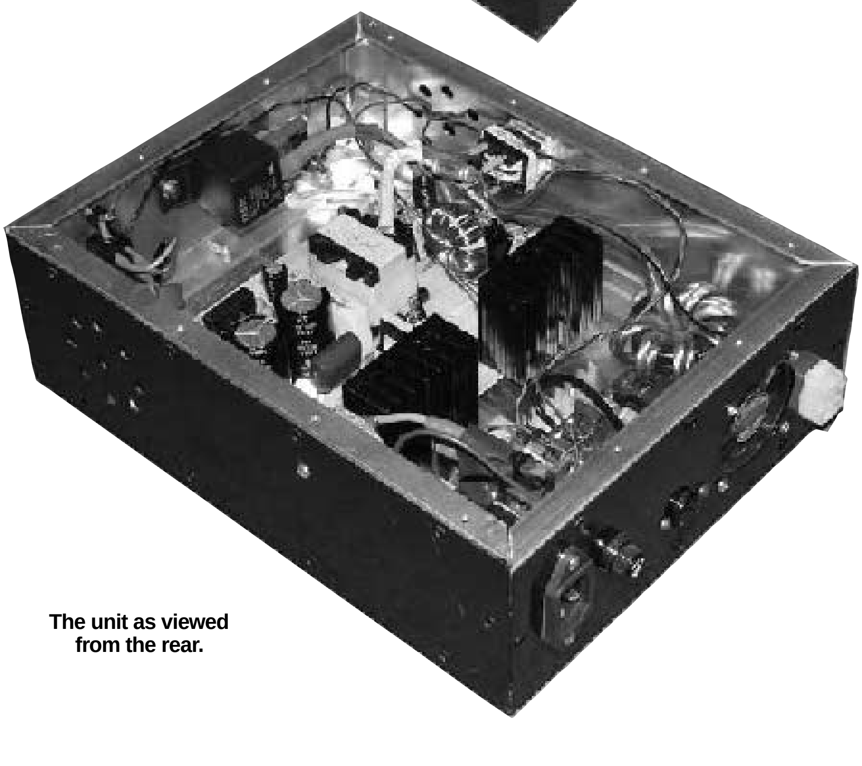
Switch mode power supplies tend to be too noisy on MF and HF for many radio environments. This author has devised a cure.

This article was originally published in Amateur Radio *magazine, the journal of the Wireless Institute of Australia, and is reprinted with their permission.*

In the March 2008 issue of *Amateur Radio*, Drew Diamond, VK3XU, described a method of reducing the RF noise emission from a low cost commercially available switch mode power supply (SMPS). Drew commented that his technique considerably reduced the strength of the noise. The level of noise suppression achieved by Drew would be adequate for most purposes, but falls short of the needs of the most demanding environments.



The unit as viewed from the front.



The unit as viewed from the rear.

Introduction

I have a trailer with a broadcast receiver installed which is frequently used for listening to distant broadcast stations such as 3WV on 594 kHz. The radio has hitherto been powered by a quiet linear power supply. The radio works well provided the ambient noise level is low.

A couple of years ago I saw advertised in the Jaycar catalog a small SMPS rated to deliver 13.8 V at 20 amps. In spite of its much higher output rating, it is actually lighter than the linear supply. As a bonus, the SMPS also seemed to be about right to power my FT-897 transceiver, which I sometimes take away with me, and which I had until then powered from a lash-up supply to keep the weight down.

I hot footed it to our local store and talked to the man there about it. I asked about the noise levels generated by the unit, but he would offer no guarantees about the suitability of the SMPS for my purpose. Fair enough. In due course I decided to take the plunge and bought one. It turned out that the SMPS matched my worst fears. There were birdies right across the MF and HF spec-

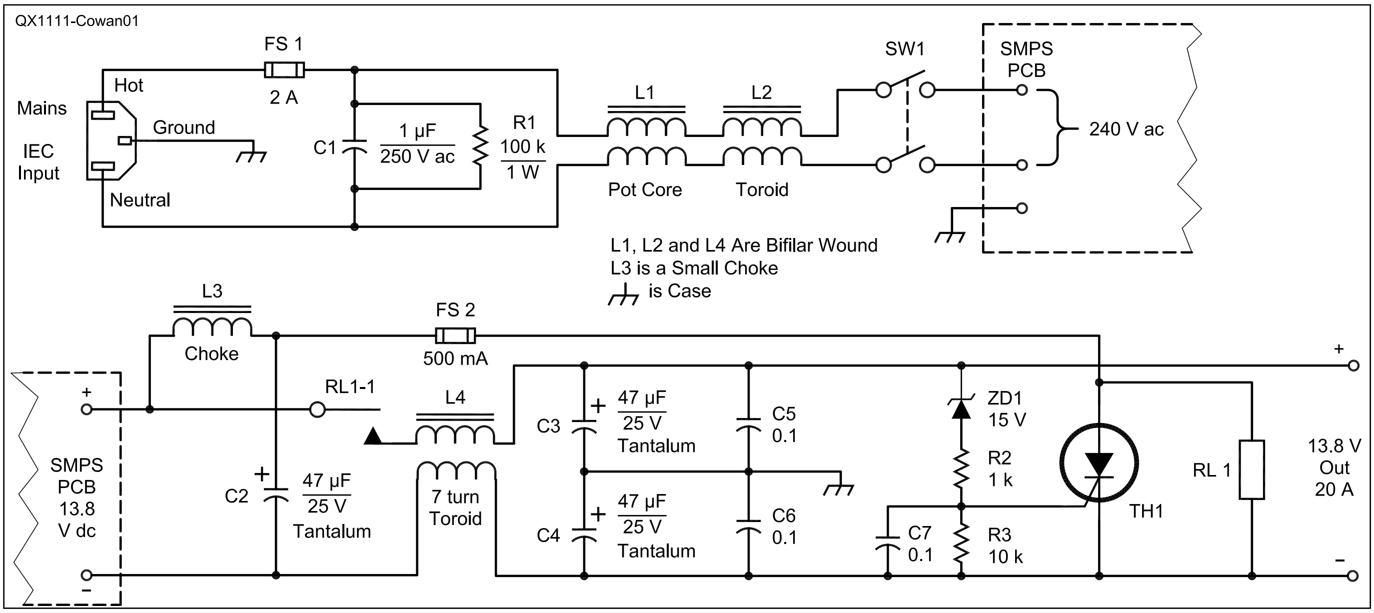


Figure 1 — The circuit diagram.

trum, so the unit was useless for my purpose. However, in other respects it seemed fine, as it delivered rated output with good regulation and no sign of overheating. It has a built-in fan and this operates in an unusual manner in that the fan speed seems to be modulated by the output load current. Slow for light loads, fast at 20 amps. And it is quiet.

So I decided to put some effort into getting rid of the RFI.

Warning: Switch Mode Power Supplies [SMPS] Can Carry Lethal Voltages

In addition to dangerous voltages, equipment using SMPSs regularly use "hot chassis" construction. That is, the chassis operates at a high voltage and so is not grounded. These power supplies should not be worked on without protective wear, both electrical and physical. Components have been known to shatter during testing. Precautions require that under no circumstances should it be possible to touch or be in contact with a working supply; especially by accident. The SMPS is not a beginner's project.

A Little Theory

SMPS power supplies are by nature prolific generators of RF noise and the reason for this is not hard to see. A typical small cheap SMPS delivers the ac input direct to a high voltage bridge rectifier via some rudimentary filtering. The bridge rectifier then charges a large electrolytic capacitor to something like 340 V dc for units used in Australia. Also connected to this electrolytic is a fast power switch – usually a power MOSFET – that switches at a frequency upwards of 50 kHz. This switch operates at a short mark-to-space ratio (that is, the "on" time is much less than the "off" time) so that the average value of the output of the switch is much less than the voltage across the main electrolytic. This very spikey waveform then passes to a ferrite core transformer for processing into regulated dc, and that is where we will leave it.

The main bridge rectifier operates from the 50 Hz ac line, so it moves into conduction 100 times every second. During the "on" time of the rectifier the main electrolytic is connected direct to the mains input via an effective resistance of only a few ohms. While this is happening the high speed switch is also belting away at a much higher frequency – its operation modulates the voltage across the electrolytic and this modulation influence also appears across the mains for as long as the input rectifier is in conduction. Thus the ac mains supply becomes the bearer of a nasty, complex RF waveform which is rich in the harmonics of the mains frequency, the high speed switch frequency, and the intermodulation products of all these frequencies.

Now you may think that the main electrolytic, which might be rated at 470 μF or so, would be big enough to prevent the dc voltage across it from being modulated by the high frequency switch. Unfortunately, these units typically have high internal resistance to high frequencies, so they are pretty useless as filters.

From the above it can be seen that cheap switch mode power supplies tend to noise modulate the voltage between the active and neutral of the incoming ac lines, and this form of noise is known as normal mode noise. This is the most copious form of noise coming from most cheap units. It is also the hardest to deal with as it appears in a lethally dangerous environment.

The dc output from the SMPS is derived from a high frequency rectifier, which is followed by a simple filter arrangement of limited efficiency. Thus there is also plenty of residual normal mode noise between the positive and negative output wires.

A second form of noise – known as common mode noise – is also conspicuous in cheap SMPS units. This is noise transferred by stray capacitive and inductive coupling into the mains and output circuits. With a little reflection it is realized that the RF power level of the fast switch must be quite high, so cross coupling of significant levels of noise is readily achieved. Common mode noise appears across both wires of the incoming mains more or less equally in both amplitude and phase, as it also does on the positive and negative of the output.

From the above it can be seen that both common mode and normal mode noise must be dealt with at both the input and the output if a cheap SMPS is to be silenced.

Warning

Before I go further into this, I must issue a warning to anyone tempted to modify an

SMPS along the lines I am about to describe. An SMPS is a very dangerous piece of equipment to work on. Circuits carrying ac mains voltages are involved, and in addition there is a large filter capacitor which may be packed with energy at over 300 V dc. There is enough charge contained here to cause instant death to the ignorant or careless. Do not attempt to modify such a supply unless you are fully aware of the safety procedures necessary for this work.

Method

The SMPS I bought was built into a small metal and plastic box which did not show much promise as an RF shield, so I decided to ditch it. I removed all the components from the case for future reuse. This included the main PCB of the supply, the fan, and the sundry terminals and components.

I then made up an aluminium box 260 × 200 × 80 millimetres using aluminum sheet and angle and pop riveted all but the top together. These dimensions were chosen to match the FT-897, so the transceiver could sit upon the SMPS when set up. I nominated one end of the box as the front, and drilled a series of 6 mm holes towards the front end of each of the side panels. These are to provide for ventilation when the fan is running. In the front panel is installed the mains switch, "Power On" LED and output terminals. On the other end I mounted my standard Molex type connector (to power the FT-897), the fan, an accessory outlet, the mains fuse, and an IEC type mains inlet socket with integral EMC filter. This inlet socket is of the same type as used by Drew Diamond (like Jaycar Cat No. MS4003). The main SMPS PCB was then installed into the bottom of the box, towards the front, on insulated standoffs. The original SMPS was configured for floating output — neither the positive nor negative sides of the dc output were grounded — and I decided to keep this configuration in the interests of reducing the number of ground loops in the installation.

The circuit arrangement inside the box is shown in Figure 1 and the physical arrangement is shown in the accompanying photographs. The main PCB is in two parts – mains input and dc output – and these are well isolated from each other. They remain well isolated in the additional filtering -- all external to the main PCB -- which I have provided.

The mains supply enters via the IEC filtered connector and fuse FS1. The active and neutral wires are then wound together in bifilar fashion firstly through an old pot core (one turn) to form L1, and then twice through a 30 mm toroid to form L2. The A and N wires then connect to the mains switch and PCB in the usual way. L1 and L2 provide a considerable amount of common mode noise suppression at RF, but do nothing for the normal mode component. This is dealt with by C1, a 1 μF 250 V ac mains-rated capacitor scrounged from an old PC switch mode supply and it is very effective in removing the normal mode component. F1 protects against possible failure of C1 or the main PCB, while R1 is there to discharge C1 for safety reasons. With this set up, noise levels injected into or carried by the mains are very low.

The dc output from the main PCB also has some degree of filtering on the main PCB but it is not enough. As can be seen from Figure 1, a filtering set up very similar to that suggested by Drew Diamond is used. L4 comprises a 40 mm ferrite core toroid that has seven turns of heavy duty twin core flexible speaker cable wound through it. As described by Drew, running the positive and negative leads through the toroid together prevents core saturation yet, in conjunction with C3 – C6 provides good suppression of the common mode noise components. These capacitors also suppress the normal mode noise components.

For the sake of completeness, Figure 1 also shows the circuit of a crowbar type over-voltage protection arrangement in my SMPS. I doubt the need for this – it seems that unlike linear power supplies, it is very rare for an SMPS to produce an over-voltage failure.

Some notes on the constructional aspects might be in order. First, there is only one connection point on the box and to this are connected the ground pin on the IEC connector, the center point of the C3 – C6 capacitors, and the ground point of the main PCB. This single point grounding arrangement prevents the flow of RF currents in the metalwork of the case, and as a result, it prevents radiation from the case. All capacitors larger than 1 μF are of the tantalum type; these are more effective as RF bypasses than the ordinary aluminium foil variety. The smaller capacitors are polyester type. Finally, the dc side of the main PCB must not be connected to the case; to do so would short out L4 at RF, and ruin the common mode suppression of the unit. It is okay to ground either side of the output of the SMPS, however.

Outcome

The results of the effort described above have been excellent. Under working conditions there is no detectable noise produced by the SMPS from the bottom end of the AM broadcast band through to the 10 meter amateur band. This is a very satisfying result, and I now have no concerns about operating the SMPS in any environment where RF noise might be an issue.

QEX

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Examining Multi-band Operation of Near End Fed Wire Antennas

Ron presents the results of analysis of feeding wire antennas off-center for ease of construction and direct connection of coax for multiple band operation.

The use of feed points near one end of wire antennas provides installation options that can be quite helpful where access to the mid-point of a long dipole is inconvenient. However, the unbalanced nature of such designs gives cause for concern about common-mode effects and this topic was explored in my article on Near End Fed antennas in *QEX* March/April 2009. Several readers have inquired about multi-band operation and, in response, the following details were developed using *EZNEC* models. I have not built many of these designs but intend to build more as time permits. In the meanwhile, it would be of great interest to hear about any actual experiments.

The general theory behind the Near End Fed idea is to take advantage of the fact that the impedance varies along the length as determined by the amplitude and order of the standing waves of voltage and current at any point. Numerous useful impedance matching opportunities can be found for dual-band operation. Many off-center-fed antennas have been described for single band or multi-band operation; the most famous being the early designs by Lauren Windom (*QST* Sept 1929).[1] My focus has been on feed points in the range of 5-25% from one end of the wire. Illustrations of standing waves are usually in terms of current and ignore the equally significant voltage distribution. This may be a legacy of the days when we moved RF ammeters along the wire. Now, modeling programs provide graphs to show the current distribution; voltage distribution can be inferred.

The models used sufficient segments to position a feed point with an accuracy of 1-2 ft, and show feed-point options at various impedances, frequency of lowest SWR and bandwidth. Near End Fed designs can be particularly useful as slopers and some promising models emerged with the feed point at the lower end. The type and proximity of the ground have profound affects, and, for this exercise, real ground was used as defined in *EZNEC.* Transformers can have a wide variety of ratios to match a wide range of feed-point impedance, and those with single turn primary and 4:1 and 9:1 impedance ratios are simple and effective. Direct feed with either 50 Ω or 75 Ω cable would be most convenient and this is seen to be possible in some cases.

Even with a simple dipole the variables and feed-point options are numerous. Therefore, I selected just those that looked the most promising. Table 1 lists those we will examine. It is fair to say that many of us at one time or another used the formula 468/f to calculate the length of a wire dipole, just erected it where we could, and fed it at the center with coax. We hoped for the best and used a tuner to keep the radio happy. Here are examples that show how misleading the resonant frequency, impedance and bandwidth can be relative to 50 Ω. See Table 2.

[1]L.G. Windom, W8GZ/W8ZG, "Notes on Ethereal Adornments", *QST*, September 1929

Table 1
Antenna Designs To Be Evaluated

Bands	*Height (Ft)*	*Feed Point Impedance (Ω)*
160/80	50	50/75
160/80	40	50/75
160/80	15 to 50	50/75
80/40	40	200
20/15	30	200

Table 2
Center Fed Dipole Characteristics

	160m (254 Ft, 1.84 MHz)			*80m (126 Ft, 3.7 MHz)*			*40m (66 Ft, 7.09 MHz)*		
Height	*Resonance*	*SWR*	*BW (kHz)*	*Resonance*	*SWR*	*BW (kHz)*	*Resonance*	*SWR*	*BW (kHz)*
Free Space	1.9	1.5	80	3.79	1.5	84	7.22	1.46	220
100	1.86	1.23	70	3.82	1.9	39	7.17	1.42	280
80	1.86	1.11	40	3.76	1.9	100	7.25	1.17	200
60	1.87	1.74		3.72	1.5	150	7.31	1.66	130
40	1.88	3.3		3.73	1.14	134	7.16	1.86	200
30	1.89	5.1		3.75	1.78	60	7.08	1.42	240

160 Meter Dipole 254 ft Wire at 50 and 40 ft

Clearly a low SWR will not be obtained for center-fed dipoles at some antenna heights above ground. This is generally true but is less severe at the higher frequency bands. What seems a pitfall for single band operation may turn out to be a distinct advantage for dual-band operation. Table 3 shows the possibilities of operating a 254 ft dipole on160 m and 80 m with either 50 Ω or 75 Ω feed line. All references to feed-point percent are from either end of the whole span (50% = mid-span).

The same antenna showed similar results at 40 ft with the optimum feed point at 20% as shown in Figure 1. I was surprised to find a location which could be fed with common coax for two band operation. Figure 2 illustrates one example why this occurs. It is a plot of the current distribution for both 160 m and 80 m. As expected, the fundamental has a single lobe and that for 80 m (being the second harmonic) has two. At 25% neither node

Table 3
254 Ft wire at 50 Ft

Feed point % from end	*Feed point Ft from end*	*160 m operation* 50Ω SWR	75 Ω SWR	*80 m operation* 50 Ω SWR	75 Ω SWR
40	101.6	2.1			
35	88.9	1.9	2.8	2.1	1.4
30	76.2	1.6	2.3	1.4	1.0
25	63.5	1.2	1.8	1.3	1.1
20	50.8	1.2	1.3	1.4	1.1
17.5	44.5	1.4	1.0	2.0	1.3
15	38.1	2.1	1.4		2.5

Table 4
254 Ft wire Sloper 15Ft-50 Ft

Feed point % from end	*Feed point Ft from end*	*160 m operation* 50Ω SWR	75 Ω SWR	*80 m operation* 50 Ω SWR	75 Ω SWR
30	76.2	2.6		1.2	1.7
25	63.5	2.0	2.9	1.3	2.0
20	50.8	1.4	2.1	1.2	1.8
17.5	44.5	1.0	1.6	1.05	1.6
15	38.1	1.2	1.3	1.1	1.3
12.5	31.8	1.6	1.1	1.4	1.0
10	25.4	2.6	1.7	2.1	1.4

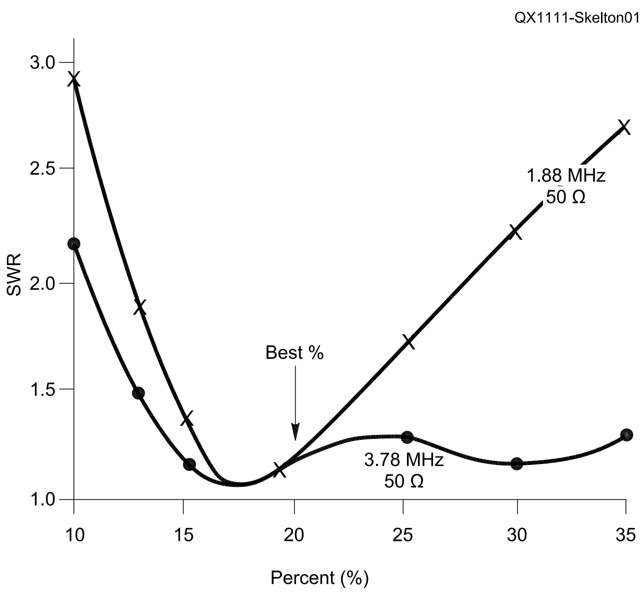

Figure 1 – 160 m and 80 m operation of a 254 ft antenna with a 50 Ω feed point showing the best position as a percentage offset. The 20% point shows the best compromise for dual band operation.

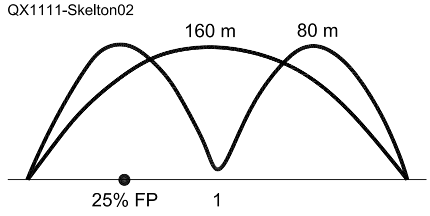

Figure 2 – Illustration of current distribution of the 254 ft antenna on both 160 m and 80 m. The 25% offset feed point shows current level for both 160 m and 80 m.

Table 5
69 Ft dipole at 30 Ft

Feed point % from end	*Feed point Ft from end*	*15 m operation 200Ω SWR*	*20 m operation 200 Ω SWR*
25	17	1.1	
20	14	1.7	1.6
17.5	12	1.9	1.3
15	10	1.8	1.1
12.5	9	1.7	1.1
11	8	1.3	1.3
10	7	1.2	1.6
8	6	1.0	
6	4	2.0	

is at a minimum (high impedance) nor at a maximum (low impedance) but somewhere between. Locations like this have been found where direct-feed with 50 Ω or 75 Ω coax will be acceptable. Moreover, the feed point does not require precise placement. Polar diagrams at these offset locations are similar to those found with center-fed dipoles. Table 4 illustrates that a typical two band sloper looks promising with a feed point about 17.5% from the low end for 50 Ω and 12.5% for 75 Ω.

80 Meter Dipole 134 ft Wire at 40 ft

The lobes for a fundamental and second harmonic are the same for other combinations so it is reasonable to suppose more examples may work well for other bands. Figure 3 confirms this is the case for the 80/40 combination but involves 4:1 transformers to convert 200 Ω to 50 Ω. As shown the optimum feed-point is 13.5% (18 ft) from one end. Any feed point which is not exactly at a point of resonance will have a reactive component in the impedance which increases the further it is from resonance. This may affect the bandwidth, but most designs showed only slightly less than when fed at an exact point of resonance.

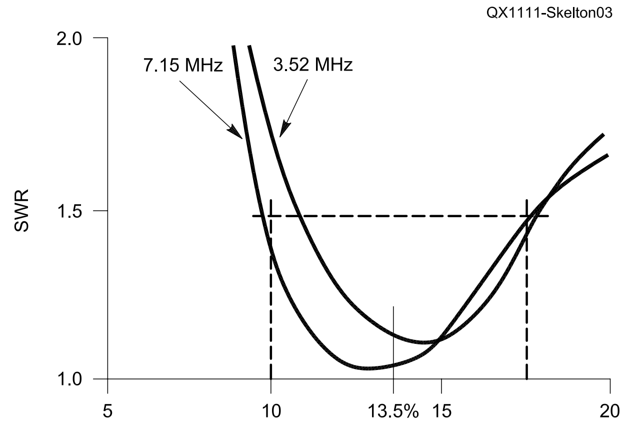

Figure 3 — Dual band operation of a 134 ft dipole at 50 Ω using a 4:1 transformer for a 200 Ω impedance. The optimum dual band position for this antenna is 13.5%.

40 Meter Dipole at 30 ft

It is often claimed a center-fed 40 meter dipole can be operated on its third harmonic at 15 meters because each band has a current node at the center. This is a reasonable proposition and it was rather surprising not to find an acceptable match to either 50 Ω or 75 Ω coax. Not only was the SWR on 15 m excessive but the point of resonance had shifted above the top end of the band. It was also expected the 40 m dipole would support the second harmonic 20 m operation much as was found on the lower bands. No such operation was found, however, as once more the resonance on 20 m was well above the top end of the band. The 66 ft wire (so well suited for 40 m single band operation) needs to be increased to 69 ft if dual band operation is to be achieved. This length moved 40 m below the band so hopes of a three band design faded. My attention shifted to finding a design that might just work on 20 m and 15 m. There are several possible feed points and in general the lower impedances are to be preferred as being less critical than those for higher ratios; 200 Ω at 11% worked well.

Future Directions

For further work on this subject, I am interested in a better understanding of how the many variables interact both on paper and by actual construction. Why, for example, did the frequency of resonance change so much in the 40/20/15 m case? Is this related to the end-effect, diameter of the wire, etc? One observation made from those models is that the node impedance is related to the order of the harmonic; the higher the order of the harmonic the higher the impedance at a node. This also might provide other multi-band options. The idea of incorporating some reactive loading seems worth following up.

Conclusions

There is always much to be learned by building models or real antennas. Sometimes something new shows up, as is the case here, at least on paper. It is quite clear the usual midpoint feed with direct coax cannot be taken for granted especially on the LF bands. It appears possible to depart from the center, directly feed with coax and in some cases to also operate on two bands. It seems conventional wisdom is not to be trusted for operation with third harmonics as illustrated in the 40/15 m case. My choice for dual-band operation would be 200 Ω, but the feed point location is not a given and shifts closer to the end as the height is reduced. Reasonable sloper arrangements came out well and may have desirable directional properties. If your shack is midway in a clearing between a pair of 150 ft redwoods, 500 ft apart you probably will not benefit much from Near End Fed designs. For the rest of us, it is a fertile area for experimentation. Just don't forget those ferrite line chokes.

Ron was first licensed in the UK as G3IHP in 1950/51. He worked for many years in Commonwealth countries in North America, Africa, Asia and Caribbean. He moved to the US in 1975 and retired following a varied career in communications and information systems. He is a Fellow of the Institution of Engineering and Technology in the UK and is a Life Member of the IEEE.

QEX

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The Design of Mixed-Coupling LC Band-Pass Ladder Filters

The author examines the frequency relationships in wideband asymmetrical mesh-coupled band-pass ladder filters, briefly discusses the design difficulties associated with them, and describes a simple procedure for designing symmetrical band-pass ladder filters using mixed coupling.

Introduction

The mesh-coupled form of LC ladder filter shown in Figure 1 has become increasingly popular for low-impedance HF use over the last three decades. It's but one of a family of filters, which includes the commonly used pair of top-coupled parallel-tuned circuits. The latter is from the node-coupled side of the family, of course, but despite individual differences in circuit topology, the members of this family have common characteristics as well as some common elements in their design procedure. The solid curves (A) and (C) in Figure 2 show the types of asymmetrical response produced by these mesh-coupled ladder filters. The capacitor-coupled version produces a mirror image (A) of the response produced by the inductor-coupled version (C). The dashed curve (B) is the response of a conventional logarithmically symmetrical band-pass filter, which is shown for comparison. The node-coupled version using capacitors to connect between shunt parallel-tuned circuits produces the same shape of asymmetrical response (C) as mesh coupling with inductors, and the mirror image response (A) is produced by mesh coupling with capacitors and node coupling with inductors. One of the main advantages of this family of LC ladder filters is that they have, or can be made to have reasonably convenient component values. The characteristic asymmetry of the frequency response can often be useful to provide more attenuation on the side where mixing images occur. Sometimes more sym-

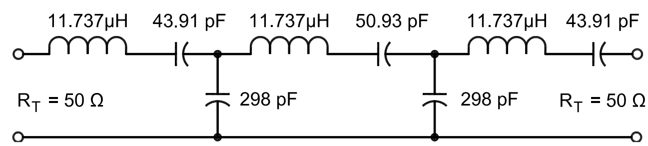

Figure 1 — Standard third-order 1 dB-ripple Chebyshev LC band-pass ladder filter with mesh coupling provided by capacitors. The center frequency is approximately 7.45 MHz.

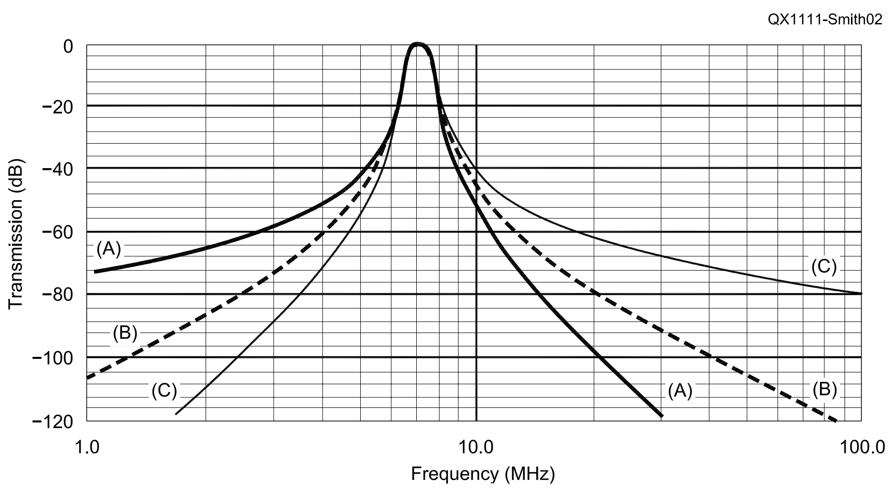

Figure 2 — Comparative frequency responses of different types of third-order LC band-pass ladder filters centered on 7.15 MHz. A) Mesh coupling by capacitors or node coupling by inductors. B) Conventional band-pass or mixed coupling. C) Mesh coupling by inductors or node coupling by capacitors.

metrical responses are required, however, when the reduction of mixing images has to be balanced against reducing the effect of strong broadcast signals on the other side of the response. Mesh-coupled LC ladder filters with coupling provided by a mixture of capacitors and inductors produce symmetrical responses just like those of conventional LC band-pass filters, but usually with more convenient component values. The transformation from low-pass to band-pass form in both node- and mesh-coupled ladder circuits, where the same type of coupling is used throughout, causes a frequency shift that is often ignored in designs where the relative bandwidth is small. However, the issue of transformation frequency shift has to be addressed in wideband ladder designs in order to achieve controlled responses of the Chebyshev or Butterworth type. Getting the mesh frequencies right can be particularly important when a high degree of matching is required and a minimum return loss specification has to be met.

Table 1

Normalized Design Coefficients for Third-Order Mixed-Coupling LC Band-Pass Ladder Filters

Ripple	*q*	k_{12}	k_{23}	*b*	$q \times k_{12}$
1.00E–02	1.1811	0.6818	0.6818	0.2324	0.8053
9.00E–03	1.1743	0.6825	0.6825	0.2329	0.8015
8.00E–03	1.1671	0.6834	0.6834	0.2335	0.7976
7.00E–03	1.1592	0.6842	0.6842	0.2341	0.7931
6.00E–03	1.1507	0.6852	0.6852	0.2347	0.7885
5.00E–03	1.1412	0.6864	0.6864	0.2356	0.7833
4.00E–03	1.1304	0.6877	0.6877	0.2365	0.7774
3.00E–03	1.1178	0.6893	0.6893	0.2376	0.7705
2.00E–03	1.1022	0.6913	0.6913	0.2389	0.7620
1.40E–03	1.0902	0.6930	0.6930	0.2401	0.7555
1.00E–03	1.0803	0.6944	0.6944	0.2411	0.7502
7.00E–04	1.0710	0.6957	0.6957	0.2420	0.7451
5.00E–04	1.0632	0.6968	0.6968	0.2428	0.7408
3.00E–04	1.0531	0.6984	0.6984	0.2439	0.7355
2.00E–04	1.0462	0.6994	0.6994	0.2446	0.7317
1.00E–04	1.0365	0.7010	0.7010	0.2457	0.7266
5.00E–05	1.0289	0.7022	0.7022	0.2465	0.7225
1.00E–05	1.0168	0.7042	0.7042	0.2479	0.7160
1.00E–06	1.0078	0.7058	0.7058	0.2491	0.7113
0.00E+00	1.0000	0.7071	0.7071	0.2500	0.7071

Note: Data for third-order mixed-coupling LC band-pass ladder filter design. The $q \times k_{12}$ column allows any design to be tailored to fit in with preferred values of coupling capacitor if some variation in performance can be tolerated — see text for details.

Frequency Relationships, Coupling and Wideband Design

The capacitor-coupled LC ladder filter shown in Figure 1 is a 1 dB ripple Chebyshev design with a working *Q* (Q_w) of approximately 5, centered on f_0 = 7.449 MHz — this is the geometric mean of its –3 dB frequencies. Its pass-band response is presented in Figure 3. Note that the middle peak of the Chebyshev response (f_{P2}) does not coincide with the center frequency, f_0, but is located at 7.500 MHz. Furthermore, the mesh frequency of the middle section (marked as f_{M2}) has to be set at 7.5405 MHz and the end sections at 7.500 MHz to achieve a reasonably precise 1 dB ripple Chebyshev response. So, both the geometric mean (center) frequency and the middle peak frequency are lower than the frequency of the middle mesh. In a conventional band-pass filter made up of series and parallel resonant circuits these three frequencies would be coincident. In the case of band-pass ladder filters coupled by inductors, the sense of the frequency shift is the other way round, and both the center frequency and the middle peak of the response are above the frequency of the middle mesh. This difference between the pass-band center frequency and the middle mesh frequency is a function of the bandwidth, the center frequency and the coupling factors. It's very small in narrow ladder filters, and, consequently, is not normally included in the design procedure. The coupling factors required for a particular pass-band ripple also decrease as Q_w is reduced, though the *q* values remain pretty close to their narrow-band values.

In wideband LC ladder filters with only one type of coupling, the mesh frequencies required for perfectly controlled responses are quite complicated to predict. Both the sense and the magnitude of these frequency differences between adjacent meshes change with pass-band ripple and filter order, so the prudent approach is very often just to do a rough design and then rely on getting the mesh frequencies right when aligning the finished filter.

When a capacitor-coupled ladder filter is combined with an inductor-coupled one to form a third-order, mixed-coupling band-pass ladder filter, as shown in Figure 4, the situation is much simpler to analyze and easier to predict. The middle mesh needs to resonate at a frequency that is the geometric mean of the two end-section frequencies, and this coincides with the filter center frequency. Since the frequency differences between the end and middle meshes are small compared with the bandwidth, they can be made equal to simplify the design procedure. The frequency offsets of the end meshes above and below the frequency of the center mesh can be approximated by a simple expression that is accurate enough (within 1%) for all practical purposes down to Q_w values of around 3.5. The end section, which couples to the middle section through a capacitor, needs to resonate higher in frequency than the middle section because of its transformation shift, and the inductor-coupled end section has to resonate lower in frequency in order to achieve an equal-ripple response. In a properly designed third-order band-pass ladder filter with mixed coupling, the middle peak frequency, the middle mesh frequency and the geometric mean of the –3 dB frequencies should all be coincident, and the response then follows the curve for a conventional band-pass filter (B) shown in Figure 2. The amount by which the end-section frequencies need to be offset with respect to the frequency of the inner mesh depends upon the pass-band ripple, the –3 dB bandwidth (BW_3), and the center frequency (f_0). Approximations for these offsets ($\pm\Delta f$) are given by:

$$\Delta f = \frac{\pm\, b \times BW_3^2}{f_0} \qquad \text{[Eq 1]}$$

The constant *b* can be calculated from the –3 dB coupling coefficients for any particular pass-band ripple and order. Coupled tuned circuits that have a frequency offset do not interact to the same degree as those that are resonant on the same frequency. Therefore, wide responses produced using the standard narrow-band *k* and *q* values are not exact for both this reason and the fact that the relative bandwidths are large. They are sufficiently close, however, to be very useful in practice, and always achieve return loss figures better than the target value for low-ripple designs. Typically, the standard *k* and *q* figures for a third-order 0.01 dB Chebyshev filter will produce 0.009 dB ripple with a return loss of over 27 dB, and the figures for a 0.001 dB Chebyshev response will yield a filter with 0.0007 dB ripple and 38 dB return loss for a design of Q_w = 4.

Third-Order LC Filters with Mixed Coupling

Designing a wide, geometrically symmetrical, third-order ladder filter is only slightly more involved than designing one with just capacitor coupling. The first step should always be to establish the total inductance (L_m) required for each mesh. This can be calculated from the –3 dB q value given for the desired level of pass-band ripple in Table 1, using Equation 2.

$$L_m = \frac{q \times Q_w \times R_T}{2\pi f_0} \qquad \text{[Eq 2]}$$

In Equation 2, Q_W and f_0 are as previously defined, and R_T is the termination resistance. Once L_m has been determined, the coupling inductance L_{23} can then be calculated from Equation 3.

$$L_{23} = \frac{k_{23} \times L_m}{Q_w} \qquad \text{[Eq 3]}$$

Here, k_{23} is the –3 dB k value in the k_{23} column in Table 1. The two coupling components C_{12} and L_{23} should have the same reactance at the filter center frequency and, therefore, must resonate at f_0, so

$$C_{12} = \frac{1}{4\pi^2 f_0^2 L_{23}} \qquad \text{[Eq 4]}$$

Since C_{12} and L_{23} resonate at f_0, L_{S2} and C_{S2} should also resonate at f_0 separately, since the whole central section should resonate at f_0 when all these components are considered together as an isolated mesh. C_{S1} in series with C_{12} must resonate with L_m at $f_{E1} = f_0 + \Delta f$, so the value given in the b column of Table 1 for the required ripple should be used in Equation 1 to establish Δf and then f_{E1}. The simplest way to calculate the value of C_{S1} is to determine the capacitance C_{T1} required to resonate with L_m at f_{E1}, and then subtract the reciprocal of C_{12} from the reciprocal of C_{T1}, and take the reciprocal of that result — this is the value of C_{S1}. The mesh at the end containing C_{S3} needs to resonate with L_m at $f_{E3} = f_0 - \Delta f$. Once this has been calculated, the design is almost complete. The final adjustment to the design is best done by checking the return loss using a circuit analysis program such as the one offered by AADE.[1] An equal-ripple response will have return loss peaks that are equal in level within the passband. If the imbalance in the return loss peaks of the design as it stands is not acceptable, only the values of C_{S1} and C_{S3} should need slight adjustment to achieve the equal-ripple condition. Very often, this can be achieved solely by altering the value of C_{S3} slightly. The resulting design should be within +1% of the required bandwidth providing Q_w is not substantially less than 4. Even at $Q_w = 2$, though, the error in the –3 dB bandwidth should not be much more than +3%, and in practice this will be masked to some degree by the effect of coil loss on the bandwidth, anyway.

[1]Notes appear on page 32.

Third-Order Mixed-Coupling Design Example

Let's say we need a design that passes the entire band from 3.5 to 4.0 MHz with a reasonable match to the load and source. A 0.01 dB design with a mismatch of up to 10% within the ripple passband would probably meet this specification. The ratio of the ripple bandwidth to the –3 dB bandwidth for this amount of pass-band ripple is about 0.533, so a Q_w of 3.6 would provide adequate bandwidth with some margin for error. The geometric mean of the band edge frequencies is 3.742 MHz, and this will be used as the center frequency of the filter. The q value from Table 1 for 0.01dB ripple is 1.1811, and so for 50 Ω terminations, L_m will need to be:

$$L_m = \frac{1.1811 \times 3.6 \times 50}{2 \times 3.142 \times 3.742 \times 10^6}$$

$L_m = 9.04223$ µH

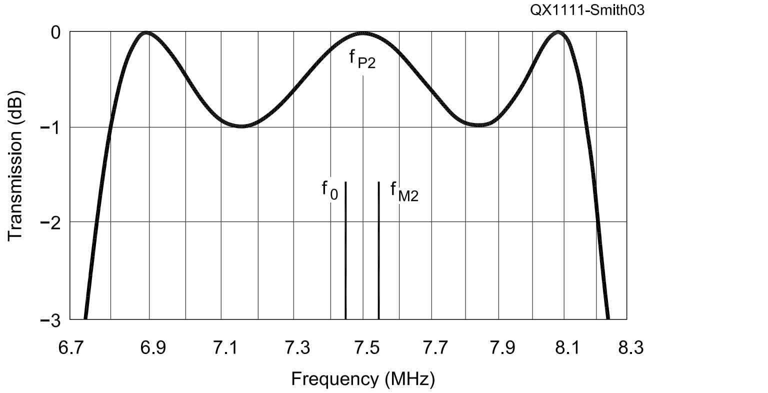

Figure 3 — The pass-band response of the third-order Chebyshev LC band-pass ladder filter shown in Figure 1. The frequency of mesh 2 is f_{M2} and f_0 is the geometric mean of the –3 dB frequencies (Q_w = 5).

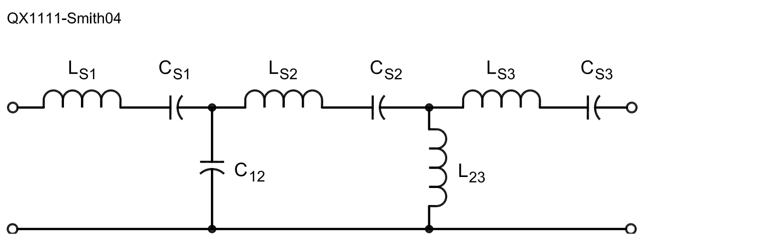

Figure 4 — Circuit diagram of a third-order LC band-pass ladder filter with mesh coupling provided by both a capacitor and an inductor — mixed coupling.

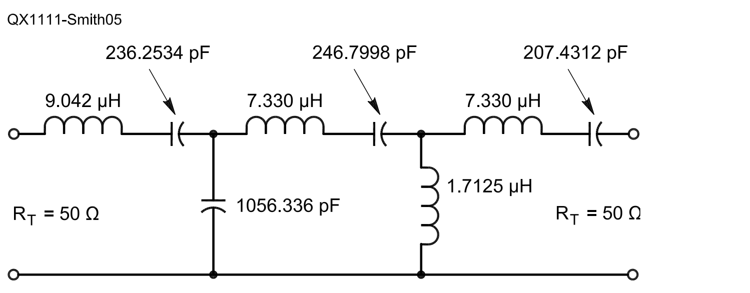

Figure 5 — Circuit diagram of the 0.01 dB-ripple band-pass ladder filter used as the third-order mixed-coupling design example.

Using the k_{23} figure from the 0.01 dB row in Table 1:

$$L_{23} = \frac{0.6818 \times 9.04223 \times 10^6}{3.6}$$

L_{23} = 1.7125 μH

C_{12} must resonate at 3.742 MHz with L_{23}, so it needs to be 1056.336 pF. The value of b from Table 1 is 0.2324, and the –3 dB bandwidth is $(3.742 \times 10^6) / 3.6$ = 1.039444 MHz. So, Δf from Equation 1 is 67.102 kHz. This makes f_{E1} = 3.809102 MHz and f_{E3} approximately 3.674898 MHz. Therefore, C_{T1} is 193.0721 pF, and taking $1/C_{12}$ away from $1/C_{T1}$ gives $1/C_{S1}$ as 1/236.2534 pF, or C_{S1} = 236.2534 pF. In mesh 2, $L_{S2} = L_m - L_{23}$ = 7.32973 μH and since C_{S2} must resonate at 3.742 MHz with this inductance, it needs to be 246.7998 pF. L_{S3} always equals L_{S2} because their meshes both share L_{23}, and, therefore, it is also 7.32973 μH. Since C_{S3} has to resonate with L_m at f_{E3}, it is 207.4312 pF. The full circuit of this design is shown in Figure 5. In practice, the values of these inductors and capacitors can be made adjustable, so the precision shown for the component values in these calculations is only there so that they can be used to check out the basic design by computer modeling.

The pass-band response and return loss plot for this filter using components without loss are shown in Figure 6, parts A and B. The filter, as designed, has a –3 dB bandwidth of 1.046341 MHz with a center frequency at 3.742237 MHz. Therefore, the error in the bandwidth is +0.66%, which is reasonable considering that the value of Q_w is below 4. The slight imbalance of the two peaks in the return loss plot indicates that the response is not quite equal ripple, but very close — the lower peak rises to 27 dB down and the upper one to 27.5 dB down. This slight difference could be corrected by adjusting the value of C_{S3}, but it is so small that it's not worth worrying about. Both of these plots show that the whole amateur band fits nicely into the region where the match to source and load is quite good.

Fourth-Order LC Filters with Mixed Coupling

Even-order ladder filters have an odd number of coupling components, and consequently can't produce a logarithmically symmetrical response with mixed coupling. Obviously, a fourth-order filter must have more of one type of coupling component than the other, and, consequently, the response will not be as symmetrical as a conventional LC band-pass filter of the same order, but it will be better than a standard ladder filter using only one type of coupling element. The two possible circuit arrangements

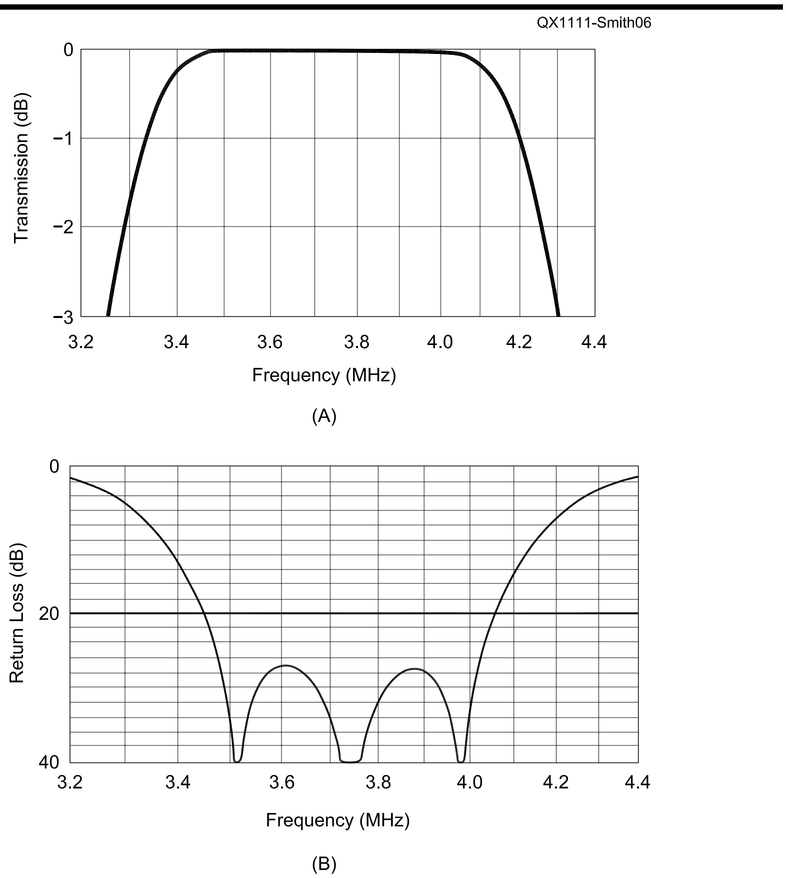

Figure 6 — Plots for the 0.01 dB-ripple third-order mixed-coupling design example. A) Pass-band amplitude response between the –3 dB points and B) the return loss down to 40 dB.

Table 2
Normalized Design Coefficients for Fourth-Order Mixed-Coupling LC Band-Pass Ladder Filters

Ripple	*q*	k_{12}	k_{23}	k_{34}	*b*	$q \times k_{12}$
1.00E–02	1.0457	0.7369	0.5413	0.7369	0.2939	0.7706
9.00E–03	1.0371	0.7391	0.5413	0.7391	0.2957	0.7665
8.00E–03	1.0279	0.7414	0.5413	0.7414	0.2975	0.7621
7.00E–03	1.0179	0.7440	0.5413	0.7440	0.2996	0.7573
6.00E–03	1.0069	0.7470	0.5413	0.7470	0.3021	0.7522
5.00E–03	0.9945	0.7504	0.5413	0.7504	0.3048	0.7463
4.00E–03	0.9803	0.7544	0.5412	0.7544	0.3080	0.7395
3.00E–03	0.9635	0.7594	0.5412	0.7594	0.3121	0.7317
2.00E–03	0.9422	0.7661	0.5412	0.7661	0.3176	0.7218
1.40E–03	0.9256	0.7716	0.5412	0.7716	0.3222	0.7142
1.00E–03	0.9114	0.7765	0.5412	0.7765	0.3263	0.7077
7.00E–04	0.8979	0.7813	0.5412	0.7813	0.3304	0.7015
5.00E–04	0.8863	0.7856	0.5412	0.7856	0.3340	0.6963
3.00E–04	0.8706	0.7916	0.5412	0.7916	0.3391	0.6892
2.00E–04	0.8600	0.7960	0.5412	0.7960	0.3429	0.6846
1.00E–04	0.8442	0.8026	0.5412	0.8026	0.3486	0.6776
5.00E–05	0.8311	0.8083	0.5412	0.8083	0.3536	0.6718
3.00E–05	0.8229	0.8120	0.5412	0.8120	0.3568	0.6682
1.00E–05	0.8087	0.8187	0.5412	0.8187	0.3627	0.6621
4.00E–06	0.7996	0.8231	0.5412	0.8231	0.3667	0.6582
1.00E–06	0.7895	0.8262	0.5412	0.8262	0.3694	0.6523
1.00E–07	0.7788	0.8337	0.5412	0.8337	0.3762	0.6493
0.00E+00	0.7654	0.8409	0.5412	0.8409	0.3827	0.6436

Note: Data for fourth-order mixed-coupling LC band-pass ladder filter design. The $q \times k_{12}$ column allows any design to be tailored to fit in with preferred values of coupling capacitor if some variation in performance can be tolerated — see text for details.

for fourth-order, mixed-coupling, LC band-pass ladder filters are shown in Figure 7, parts A and B. The value of L_m can be calculated using Equation 2. The two inner meshes should be resonated at the center frequency, just as the middle mesh is in third-order filters, but the frequencies of both the end sections need to be offset in the same direction. The frequency offset (Δf) can be worked out using Equation 1 and the value of *b* given for fourth-order filters in Table 2. For the circuit given in Figure 7, Part A, the offset is positive since the coupling to the end sections is through a capacitor. The offset for the circuit in Figure 7, Part B, of course is negative. The inductor and capacitor coupling components in a mixed fourth-order filter are not in equivalent positions, and consequently don't have the same reactance at the center frequency. So, although L_{23} can be worked out from Equation 3 for a fourth-order CLC design, there must be a correction for working out the value of the coupling capacitors, and

$$C_{12} = C_{34} = \frac{k_{23}}{4\pi^2 f_0^2 k_{12} L_{23}} \qquad \text{[Eq 5]}$$

This just calculates C_{12} and C_{34} as if they have the same reactance as L_{23} at the center frequency, and corrects for the difference through k_{23} / k_{12}. Once L_m, L_{23}, C_{12} and C_{34} have been worked out, the values of the remaining components can be calculated by isolating each mesh in turn and working out the value of C_T for resonance at the required mesh frequency while keeping the total mesh inductance constant at L_m. Each C_S can then be determined from its mesh C_T using the reciprocal rule for series capacitors.

The procedure for the design of the LCL version requires L_{12} and L_{34} to be worked out from Equation 6.

$$L_{12} = L_{34} = \frac{k_{12} \times L_m}{Q_w} \qquad \text{[Eq 6]}$$

This can then be used to derive the value of C_{23} from Equation 7.

$$C_{23} = \frac{k_{12}}{4\pi^2 f_0^2 k_{23} L_{12}} \qquad \text{[Eq 7]}$$

Taking a 40-meter CLC fourth-order band-pass filter with a Q_w of 10 and ripple of 0.001 dB as an example, the value of mesh inductance is:

$L_m = 0.9114 \times 10 \times 50 / 2 \times 3.142 \times 7.148 \times 10^6 = 10.14646$ µH.

Using Equation 3 for the circuit in Figure 6, part A,

$L_{23} = 0.5412 \times 10.14646 / 10 = 0.54913$ µH.

Then working out the coupling capacitors with Equation 5 gives:

$C_{12} = C_{34} = 0.5412 / (4\pi^2 \times 7.148^2 \times 10^{12} \times 0.7765 \times 0.54913 \times 10^6) = 629.2342$ pF.

Meshes 2 and 3 must resonate at 7.148MHz, so C_{12} in series with C_{S2} must be C_{T2} = 48.8604 pF. Since C_{12} is 629.2342 pF, C_{S2} must be 52.9738 pF. C_{S3} is the same value. The –3 dB bandwidth is 7.148 MHz / 10 = 714.8 kHz. Therefore, the frequency offset using Equation 1 is given by:

$\Delta f = + 0.3263 \times (714.8 \times 10^3)^2 / (7.148 \times 10^6) = 23.327$ kHz.

Both end sections must resonate at $f_0 + \Delta f$ = 7.171327 MHz. Therefore, C_{T1} and C_{T4} should be 48.543 pF. Taking $1/C_{12}$ away from $1/C_{T1}$ gives $1/C_{S1}$, which when inverted results in C_{S1} = 52.611 pF. C_{S4} should be the same value. L_{S1} and L_{S4} are the only inductors in meshes 1 and 4, so should both be 10.14646 µH. The inductor L_{23} is shared by meshes 2 and 3, so L_{S2} and L_{S3} should be equal to (L_m – L_{23}), or 9.59733 µH. The circuit of the completed design is presented in Figure 8. Analysis using AADE 4.42 indicates that the center frequency is 7.147408 MHz and the bandwidth 715.154 kHz (+0.05%). The amplitude response between the –3 dB points is shown in Figure 9, Part A. The ripple is not quite equal across the passband, however, as indicated by the unequal peaks in the return loss plot of Figure 9, Part B. At a minimum return loss of 35 dB down, though, the match across the entire passband is unlikely to be a problem.

Fifth-Order LC Filters with Mixed Coupling

There is an even number of coupling components in a fifth-order ladder filter, as shown in Figure 10, and an equal number of capacitors and inductors can be used to produce a logarithmically symmetrical response. Table 3 provides *k*, *q* and *b* values for fifth-order Butterworth and Chebyshev filters up to a ripple of 0.01 dB. Again, as with the third-order design procedure, use can be made of the symmetry of the circuit. Therefore, L_{12} needs to resonate at the center frequency with C_{45}, and L_{34} with $\mathbf{C_{23}}$. The starting point, as always, is to calculate L_m using Equation 2, and then L_{12} can be worked out from Equation 6 and C_{45} from Equation 8.

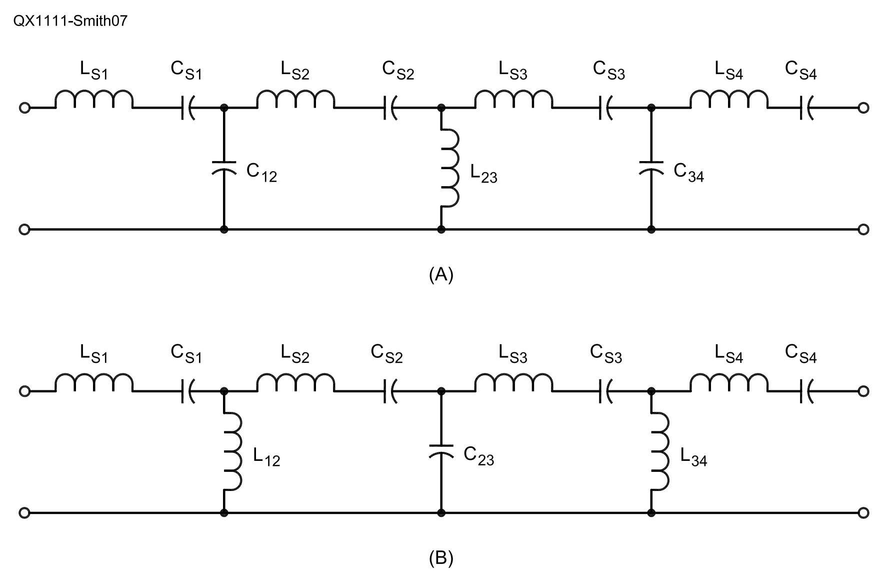

Figure 7 — The two versions of a fourth-order mixed-coupling LC band-pass ladder filter; A) the CLC type and B) the LCL version.

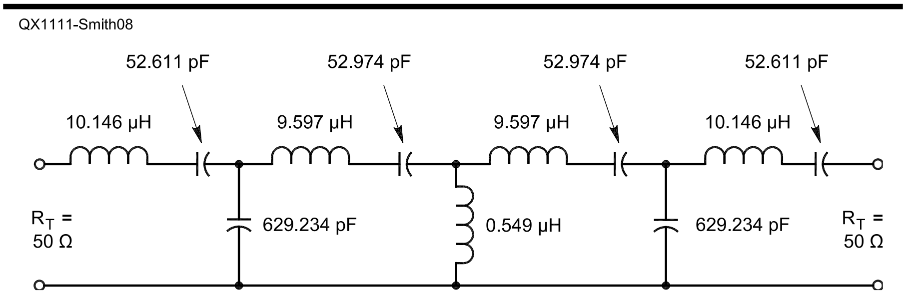

Figure 8 — Component values for the fourth-order CLC 0.001 dB-ripple mixed-coupling band-pass ladder filter design example discussed in the text.

Table 3
Normalized Design Coefficients for Fifth-Order Mixed-Coupling LC Band-Pass Ladder Filters

Ripple	*q*	k_{12}	k_{23}	k_{34}	k_{45}	$q \times k_{12}$	*b*
1.00E–02	0.9766	0.7796	0.5398	0.5398	0.7796	0.7614	0.3034
9.00E–03	0.9670	0.7829	0.5400	0.5400	0.7829	0.7571	0.3065
8.00E–03	0.9567	0.7866	0.5403	0.5403	0.7866	0.7525	0.3094
7.00E–03	0.9454	0.7907	0.5405	0.5405	0.7907	0.7475	0.3126
6.00E–03	0.9329	0.7955	0.5408	0.5408	0.7955	0.7421	0.3164
5.00E–03	0.9190	0.8010	0.5412	0.5412	0.8010	0.7361	0.3208
4.00E–03	0.9028	0.8076	0.5416	0.5416	0.8076	0.7291	0.3261
3.00E–03	0.8834	0.8160	0.5422	0.5422	0.8160	0.7209	0.3329
2.00E–03	0.8588	0.8274	0.5430	0.5430	0.8274	0.7106	0.3423
1.40E–03	0.8392	0.8369	0.5436	0.5436	0.8369	0.7023	0.3502
1.00E–03	0.8224	0.8456	0.5442	0.5442	0.8456	0.6954	0.3575
7.00E–04	0.8062	0.8545	0.5448	0.5448	0.8545	0.6889	0.3651
5.00E–04	0.7923	0.8624	0.5454	0.5454	0.8624	0.6833	0.3719
3.00E–04	0.7732	0.8739	0.5462	0.5462	0.8739	0.6757	0.3819
2.00E–04	0.7597	0.8824	0.5468	0.5468	0.8824	0.6704	0.3893
1.00E–04	0.7396	0.8958	0.5478	0.5478	0.8958	0.6625	0.4012
3.00E–05	0.7116	0.9161	0.5493	0.5493	0.9161	0.6519	0.4196
1.00E–05	0.6921	0.9314	0.5505	0.5505	0.9314	0.6446	0.4338
4.00E–06	0.6790	0.9422	0.5513	0.5513	0.9422	0.6398	0.4439
1.00E–06	0.6637	0.9555	0.5524	0.5524	0.9555	0.6342	0.4565
1.00E–07	0.6465	0.9714	0.5536	0.5536	0.9714	0.6280	0.4718
1.00E–08	0.6358	0.9818	0.5544	0.5544	0.9818	0.6242	0.4820
0.00E+00	0.6180	1.0000	0.5559	0.5559	1.0000	0.6180	0.5000

Note: Data for fifth-order mixed-coupling LC band-pass ladder filter design. The $q \times k_{12}$ column allows any design to be tailored to fit in with preferred values of coupling capacitor if some variation in performance can be tolerated — see text for details.

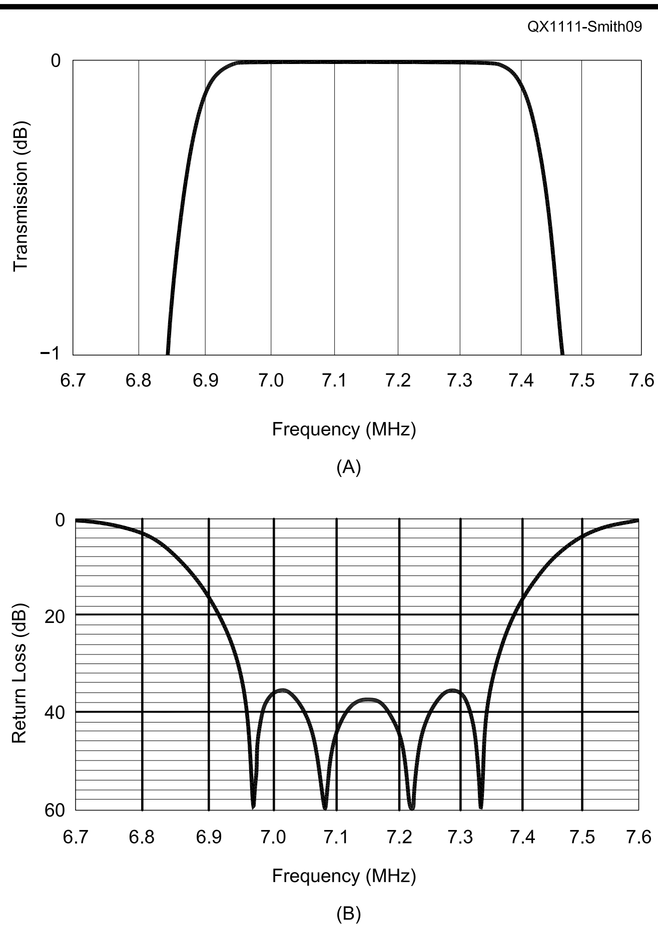

Figure 9 — Plots of A) the amplitude response between the –1 dB points and B) the return loss down to 60 dB for the 0.001 dB-ripple fourth-order CLC band-pass filter derived in the design example.

$$C_{45} = \frac{1}{4\pi^2 f_0^2 L_{12}} \quad \text{[Eq 8]}$$

Similarly,

$$L_{34} = \frac{k_{34} \times L_m}{Q_w}$$

and

$$C_{23} = \frac{1}{4\pi^2 f_0^2 L_{12}}.$$

Initially, the three inner meshes should be resonated at the center frequency, so C_{T2}, C_{T3} and C_{T4} are all equal and need to resonate at f_0 with L_m. The values of C_{S2}, C_{S3} and C_{S4} can then be worked out from the calculated value of C_{T2} using the reciprocal rule for series capacitors and their respective mesh coupling capacitors, C_{23} or C_{45}. In the case of very wide filters, the middle mesh may need to be offset very slightly to obtain a more precise fit to the desired response, but that can be finalized, if needed, at the circuit analysis stage when the design is checked for response shape and bandwidth. The inductance of each isolated mesh must equal L_m, so $L_{S1} = (L_m - L_{12})$, $L_{S2} = (L_m - L_{12})$, $L_{S3} = (L_m - L_{34})$, $L_{S4} = (L_m - L_{34})$ and $L_{S5} = L_m$. The resonant frequency of the end meshes can be calculated using Equation 1 and the value of *b* given for the required pass-band ripple in Table 3. Mesh 5, being coupled to the rest of the filter by a capacitor, must be set at $f_0 + \Delta f$, and mesh 1 being inductively coupled at $f_0 - \Delta f$.

Figure 11 presents a fifth-order Butterworth band-pass filter design with mixed coupling that has been produced using this procedure. The filter has a Q_w of 8 and center frequency of 7.150 MHz. The pass-band response and return loss plot for this filter are shown in Figure 12, parts A and B. Although not perfect, the return loss is better than 55 dB down over the main part of the passband. This corresponds to an SWR of better than 1.004:1.

Summary and Final Remarks

Standard narrow-band *k* and *q* figures can be used to design quite closely controlled LC ladder filters with wide, symmetrical band-pass responses using mixed coupling, provided the working Q is not substantially less than 4. Even wider mixed-coupling band-pass ladder filters with Q_w in the range 2 to 4 can produce quite good pass-band responses with acceptable levels of error using the standard k and q values as a starting point. The deviation of the pass-band response from the ideal shape increases as the working Q and the pass-band ripple are reduced, and the order of the filter increased. Despite these limitations, however, there are many applications where LC ladder filters with mixed coupling offer a more convenient solution than the conventional symmetrical band-pass filter circuit.

The procedure for designing wideband ladder filters is relatively simple, but differs slightly to that used for narrow-band ladder designs where approximations are used to further simplify the process. The main difference between the two is that the coupling capacitor values in wideband designs are based on the value of the mesh tuning capacitance, C_T, and not on C_S as they are in designs with very narrow bandwidths, where the two values are pretty much the same because C_S is so very small compared to the coupling capacitor values. The wideband design procedure is the more general case and could be applied to narrow-band design, whereas the converse is not true.

The input and output impedances of mixed-coupling band-pass ladder filters can be transformed by extracting part of C_{S1} and C_{Sn} as series capacitors, and doing a series-parallel conversion with the natural source and load resistance, just as they can with other forms of mesh-coupled ladder filter. This technique allows the input and output impedance to be increased above the natural level, or conversely, the L/C ratio reduced for a particular impedance and center frequency. The latter can be useful for adjusting the inductor values so they to conform to standard values that are readily available, but requires at least two additional capacitors to implement the change.

The tables of filter design data presented in this article were put together using information from a number of projects that span a considerable number of years. The tables only stop at 0.01 dB ripple because they were all derived for applications where a reasonably good match to source and load were required, and they cover a wide range of ripple values because that sometimes allowed the selection of a design that coincided with a convenient standard value of coupling capacitor (SVC). The $q \times k_{12}$ column in all three tables was used for working out the ripple value that might give a preferred value from an initial starting ripple that didn't. The ratio of the nearest standard value to the calculated value of coupling capacitor for the initial starting ripple was multiplied by the product in the $q \times k_{12}$ column for that ripple to give a new product that could be looked up in the $q \times k_{12}$ column to identify which ripple level might give a convenient coupling value.

Of course, mixed coupling can be used for higher ripple band-pass filters as well, if attenuation is more important that good matching. Standard narrow-band k and q figures for higher levels of ripple have been published by the author in a previous article (1dB) and also by Zverev in his filter handbook (0.1 and 0.5 dB).[2, 3]

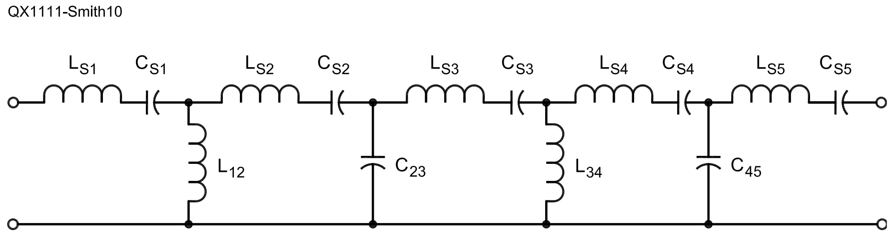

Figure 10 — Circuit diagram of a fifth-order mixed-coupling LC band-pass ladder filter showing component designations.

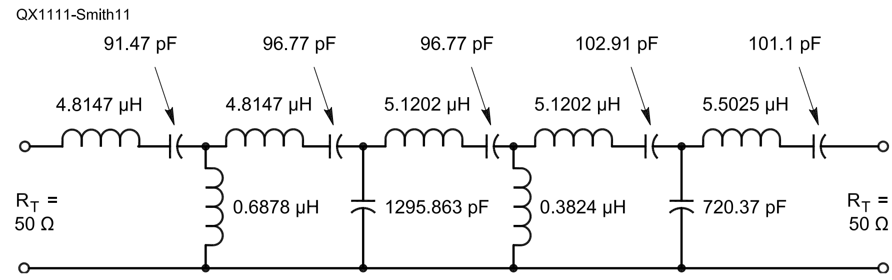

Figure 11 — Component values for the Butterworth LC band-pass ladder filter used as the fifth-order mixed-coupling design example.

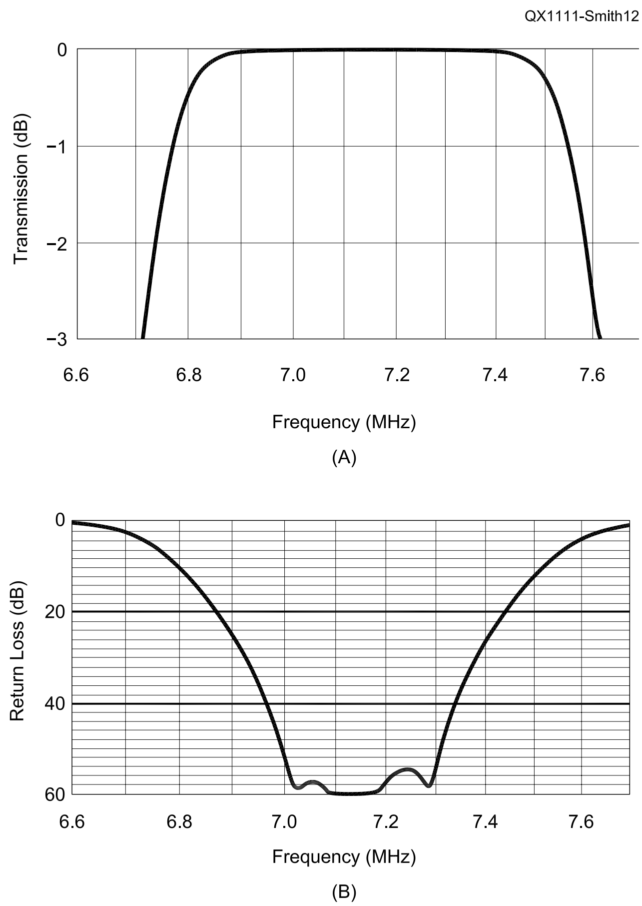

Figure 12 — Plots for the Butterworth fifth-order mixed-coupling design example; A) pass-band response between the –3 dB points and B) the return loss down to 60 dB.

Dave Gordon-Smith; G3UUR, was first licensed in 1965 and concentrated mainly on home construction and 160 m CW DX during his first few years on the air. Constructing his own equipment was a necessity in those days because he was an impoverished schoolboy, but it later became a source of great fun and satisfaction. Filters, antennas, and propagation have always fascinated him, and much of his Amateur Radio construction and experimentation has been driven by a desire to understand them. He tends to use theory to make up for a lack of test gear and enjoys playing with novel and unusual circuits. The RSGB awarded him the Courtenay Price Trophy for his work on variable bandwidth crystal ladder filters in 1981.

He served with VSO (British Peace Corps) on the Caribbean island of Grenada, where he taught math and chemistry at Presentation College in St Georges and operated as VP2GBR. He holds a PhD in materials science and was a tenured member of the academic staff at the University of Warwick for many years, specializing in the characterization and study of defects in crystalline solids. During his tenure there, he also spent quite a lot of time in the United States, mainly doing research at Brookhaven National Laboratory (BNL) and SUNY (Stony Brook) on Long Island. Dave has never entirely given up the use of tubes and AM, and following his discovery in about 1986 of the old club station at BNL, which was still crammed full of old Hammarlund, Johnson, and Collins tube gear from the 1950s and '60s, he became an ardent vintage radio enthusiast, collecting and restoring old tube transmitters and receivers. After working part time for several years, he decided to fully retire in 2010 and devote more time to his other interests.

Notes

[1]The filter design and circuit analysis software from AADE can be downloaded from **www.aade.com**.

[2]Dave Gordon-Smith, G3UUR, "Ceramic Resonator Ladder Filters," *QEX*, March/April 2007, pp 55-58.

[3]Anatol I. Zverev, *Handbook of Filter Synthesis*, John Wiley & Sons, Inc, 1967, ISBN 0 471 98680 1.

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Ray Mack, W5IFS

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SDR: Simplified

We look at an alternative development platform and DSP generated DDS (direct digital synthesis).

An Inexpensive DSP Evaluation Kit

We started with the Blackfin Stamp because it was the most cost effective evaluation system at the time. It is still the most capable platform I am aware of. There is an inexpensive but less capable board available from Texas Instruments, however, that looks like an interesting experimentation platform. TI started a promotion right in the middle of our publication cycle that offers the board and tools for $55 until October 24. (That is probably too late for readers of this edition, but perhaps they could extend the offer or repeat it later.) After that, the price goes back to $99, which is still a pretty small price for such a package. The TMS320C5535 eZdsp board (TMDX5535EZDSP) can be purchased from the TI eStore at the reduced price (or full price after the promotion ends).[1] If you miss out on the promotion, it is also available through distribution. Mine arrived just as I was starting to write this installment, so I have not gotten all the way through my evaluation.

The eZdsp board is manufactured for TI by Spectrum Digital. It contains a TMS320C5535 DSP chip along with a USB audio device and an embedded JTAG debugger that connects to your development PC by way of a USB cable. The C5535 is a low power DSP intended for use in USB sound card applications or other limited power DSP and microcontroller applications. The evaluation board has numerous facilities that make it a candidate for soft rock types of systems. The Codec can produce 192 kHz stereo record or playback, which fits well with the *PowerSDR* software.

There are a few features that make the system worthwhile even at the regular price.

[1]Notes appear on page 36.

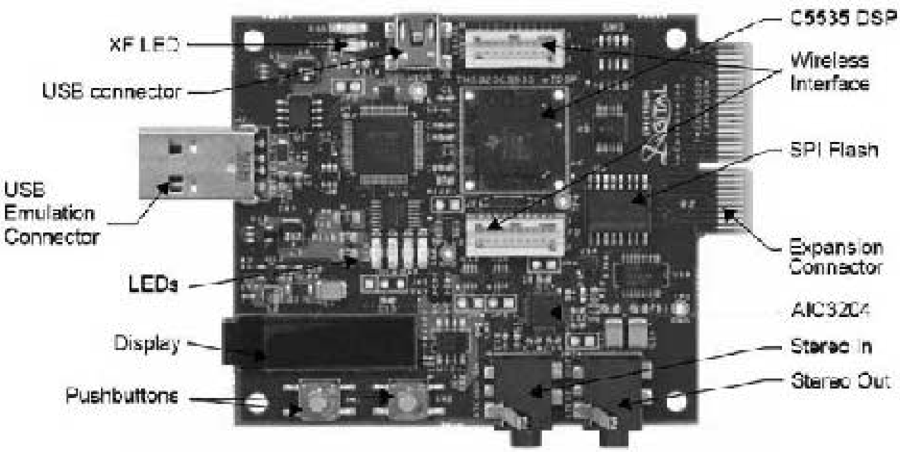

Figure 1 — This photo shows the Spectrum Digital Technologies eZdsp evaluation board for the Texas Instruments TMS320C5535 DSP chip. The various connectors and controls are identified.

It ships with a Lenovo microphone plus stereo earbuds, a complete DSP software development system (a time limited version) on DVD, a 2 GB memory card, and a very small Organic LED graphic display. Shipping and handling is included in the price. Although shipping method is not mentioned when purchasing, mine arrived by Federal Express in a very short time. I am convinced this product is a loss leader for TI to get us to design in their products. I won't argue with their plan! Figure 1 is a photo of the board with pointers to the various resources on the board.

When we did the Blackfin Stamp installation we ran into numerous issues where the installation did not work as expected. This install is no different. I will walk you through the installation since the Quick Start guide left a few things to your discretion that will make the process difficult, and some things just did not work. The Quick Start instructions ask you to install the microSD card, but mine came already installed. Then you should connect the USB A/USB mini B cable to the evaluation board and then to the computer. *Windows XP* and *Windows 7* gave me errors installing the hardware the first time I attached the USB connector. I removed the USB connector and reinserted it. This time it installed generic drivers and then told me that my hardware could work faster if I plugged into a USB2.0 port. You can ignore that warning since it will work (more or less) with USB 1.1. There is a driver on the DVD for 32 bit and 64 bit operating systems under <DVD>:\SpectrumDigital\usbdrivers. You can install that driver, but I have an *XP* laptop and a *Windows 7 Home Premium* desktop where the driver installed and the system never recognized the new driver.

The Quick Start instructions have you

Glossary

Sinc — a shorthand name for the function sin(x)/x

DDS — direct digital synthesis

JTAG — Joint Test Action Group. They define a way to access the internal devices of an IC. It is one of the best ways to connect to a CPU for software debugging.

Codec — ***Co**der-**dec**oder*. In general this refers to an IC that combines both a DAC and ADC for use as audio in and audio out. The device on the evaluation board uses the I2S serial protocol to transfer data in and out of the DSP.

DDS-FM Transmitter Code

```
#include <stdio.h>
extern get_and_display_sound_card_data(void);
extern setup_record(void);
extern setup_play(void);
extern int get_audio_sample(void);
extern play_audio_sample(int);
extern int sine_table[16384];
int deviation;
int frequency;
int accumulator = 0;
int play_sample;
int mic_sample;
int sine_index;
#define ACCUMULATOR_LIMIT 0x3ffff  // 18 bit accumulator
#define HALF_SINE 0x4000
int main(void)
{
   get_and_display_sound_card_data();
   setup_record(); // call the Windows APIs necessary to
                   // connect to the sound card microphone input
   Setup_play();   // call the Windows APIs necessary to
                   // connect to the sound card line out
   printf("Enter the deviation value:");
   scanf("%d", &deviation);
   printf("Enter the Frequency value:");
   scanf("%d", &frequency);
   while (1)
   {
      mic_sample = get_audio_sample();
      mic_sample *= deviation;
      accumulator += (mic_sample + frequency);
      if (accumulator > ACCUMULATOR_LIMIT)
      {
            accumulator -=ACCUMULATOR_LIMIT; // make accumulator
                                             // roll over
      }
      sine_index = accumulator >> 4;
      if (sine_index > HALF_SINE)
      {
            sine_index -= HALF_SINE;
          play_sample = -sine_table[sine_index];
      }
      else
      {
            play_sample = sine_table[sine_index];
      }
      play_audio_sample(play_sample);
   }
}
```

play the demo file on the DVD. When you use a media player such as *Windows Media Player*, the small display gives a spectrum display showing the sounds that are playing. I had to use streaming in *Internet Explorer* with **trueoldies.com** rather than *Windows Media Player* on the stubborn *XP* machine. The sound card implementation seems to be quite good if you can get *Windows* to use the USB 2.0 driver. I have an album of 80s songs that I converted to MP3 that plays with skips and pauses on my eMachines Netbook internal sound card. It plays correctly, however, on the netbook/USB card combination. The spectrum display is meant to show off the hardware FFT capability that is embedded in the TMS320C5535. The update is rather slow, but I suspect that it is a consequence of *Windows* or the display driver rather than slowness of the FFT. I am not familiar with the internal code of *PowerSDR*, but I understand that it uses FFT and inverse FFT for filtering rather than implementing continuous time FIR filters. The embedded FFT would be a big advantage for such a system. SDR is one of the features that TI mentions for this DSP. The FFT does eight to 1024 (powers of 2) real or complex conversions.

Once you have tried the USB sound card demo, you will install the *Code Composer Studio*. Launch the install program at <dvd>:\setup_ccs_4.2.2.00033.exe. You should take notice that the software will install with a 30 day limited license (more on this later). The Quick Start guide does not mention that you have to decide which type of installation to perform. I chose the Platinum Edition for installation and then deselected all but the C55X products. The next screen lists the products to install. **Be sure to unselect XDS560 or the install will stall indefinitely.** Once the install begins, the window will show the pieces in a list. A pop up window will show progress for each piece. You will need to attend to the install in order to confirm installation for several pieces. The pop up windows occur correctly, but the check marks for each piece in the initial window do not keep track.

Now you have *Code Composer Studio* installed on your system. All of the items you need for this board are part of the *Eclipse* and *Apache* system which are publicly licensed. That means they are free. The next step is to open *Code Composer* from your desktop since the installer does not put anything in your "Start Menu". Select the "Help" menu and then "Licensing Options…" Select the "Register" button which will open a window in your Internet browser at the TI licensing site. You can register for a TI account for free, which will allow you to get a license for the Platinum Edition with XDS100 support. You need to fill out all of the forms and agree to the licensing items. The last step is to give the site one of the MAC addresses from the licensing dialog. You can also license the software to another computer by giving its MAC address. Use the command "ipconfig / all" in a command window to get your MAC address (displayed as Physical Address.) I licensed the USB Wireless N adapter attached to one of my desktop machines, so I can move the software to any machine along with the USB adapter. The site will send you the licensing file to your email. Save the file to a convenient directory. Select the license file in the license dialog and complete the process. The software is now permanently licensed.

Spectrum Digital provides the code platform for the *Windows* part of the system. You install this from <dvd>:\DemoAudio\ C55_Connected_Audio_Framework_EZDSP-01.51.00.03-Setup.exe. You can rebuild the software by opening a command window and running click2build.bat. Unfortunately, if you choose a directory other than "Program Files" as I always do, the batch file will not work. I consider this to be really rude of Spectrum Digital. I tried to edit the batch file in *Code Composer*, but attempting to open a batch file for editing runs it instead. This is also pretty rude behavior. I had to use my favorite code editor to add the following two lines to click2build.bat to match the directory to which I installed my software: C:\TI:

Set CCS4_DIR="c:\ti\ccsv4"

Set HEX55= %CCS4_DIR%"\tools\compiler\c5500\bin\hex55.exe"

These two lines allowed the build to run correctly.

TI sells a version of Platinum Edition for $795 that includes support for XDS510, XDS560 or MSP430 JTAG debugging as well as a subscription for annual support. The XDS100 debugger is implemented in dedicated hardware from Blackhawk on the development board. The XDS100 debugger connects to the PC via USB on the evaluation board. Blackhawk also produces a standalone XDS100 debugger for $99 that supports all but the Stellaris product line. I am looking at the TI line for multiple DSP and microcontroller projects. I will keep you informed regarding how useful they really are.

Direct Digital Synthesis in Software

There are a few different ways to generate a sine wave in software. The one piece that is required in each method is the phase accumulator. *Accumulator* is just a computer term for a register, where the value in the register is acted upon (add, subtract, multiply, divide) by another value and the result is stored back in the original register. A DDS phase accumulator holds the present phase value of the sine wave you are generating. The value of the phase accumulator is a stepwise approximated sawtooth waveform that takes values

Table 1
32 Hz Sine Wave Actual Values and Integer Output Values for DDS Generator

Sample	*Actual*	*DDS Out*	*Error*	*Sample*	*Actual*	*DDS Out*	*Error*
0	0.0	0	0.0	16	0.0	0	0.0
1	6392.5	6393	0.5	17	–6392.5	–6393	0.5
2	12539.4	12539	–0.4	18	12539.4	–12539	–0.4
3	18204.4	18204	–0.4	19	–18204.4	–18204	–0.4
4	23169.8	23170	0.2	20	–23169.8	–23170	0.2
5	27244.8	27245	0.2	21	–27244.8	–27245	0.2
6	30272.8	30273	–0.2	22	–30272.8	–30273	–0.2
7	32137.4	32137	–0.4	23	–32137.4	–32137	–0.4
8	32767.0	32767	0.0	24	–32767.0	–32767	0.0
9	32137.4	32137	–0.4	25	–32137.4	–32137	–0.4
10	30272.8	30273	–0.2	26	–30272.8	–30273	–0.2
11	27244.8	27245	0.2	27	–27244.8	–27245	0.2
12	23169.8	23170	0.2	28	–23169.8	–23170	0.2
13	18204.4	18204	–0.4	29	–18204.4	–18204	–0.4
14	12539.4	12539	–0.4	30	–12539.4	–12539	–0.4
15	6392.5	6393	0.5	31	–6392.5	–6393	0.5

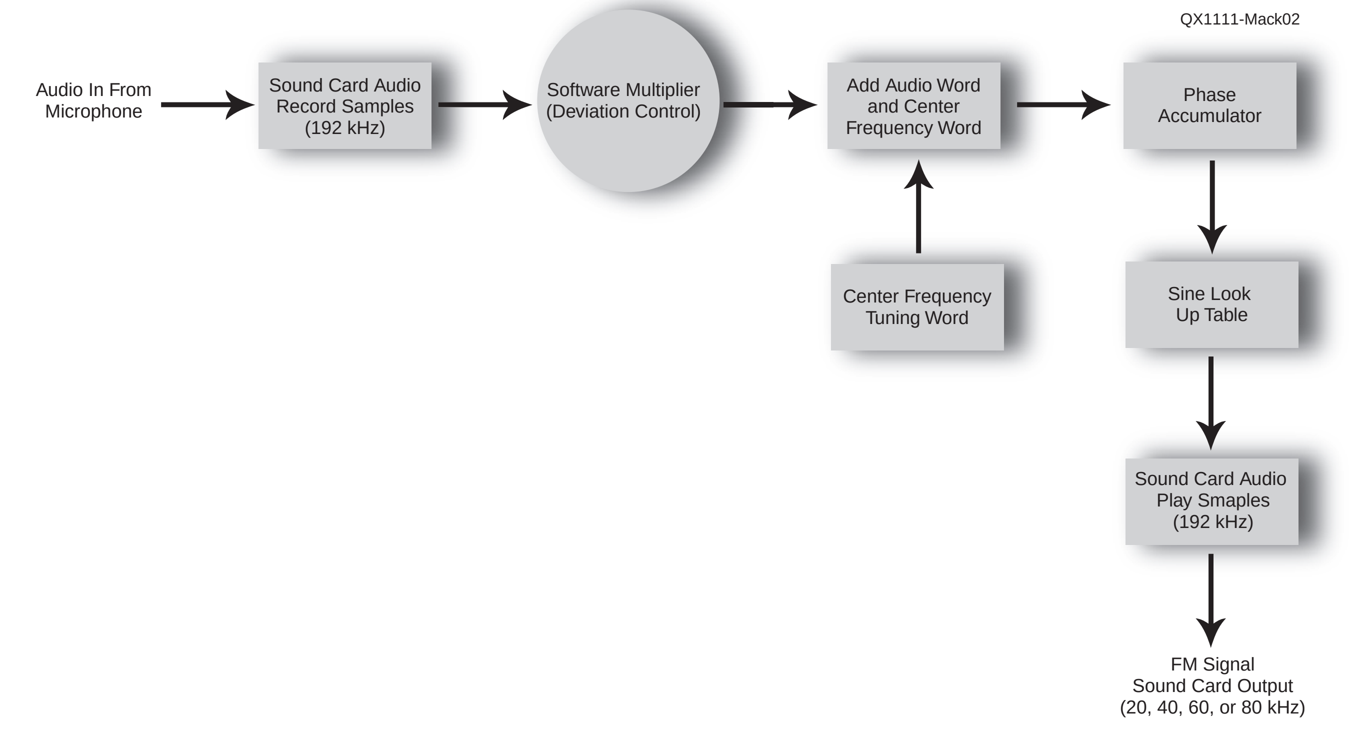

Figure 2 — This block diagram is the simple implementation of an FM transmitter described in the text.

from 0° up to just less than 360° and then the value goes back to near 0° and repeats. The phase change per sample is always a linear change.

Let's look at two simple examples. We will set our sample rate to 1024 samples per second and design our phase accumulator to have 5 bits. That means that our phase accumulator will have the values zero through 31. If we add one to the accumulator for each sample, the value will count up every value from zero to 31 and then return to zero and repeat. That means that we will produce one full cycle every 32 samples or 1024/32 which is 32 Hz. If we add three to each sample, the system counts starting at zero up to 30 and then rolls over to one. It counts up to 31 and then rolls over to two. It counts up to 29 and then rolls over to zero. This yields a frequency of 1024 × (3/32) = 96 Hz. So, we see that the equation for a DDS is:

$$\text{Frequency} = f_S \times (N / register_{max})$$

where f_S is the sample rate in Hz, N is the phase increment, and $register_{max}$ is the largest value of the register plus one (remember, we start counting at zero). We can use the entire range of the DSP data word to minimize quantization error. For a 16 bit DSP, that means the peak value can be 32767 and the negative peak will be –32767. Table 1 shows the actual value of a sine wave along with our integer output value for all 32 samples of a 32 Hz sine wave. The error is 96 dB below the carrier when we have a 16 bit word size. A sine wave is odd symmetric, so we can take advantage of the identity that sin (–*x*) = –sin (*x*). The symmetry lets us use half as much storage for the sine wave look up table as needed for one full cycle; we use positive values from the table for 0° through 180° and negate each value for 180° through 360°. A 16 bit accumulator will require 64 KB of memory to implement 16 bit data values. We do not require a phase table entry for every sample if we are willing to deal with significant Sinc droop and more quantization noise. It is possible to use only some of the bits of the phase accumulator to generate the sine approximation. An 18 bit accumulator will give 0.732 Hz for each increment of the tuning word with a 192 kHz sample rate. We can implement the waveform array as 16384 words (32 KB) by shifting each accumulator value to the right by 4 bits (equivalent to dividing by 16). On modern CPUs, such a shift is a single instruction and usually completes in a single instruction cycle.

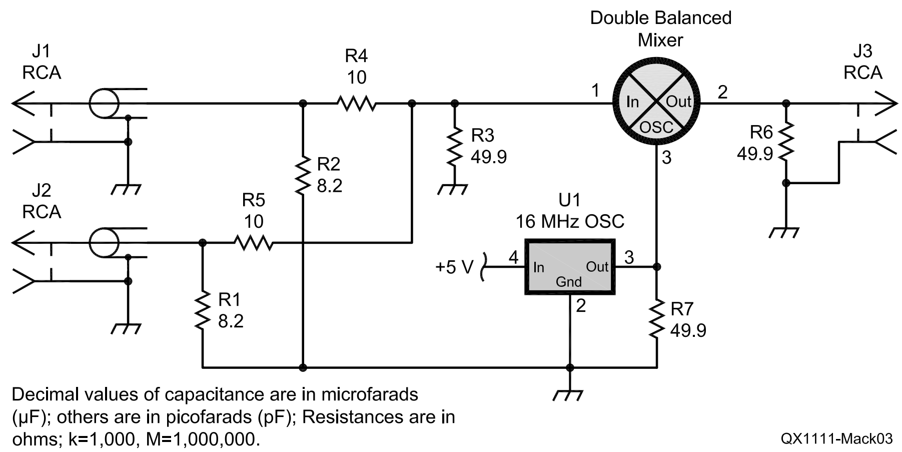

Figure 3 — The schematic diagram of the simple FM transmitter is represented here. This circuit is implemented in DSP using the evaluation kit and computer sound card.

Figure 2 shows the block diagram of an FM transmitter using DSP to produce the carrier, and using addition to create true FM from the audio input. The schematic diagram in Figure 3 is the actual circuit I am using. I imagine that nearly every reader has a two meter FM rig of some sort, so the system is designed to translate the sound card output up to two meters. The oscillator shown is 16 MHz, which multiplies in the double balanced mixer to 144 MHz. If you have a high-output signal generator, you can substitute that for the 16 MHz oscillator. We can generate an FM signal in the DDS by simply adding or subtracting a small value that corresponds to the audio voltage to the tuning value used for the DDS accumulator. The frequency deviation is adjusted by multiplying the gain applied to the audio signal. This method produces true frequency modulation.

The code in the sidebar implements the FM transmitter using the PC sound card. The code starts by displaying the parameters of the sound card, so you can abort it if the sound card cannot sample at 192 kHz. Next, it asks you for the desired output frequency in kHz (20, 40, 60, or 80). The last operation is to input the deviation value. This is a number from 2 to 10 that is used to scale the audio just as an analog deviation control would. As soon as you enter the deviation, it will begin generating the FM signal. This code is just the main loop and omits all the details of connecting to the sound card through *Windows*. The full code is available for download from the *QEX* files website.[2]

Notes

[1] You can order the Texas Instruments TMS320C5535 eZdsp board at the TI eStore. Go to **www.ti.com/tool/tmdx5535ezdsp** to order the kit.

[2] The complete program file to implement the simple FM transmitter on the DSP board is available for download from the ARRL *QEX* files website. Go to **www.arrl.org/qexfiles** and look for the file **11x11_Mack_DSP.zip**.

QEX

Letters to the Editor

Select IEEE C95 Standards Available at No Charge

Hi Larry,

I received an e-mail from IEEE, about the availability of some IEEE C95 Standards documents, with regard to safety levels with respect to human exposure to RF electromagnetic fields. I suggest that the availability of these documents be noted in *QEX*. I have copied parts of that announcement below, for your readers.

— 73, John Montague, W9JM, ARRL Technical Advisor, 818 Adger Rd, Columbia, SC, 29205; **montague@IEEE.ORG**

Get IEEE C95™ STANDARDS: Safety Levels with Respect to Human Exposure to Radio Frequency Electromagnetic Fields

Over the past few decades applications of electrical and electromagnetic energy continue to improve our quality of life, e.g., electric lighting, radio and TV broadcasting, microwave ovens, magnetic resonance imaging (MRI), mobile telephones, Wi-Fi, and Smart Meters, etc. Occasionally questions regarding safety of new technologies arise following their introduction.

For more than 50 years, IEEE SCC39 (International Committee on Electromagnetic Safety-ICES) and its predecessor (ANSI Accredited Standards Committee C95) have been developing recommended exposure limits and assessment methods to help protect against established adverse health effects associated with exposure to the electric, magnetic and electromagnetic fields produced by these technologies.

Because of interest raised by the public and others, the following IEEE C95™standards are now publicly available at no cost through the IEEE Get Program:

• IEEE C95.1TM-2005 — Safety Levels with Respect to Human Exposure to Radio Frequency Electromagnetic Fields, 3 kHz to 300 GHz.

• IEEE C95.3™-2002 — Measurements and Computations of Radio Frequency Electromagnetic Fields with Respect to Human Exposure to Such Fields, 100 kHz to 300 GHz.

• IEEE C95.3.1™-2010 — Measurements and Computations of Electric, Magnetic, and Electromagnetic Fields with Respect to Human Exposure to Such Fields, 0 Hz to 100 kHz.

• IEEE C95.6™-2002 — Safety Levels with Respect to Human Exposure to Electromagnetic Fields, 0 to 3 kHz.

• IEEE C95.7™-2005 — Radio Frequency Safety Programs, 3 kHz to 300 GHz.

To access these standards, visit: **http://standards.ieee.org/about/get/index.html#getC95**.

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Hi John,

Thank you very much for sharing this information with our readers.

— 73, Larry Wolfgang, WR1B, QEX Editor; **lwolfgang@arrl.org**

VSC-X: A Virtual Serial Cable to Remotely Program Your Mobil Radio (Sep/Oct 2011)

Hi John,

I had purchased a couple of the XBee Pro modules and was just about to breadboard my own version of a Virtual Serial Cable when your timely VSC-X article came along in the Sep/Oct *QEX*. Not wishing to reinvent the wheel, I followed your lead, and now have a nicely functioning set of VSC system! I'm realizing a range of about 1/3 of a mile in the real world using the whip antennas.

I plan to build additional sets for some upcoming projects, and wish to thank you for making my job easier by making the circuit boards available. Thanks!

— 73, Tom Tcimpidis, K6TGT, 12569 Mendel Dr, Granada Hills, CA 91344; ***k6tgt@arrl.net***

Transmission and Reception of Longitudinally Polarized Momentum Waves (Jul/Aug 2011)

Dear Larry,

On Aug 11, I toured ARRL HQ, and the tour guide, Skip Colton, W1FTE, stopped by your office. I have received *QST* for a long time but had gotten *QEX* only occasionally at radio stores. I expressed an interest and you were nice enough to round up a copy of the August issue for me. I figured I would subscribe eventually. On reading the article by Bob Zimmerman, NP4B, I got very excited about the subject of his article and have already exchanged emails with Bob, who was at Aricebo, in Pureto Rico. He and I had been there at the same time once back in 1987. So I am hooked, and my *QEX* subscription was submitted just a few minutes ago. I look forward to receiving it regularly now and for a long time. My background is electronics engineering (retired), hence my interest.

— 73, Chuck Lahmeyer, AF6GT, 6319 Roling RdJefferson City, MO, 65101; ***clahmeyer@earthlink.net***

Hi Chuck,

It was nice to meet you that day, and I am glad you found the sample issue of *QEX* so interesting, and especially the article by Bob Zimmerman.

It turns out that Bob's article, "Transmission and Reception of Longitudinally Polarized Momentum Waves," or vector potential waves as Bob likes to call them, generated more interest than probably any other *QEX* article in recent memory! Within a few days of the July/August issue being mailed, Bob received over 20 e-mails. I had been somewhat uncertain how much interest that article would generate. Originally Bob had submitted a manuscript that went into much more mathematical detail, and included more information about the experiments he had been involved with. I encouraged him to write a shorter article that focused on the concept rather than the full details, which resulted in that July/August article.

I found it very interesting to read the e-mails that I received about that article, as well as some of the ones Bob received. There seemed to be two distinct types of correspondence. A lot of readers wanted to learn more, and asked Bob to send them copies of his paper from *Modern Physics Letters B*. Several people became interested in duplicating Bob's equipment setup, and in trying to replicate his experimental results, or even in trying some new ideas for the reception of these VP waves. For example, one reader suggested that Bob consider using graphene in his detector. Graphene is a material that consists of a single layer of carbon atoms that has been in the news recently. Electrons are conducted much faster in graphene than in semiconductor materials. This led to some experiments with detectors made from graphite blocks. Bob expected the graphite blocks to produce a much more sensitive detector, but the experiments did not show that to be the case.

There were also a number of readers who wrote in disbelief, claiming that Bob did not understand basic physics or that he was actually detecting transverse electromagnetic (TEM) waves — normal RF. One writer indicated that vector potential was simply a mathematical tool used to help solve Maxwell's equations, but that it was not a physical reality.

Bob has been sending me frequent updates with regard to his further research as well as his correspondence with several other hams. He reports that there are currently at least three individuals or groups of people in various stages of building equipment and antennas and conducting experiments with vector potential radiation. I hope to have more news, and perhaps a follow-up article in a few months.

Here is a sampling of the correspondence Bob received.

— 73, Larry, WR1B

Hello Robert,

First, I'd like to congratulate you on your very clear and well written article describing vector potential radiation. Professionally, I am a signal integrity engineer, so I only casually deal with Maxwell's equations when doing finite element simulations of EMI, PCBs, connectors, and so on for gigabit serial data interfaces. Over the years, I thought I finally developed a fair understanding of Maxwell's equations, but your article totally destroyed that thought. I found your article so intriguing, that I stayed up almost all night researching the Aharonov-Bohm Effect and found myself learning more than I ever thought I would care to know about physics, quantum physics, and gauge theories, all while trying to bone up on the details of Maxwell's equations. While I've gained only a rudimentary understanding of what's going on, this phenomenon looks like fertile ground for future experimentation, as you mention in your article. I hope this conveys at least some of the excitement I see for this new communications mechanism. Thanks for bringing this phenomenon to the attention of the Amateur Radio community.

— 73, Ray Pavlak, N1GEX, 60 Beverly Hill Dr, Shrewsbury, MA, 01545;
RayPavlak@aol.com

Hello Robert,

I had to push my jaw back up off the floor after reading your article in the July/August issue of *QEX*. I have used the vector potential for mathematical analysis but it never occurred to me it could be used directly for communication. My thanks to you and Dr. Nikolova for this practical description. I have every expectation of trying this relatively simple and certainly unique mode, and I would be most interested to read the paper you mention as well. Thanks again.

— 73, Elwood, WBØOEW, 8274 N Sunset Ranch Loop, Tucson, AZ, 85743;
ecdowney@clearskyinstitute.com

Hi Bob,

I love the article in the July/August issue of *QEX*. The phenomenon you describe has been known in some circles for many years as "scalar waves." There has been a lot of mystique surrounding them, mainly because no one in the general public really knew how to transmit or receive them, and assisted by wild stories by people like Colonel Tom Bearden.

The quaternion forms of Maxwell's equations in his original work are supposed to best describe vector potential waves. Later versions of his work used vector calculus instead of quaternions, and something was lost in the translation. I checked out the original 1874 copy of *A Treatise On Electricity And Magnetism* from the University of Illinois library a few years ago but they only let me have it for a week and I didn't get to spend enough time with it.

I've played around with various detectors for receiving, involving neon bulbs and the Barkhausen effect detector (Bob Shannon). The Greg Hodowanec detector is also supposed to detect these waves. I heard some unusual things on these detectors now and then, but I didn't really know what I was receiving, and I had no idea how to transmit them.

You've certainly cleared that up. I will have to make a trip to the hardware store and do some experiments with my 1296 MHz equipment.

I wonder if many others really grasp what you've done here?

— 73, Zack Widup, W9SZ, 1003 E Washington St, Urbana, IL, 61801;
w9sz.zack@gmail.com

Robert,

I read your article, "Transmission and Reception of Longitudinally-Polarized Momentum Waves," in QEX, July/August 2011, pages 31-35, with great interest. I have several observations, questions and/or conceptual difficulties that perhaps you can respond to with either additional information or provide references. These questions/difficulties can be categorized into these areas: the receive antenna gain horn, the physics of the receiver, the transmit antenna description, and link budget considerations with experimental procedures that excludes other EM effects.

Receive antenna gain difficulties — A 20 dBi receive horn formed from sheet metal was needed to achieve the full 1500 meter range. Earlier in the article, however, you stated that: "VP waves cannot be received with metallic antennas" and "Electrons in copper and other metals can never move at velocities greater than about 3 mm/s…. This is entirely too slow to couple to vector potential radiation." Therefore, in the absence of conventional E or H fields at the receiver, how can a metal horn provide gain for the VP waves?

Physics difficulties in the receiver — The terms used and the physical effects described do not seem to match any of the conventional meaning or described effects. You indicated that the basis for detection of the VP waves was that: "The A vector potential will couple linear momentum to free-electrons in its path." and said that this was the Aharonov-Bohm Effect. I skimmed a fair number of documents resulting from a search of that term as suggested by the article and none of them associate momentum coupling to that effect. The Aharonov-Bohm Effect is related to a phase shift within the wave-function of the electron, which can only be observed within certain kinds of electron self-interference experiments. The phase function in question is not in any way related to the phase of bulk currents. Also, in all the references I skimmed, there was no indication that the Magnetic Vector Potential A would couple momentum into free electrons directly (despite it having units of Momentum per unit charge); instead momentum only appears in a differential equation (search terms: magnetic vector potential Hamiltonian) and changes in electron motion only appear as combinations of both spatial and temporal derivatives of A. When the continuity condition that relates accumulated electric charge to current flow is applied to these relations (the "Lorenz Gauge Condition" as indicated by every EM textbook I consulted) the changes in electron motion can be shown to be equivalent to responses to E and H fields — only — without any extra left over.

The upshot is, according to conventional EM theory (and even quantum electrodynamics) even though it is possible to create conditions for a Magnetic Vector Potential A without E or H present, there will be no change to electron motion influencing current under those conditions. To produce changes in electron motion, E, H or both must be present. In summary: direct coupling of A to electron momentum is not called the Aharonov-Bohm effect and the conventional literature that I skimmed indicates that even in the presence of A there still has to be E, H, or both to induce changes in electron motion. Consequently if such an effect exists it would be called something else. Do you have references or other information on experiments that would show the existence of such momentum and bulk current or plasma coupling when E and H are both conclusively demonstrated to be absent?

Transmit Antenna — The transmit antenna will probably radiate far more TEM radiation than suggested since it works on a different principle than indicated. The transmit antenna is a wire that is supposedly shielded by the 7 inch diameter stove pipe circular waveguide and consequently only allows VP radiation out the end. When calculating the magnetic vector potential "A," however, all of the currents in the antenna structure have to be considered, not just the wire "probe." If the shielding actually worked as described then it is far from certain that there would be any axial VP radiation either. This is because the currents in the walls of the stove pipe would cancel the A vector in the axial direction also yielding no VP radiation. Due to the size of the circular guide, the wire is not really "shielded" and it appears that significant amounts of conventional TEM radiation will be radiated out the end of the pipe. The "probe" extends the full length of the stove pipe so this is not really a probe in a wave guide but the center conductor of a large diameter coax cable. The dominant mode in a coax is TEM of course and since the velocity factor is close to 1 for air dielectric, the 69 cm long stove pipe is close to 3 wavelengths long.

In the absence of other modes this would be more likely to appear as an open circuit rather than a short circuit. A TM_{01} mode should not be possible in this configuration, since it requires an axial component of E field, which is prevented by the center wire. Far from completely shielding the wire, the TEM coax mode can radiate conventional TEM waves out the end of the coax line. This is a common illustration or problem in advanced EM textbooks. A quick approximate calculation based on inserting reasonable assumptions in the formulas found in textbooks suggests that the total radiated power from this mode might only be down about 13 dB from the forward power in. This is probably off by quite a bit, however, since the interior circumference of the stove pipe is more than a wavelength and the calculation is valid only for small (in terms of a wavelength) coax cable.

It is probable that calculations based on annular slot antennas would instead show that the total power radiated would be limited only by mismatch between the 50 Ω transmitter and the antenna system.

Although this mode ideally does not transmit energy along the axial direction as it would be off to the side, reflections and scattering off of other objects could redirect this energy towards the receiver. Other modes could radiate even better.

Even though the TM_{01} mode is not possible, the TE_{11} mode is possible and has a cutoff frequency around 1074 MHz. Irregularities in the configuration of the wire in the guide are likely to launch this mode in addition to the TEM coaxial mode. The TE_{11} mode field pattern does result in transmit energy on the axis of the stove pipe. Transmit power by this mode would also seem to be limited only by impedance mismatch. Upshot: even if VP radiation were to be shown to be the mode of operation of the receiving unit, it is indeed very well advised to "…operate in the Amateur Radio allocated frequency bands…" since it is quite possible that significant amounts of transmit energy would be radiated as conventional EM waves. This leads directly to the next question.

To what extent were conventional receiving antennas used in the experiments to characterize how much conventional EM waves were present? That is: was it conclusively shown that it was not possible to receive the transmissions by ordinary means? The value of this would be as a "control" in the experiment. The inability to detect any conventional TEM radiation would go a long way in demonstrating that something different was happening within the plasma tube receiver.

Without that kind of control demonstration, it is easy to account for the results as being from conventional TEM radiation. Quick free space calculations show that at 1296 MHz using isotropic (0 dBi gain) antennas at both ends of the link would result in path loss of around 89 dB for 500 meter separation and 98 dB loss for 1500 meter separation. With a 1 W transmitter through a 3 dB pad (0.5 W) and assuming a typical handheld transceiver 0.2 microvolt receive sensitivity, there would be in excess of 59 dB of margin at 500 meter separation and about 50 dB excess at 1500 meters. So between 50 and 59 dB more loss could be inserted into a system using 0 dBi gain antennas and still communicate. It is quite easy to believe the combined system of transmitting and receiving antennas described in the article with their mismatch losses, inefficiencies, and scattered RF energy could provide that level of attenuation.

In summary — no one would be happier than me to learn of a "new" way to do radio. There are many concerns and questions, however, that I believe should be addressed. Even if there is no conventional theory to explain the results, convincing tests would need to include clear demonstrations that conventional E and H fields were far too weak to produce the results observed.

— 73, Ray L. Cross, WKØO, 4018 Council St NE, Cedar Rapids, IA, 52402;
radio.wk0o@*raylcross.net*

Hi Larry,

I am having so much fun corresponding with readers about my vector potential article. I have found a most interesting historical paper you will love: "Maxwell and the Vector Potential." This paper follows Maxwell's thought process as he wrote his various articles. Maxwell fully believed vector potential to be a real physical quantity. It was Heaviside, and to some extent Hertz, who completely diminished its importance and brought on the current-day thinking. Access to this paper is available from Chicago Journals History of Science Society at **www.jstor.org/stable/228226.**

Maxwell believed his introduction of vector potential (from ideas first developed by Michael Faraday) to be his most important contribution to electromagnetic theory!

With regard to the letter from Ray Cross, WKØO, I wrote back to Ray with answers to his questions. The main concern of his letter seems to be the existence of collateral TEM radiation at the receive site. Of course there will be some TEM radiation there — I know of no way of avoiding this. The receive apparatus is mainly sensitive to longitudinal vector potential radiation, however. This is easily verified by reversing the tube bias current, which always results in a phase reversal of the detected signal. (This would not occur for TEM radiation.) As a phase reference, I used a 1 kilometer roll of fiber-optic cable — it is much cheaper than coax and lower loss as well.

— 73, Bob Zimmerman, NP4B;
rkzimmerman@gmail.com

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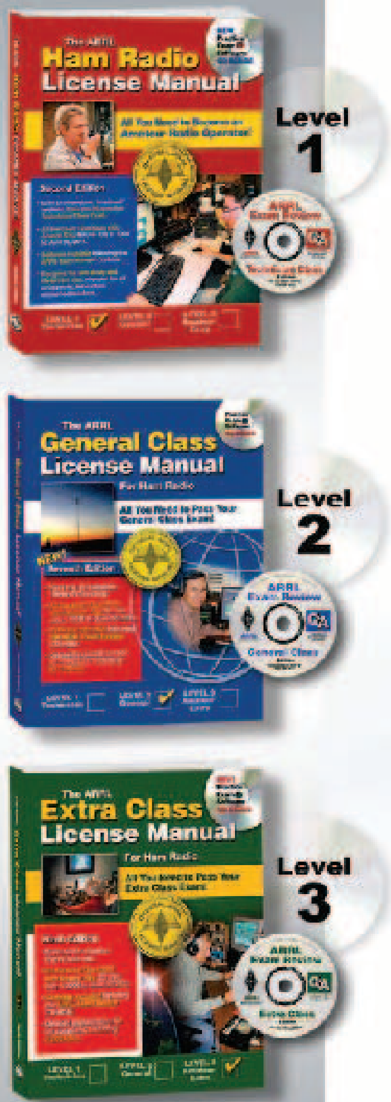

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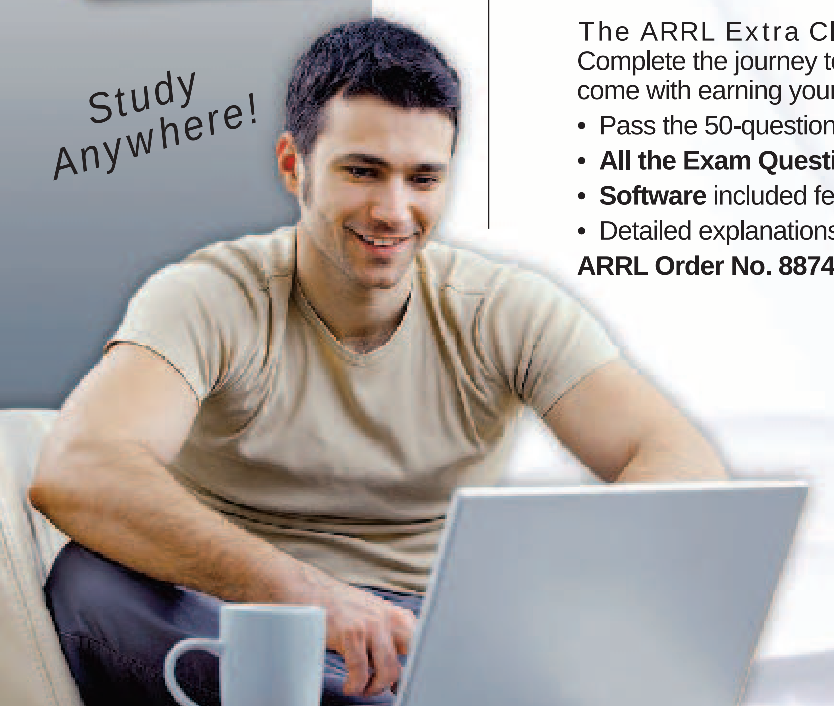

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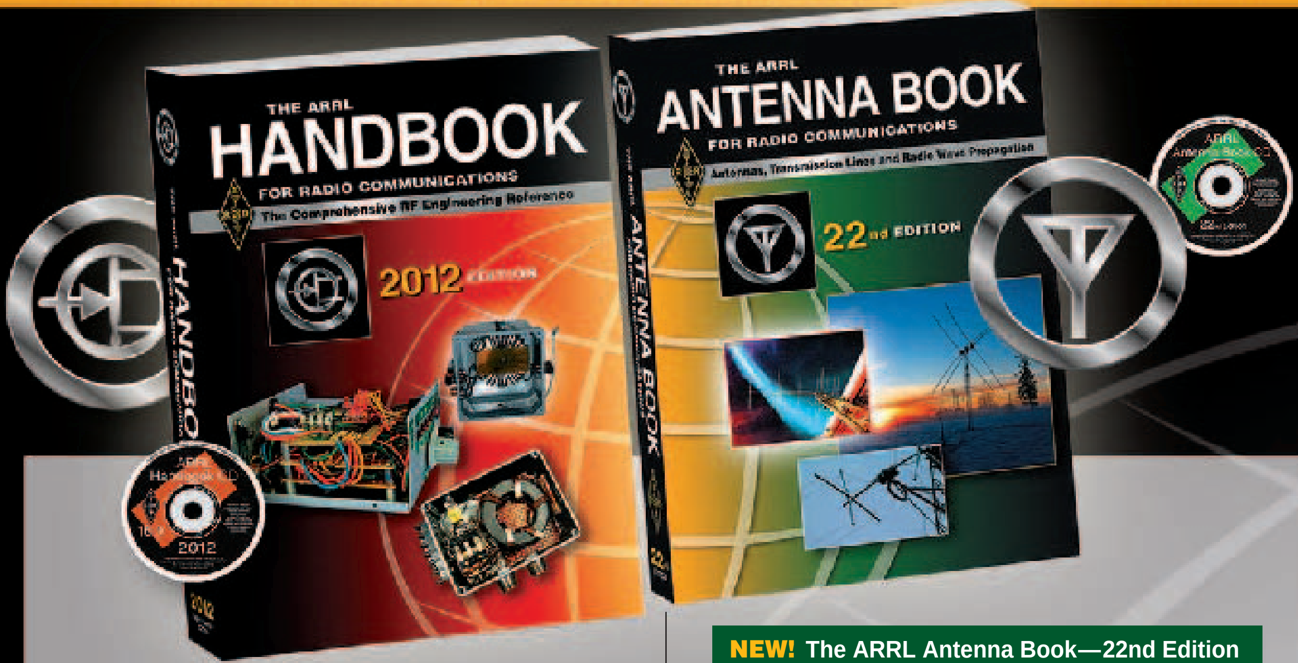

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