

# Kanga US

3521 Spring Lake Dr.  
Findlay, OH 45840  
419-423-4604

[kanga@kangaus.com](mailto:kanga@kangaus.com)  
[www.kangaus.com](http://www.kangaus.com)

## miniR2 Construction Notes

Components (except for surface mount components) are mounted on the side of the board marked "ARS miniR2".

The three surface mount components are taped to the back of the kit label..

The single turn side of T1 is mounted towards the 100 ohm resistor. The 4 turn side goes towards the antenna.

There are three jumpers on the PC board. Two are located between U4 and U5, and the third goes between the pads marked "s" and "f".

There is no pad available for sidetone input. It can be connected to the end of R50 - either above or below the PC board.

The volume control is connected as follows: One side of the pot goes to the pad marked "v", the center (wiper) goes to the pad marked "w", and the third terminal of the pot goes to ground. A spare ground pad near C29, C27, or C25 can be used.

The article (used here with permission) on the following pages is a draft copy of an unpublished article by Rick Campbell, KK7B, on the miniR2. It contains some interesting information on the process that went into the design of the miniR2. The figures, tables, and diagrams that are mentioned in the article were not included with the copy Rick gave me, so they are not included here. Most of the circuits he mentions can be found (with values) in books like the ARRL Handbook, Solid State Design, and Experimental Methods in RF Design.

The miniR2 circuit was first published as part of the article by KK7B in November 1995 QST "A Small High Performance CW Transceiver", pp 41-46. The miniR2 is also well covered in Experimental Methods in RF Design.

Also included with the documentation are two "Technical Cartoons" published in SPRAT by KK7B. They are reprinted with permission. The parts that are mentioned in the two technical cartoons are not part of the miniR2 kit. The Technical Cartoons are included for reference only.

The schematic at the end has been split in two pieces to fit the page size. A single sheet schematic is also included. Note that one new part has been added to the miniR2 that is penciled in on the schematic. It is C 41, a .47 uf coupling capacitor that goes between the wiper of R38 and the input of U6B.

## A 2<sup>nd</sup> Generation Single-Signal Phasing Receiver

Two high performance direct conversion receivers were described in QST a few years ago (1,2). The first, R1, was designed to establish a baseline for high performance in a reproducible direct conversion receiver. The second, R2, added single-signal capability to the R1 using phasing to eliminate the opposite sideband. The R1 and R2 have been very successful receiver designs, with hundreds of each version reproduced around the world.

As stated in the original QST article, the R1 was not intended to be the "ultimate" design. During the past few years, the game of reproducing, modifying and improving the R1 and R2 circuits has attracted some of the best minds in radio engineering. The resulting discussions have been most educational. This paper presents some new understanding of the technical details of the R1 and R2; a few simple modifications to squeeze a little more performance out of the original circuits; and a new "miniR2" circuit board that illustrates evolution of the original phasing receiver in the "lower cost" and "smaller size" directions.

### TECHNICAL DISCUSSION

Some progress has been made in four design areas since the original series of papers was published: mixers; audio gain; filters; and phase shift networks. A new design tool, ARRL Radio Designer software, has also become available. ARRL Radio Designer is an excellent way to study phase shift networks and filters, and circuit files are included in this paper wherever applicable.

### MIXERS

At the end of the R1 paper a discussion of higher-than expected noise figure concluded that the limitation was  $1/f$  noise. This conclusion was supported by quoted  $1/f$  noise specifications for mixer diodes and a comparison of the receiver audio output noise spectrum with the diodes on and off. Wes Hayward then built an R1 circuit and made an improvement in noise figure by changing the audio gain distribution

The improvement Hayward noted would not have occurred if mixer diode noise were the limiting factor. Some very careful receiver audio output noise spectrum measurements were made. Rather than just turning the mixer diodes on and off by disabling the LO, input to the audio diplexer was switched between a room temperature 50 ohm termination and a diode ring mixer with a 50 ohm termination on the RF port and a +7.0 dBm 50 ohm local oscillator. The audio spectra were measured and compared using an HP3582A FFT audio analyzer. There was no significant difference in the shape of the spectrum with the mixer connected to either the mixer or the 50 ohm resistor, The " $1/f$ " spectrum reported in the R1 paper was determined to arise from the difference in mixer IF port impedance with the diodes on and off.. Apparently the  $1/f$  noise in this application is significantly lower than indicated in the HP diode specifications quoted in the original paper [p. 4-69 Microwave & RF Designer's Catalog 1990-1991].

Careful analysis of Hayward's results indicates that a noise figure of 11 or 12 dB should be possible using a direct conversion receiver with a diode ring mixer, grounded base preamplifier, and no RF amplification. That is an improvement of 7 or 8 dB over the reported noise figures of the original R1 and R2. Modifying the audio gain distribution for improved noise figure is discussed in the following section.

## AUDIO AMPLIFIERS

The audio gain distribution for the original R1 circuit was designed to get as much selectivity as early in the receiver as possible. This is simply good engineering practice. One of the engineering decisions was to sacrifice a dB or two of noise figure by keeping the gain before the main selectivity element to a minimum. The exceptional close-in dynamic range of the R1 and R2 is evidence of the success of this approach. Unfortunately, the "dB or two" grew as lossier filters, series filter terminating resistors, and volume controls further reduced the gain of the early audio stages of the receiver. The noise figure of the first R1 and R2 receivers ended up around 17 to 20 dB, rather than the expected 11 or 12. After Hayward reported a significant improvement in sensitivity by raising the gain of the audio preamp stage, the gain distribution was re-examined. It appears that about 10 dB more audio preamp gain is a good compromise between improved noise figure and reduced close-in dynamic range.

For the R1, it is hard to beat Hayward's modification for improved noise figure - just rearrange a few parts and add a resistor and capacitor. The additional gain is adjustable by changing the AC emitter resistor. The original and revised circuits are shown in figure 1. All of the R1 boards sold by Kanga Products have included a minor artwork modification that allows building either version with all parts on top of the board.

Hayward's modification is not quite as attractive for the R2. The RC circuits in the emitter and collector of Q3 in figure 1 are incidental phase-shift networks with a random phase error that is difficult to control because of the notoriously wide tolerances of electrolytic capacitors. If Hayward's modification is applied to the preamp stages in an R2, narrow tolerance 47 uF tantalum capacitors are suggested for both the emitter and collector capacitors. Another concern with the R2 is that the audio phase-shift networks are designed to be driven by a low impedance source. The source impedance after modification is near 500 ohms. A careful computer simulation shows that 500 ohms is low enough that the opposite sideband suppression is not degraded in an R2 system, although the LO phase-shift network may have to be tweaked to 90.5 degrees instead of the textbook 90.

Another option for the R2 board is to add a dual low-noise op-amp audio preamp. This is particularly attractive since all of the inputs, outputs and power supply connections are already on the R2 board. Figure 2 is the schematic of an audio preamp added to an R2-T2 20 m transceiver. The measured noise figure of the receiver, including a double-tuned RF input filter, is 14 dB.

The audio phase shift networks are before the low-pass filter in the R2 circuit. Simply changing the quad op-amps from NE5514 to the lower noise TLO84 or TLO74 will improve the noise figure of the original R2 circuit by a few dB. (*TLO84's are supplied*

*with all the Kanga kits).*

## **FILTERS**

The audio filters in the original RI article are excellent. With ARRL Radio Designer, however, there is no longer any need for advanced amateurs to simply duplicate existing designs. Figure 7 shows three different filter schematics. The first two will fit on the original RI and R2 boards with no modifications. The third is a simpler version that incorporates high-pass and low-pass elements. Table 1 is an ARRL Radio Designer circuit file for all three filters. The equations for the components are given in the schematic. The design procedure is to pick high-pass and low-pass frequencies, calculate the desired component values, find the nearest available components from the Digi-Key catalog, look up the series resistances of the inductors in the Toko data book, and run the program. Once the desired filter response is obtained, a phone call to Digi-key will have the parts on the bench in three days. What if one of the parts is out of stock, (a common occurrence after the RI and R2 articles appeared in OST)? A quick look at ARRL Radio Designer will show the effect of substituting the nearest higher and lower value.

Figure 8 is a double-tuned RF input filter with design equations. This topology, with common powdered-iron toroids and trimmer capacitors, works well from 160 to 6 m. Table 2 is the ARRL Radio Designer circuit file.

Several experimenters have reported great success using DSP audio filters with the RI and R2 circuits. As prices drop and DSP chips become more and more capable, this will certainly be the cost-effective method of obtaining outstanding selectivity with extreme flexibility. As discussed in the audio amplifier section, some additional bypassing is needed in the audio gain stages if the RI or R2 is operated near a computer. A big filter capacitor and good decoupling of the power supply leads are also needed to keep digital noise out of the RI or R2 audio outputs.

There is a philosophical drawback to adding DSP to the RI and R2 receivers. Part of the appeal of simple radio equipment is that a reasonably well educated and experienced radio experimenter can understand the function of every single component. DSP chips are little appliances, and a simple and elegant radio circuit loses something intangible when one is added.

There is nothing inherently bad in appliance technology--in fact, such appliances as a personal computer running ARRL Radio Designer can greatly enhance our understanding of what each component in a simple and elegant circuit does. Human beings are strange and wonderful creatures. A paradigm for the 90s is the mid-level manager who sits at a Pentium machine during his lunch hour designing a wooden boat he will build with his hands.

## **PHASE SHIFT NETWORKS**

Two phase shift networks are used in a phasing receiver or exciter--one at audio and one at radio frequencies. Either or both may use analog, digital, or computational

(DSP) techniques. Traditionally these networks have been an area of great interest among radio amateurs, and this interest is likely to continue as inexpensive, high frequency DSP hardware forces conventional filter SSB generation and demodulation out of commercial equipment. After publication of the R2 and T2 articles in QST, I received considerable correspondence from phasing method experts around the world. Of particular value to builders of simple analog systems like the R2 and T2 were suggestions by Was Hayward, Rodger Rosenbaum, and Byron Blanchard.

Hayward questioned my use of a 'high-pass' topology op-amp phase shift network when the low-pass, topology eliminated the need for a separate 6 volt supply on each op-amp. My principal references for the early R2 work were papers by Oppelt and Breed. Oppelt used low-pass' phase shift networks and Breed used "high-pass" networks. When I began analyzing the operation of the networks, I arbitrarily chose the "high-pass" configuration. For more recent designs I have switched to the "low-pass" configuration, as it eliminates a connection to each op-amp and thus simplifies the circuitry and board layout. There is no other advantage I have found for either configuration. Rosenbaum and Blanchard both questioned the use of high impedance drive to the op-amp phase shift networks in the T2 circuit published in April 93 QST. They are right. op-amp phase shift networks of the type used in the R2 and T2 need low impedance drive. The first T2 prototype used a 2 k amplitude balance pot, but the one I sent to ARRL for photographs had a 10 k pot, so that is what I put in the published schematic diagram. Blanchard also performed a computer analysis that showed that the phase errors I incurred in the published T2 circuit made it impossible to obtain 40 dB opposite sideband suppression. This was puzzling, as I had carefully measured the opposite sideband suppression using a laboratory standard 40 dB pad substitution method and the T2 on my bench clearly had a little more than 40 dB of suppression across the band. I duplicated Blanchard's analysis using PSPICE and verified that the circuit in QST had audio phase shift errors as high as 6 degrees. At this point, the entire theory of phasing SSB generation was derived again from scratch. Instead of using the usual sines and cosines approach, I used  $\cos(sts+ss)$  for  $n$  channels. That way I could determine the effect of phase and amplitude errors anywhere in the system and quantify any advantages to summing more than two channels to eliminate the opposite sideband. The first order of business was to discover why the T2 worked. The answer is that it is possible to perfectly compensate for a fixed phase error in the audio phase shift network by introducing an opposite error in the LO phase shift network. Of course, the phase error in the T2 is not fixed, so perfect compensation can only be achieved for one modulation frequency. By adjusting the T2 for optimum opposite sideband suppression in the middle of the audio passband, as I did, the suppression was falling below 40 dB just about the time the audio bandpass filter was beginning to drop off at the high and low frequency ends of the voice passband. Thus, an LO phase shift of about 93 degrees resulted in the stated opposite sideband suppression of "more than 40 dB across the audio passband."

More computer modeling showed that maximum opposite sideband suppression using the particular RC components in the T2 audio phase shift networks resulted from a drive impedance of about 500 ohms and an LO phase shift near 91 degrees. It is easy to lower the drive impedance on the T2 board by reducing the amplitude balance pot to 1 k and connecting a 2.7 k resistor from each end of the pot to the 6 v supply line. This modification has been included with all T2 boards supplied by Applied Radio Science

and is incorporated in the revised boards supplied by Kanga Products. The modified T2 circuit provides slightly better opposite sideband suppression and is much easier to adjust, as the audio phase shift is relatively constant across the audio passband.

**Notes:**

1. Rick Campbell, KK7B, "High Performance Direct-Conversion Receivers", QST, August 1992, pp 19-28
2. Rick Campbell, KK7B, 'High Performance Single-Signal Direct-Conversion Receivers", QST, Jan. 1993, pp 32-40

**miniR2 Parts List - 5/11/2003**

Qty	Description	Schematic Label	Number on Part
2	51 ohm 1/4w resistor	R2, 8	
1	100 ohm chip resistor	R1	
1	100 ohm 1/4w resistor	R52	
1	470 ohm 1/4w resistor	R40	
5	2.7k ohm 1/4w resistor	R3, 9, 36, 37, 51	
2	3.3k ohm 1/4w resistor	R5, 11	
2	4.7k ohm 1/4w resistor	R45, 48	
2	5.6k ohm 1/4w resistor	R6, 12	
8	10k ohm 1/4w resistor	R4, 10, 39, 41, 42, 43, 47, 49	
2	27k ohm 1/4w resistor	R14, 16	
6	100k ohm 1/4w resistor	R7, 13, 15, 17, 44, 50	
1	1M ohm 1/4w resistor	R46	
12	10.0k ohm 1/4w 1 % resistor	R18, 19, 21, 22, 24, 25, 27, 28, 30, 31, 33, 34	
1	1.52k ohm 1/4w 1 % resistor 1.50k and 20 ohms	R20	
1	11.3k ohm 1/4w 1 % resistor	R23	
1	51.1k ohm 1/4w 1 % resistor	R26	
1	5.23k ohm 1/4w 1 % resistor	R29	
1	23.2k ohm 1/4w 1 % resistor	R32	
1	178k ohm 1/4w 1 % resistor	R35	
1	1k trimpot	R38	102
1	500 ohm pot	vol control	

6	.01 uf 1% matched capacitors. The six .01 caps included in the kit have been matched to within 1%	C16, 17, 18, 19, 20, 21	103
2	390 pf capacitor	C34, 38	
2	.1 uf capacitor	C32, 35	104
4	.12 uf capacitor	C14, 15, 33, 37	124
2	.33 uf capacitor	C2, 9	334
3	.47 uf capacitor	C3, 10, 41	474
2	1 uf capacitor	C30, 31	105
1	.022 uf capacitor	C26	223
1	.033 uf capacitor	C28	333
1	.15 uf capacitor	C29	154
1	.18 uf capacitor	C25	184
1	.27 uf capacitor	C27	274
7	10 uf electrolytic	C5, 6, 12, 13, 22, 23, 36	
5	33 uf electrolytic	C4, 7, 11, 24, 39	
1	100 uf electrolytic	C40	
2	220 pf chip capacitors	C1, 8	
2	1.2 mh inductor	L1, 2	122j
2	33 mh inductor	L3, 4	333j
1	120 mh inductor	L5	124j
2	TUF-1 double balanced mixer	U1, 3	

1	TK2518 balun transformer	U2	
4	2N3904 transistor	Q1, 2, 3, 4	
1	2N5457 FET	Q5	
1	1N4148 (or equiv) diode	D1	
2	NE5514 or TL084 IC	U5, U6	
2	NE5532 IC	U4, U7	

## KK7B SPRAT Technical Cartoon #1

### A Passive Phase-Shift Network to Cover the Whole band

Rick Campbell - KK7B

Published in SPRAT #81, pp 20-22, reprinted with permission

Here is an LO phase-shift network that I've been using with single-band phasing receivers and transmitters lately. The quadrature hybrid (C1, T1, and C2) provides 90 degree phase shift over a wide range, but equal amplitude signals only at the design center frequency. If this phase shift network is used to drive the LO ports of saturating mixers (like diode rings and Gilbert cell ICs) the amplitude balance is not critical, so in practice an entire amateur band may be covered without readjusting the network.

Practical phasing receivers and transmitters need a few degrees of adjustment of the LO phase-shift network to compensate for small phase errors elsewhere in the system. I added a pair of identical 1/8th wavelength lumped-element transmission lines to the output ports of the hybrid, using fixed ceramic capacitors and toroid inductors. By squeezing the turns of the inductors I get a few degrees of phase adjustment without affecting the amplitude.

The complete circuit is shown in figure 1, along with design equations. 10% tolerance capacitors are fine, as the network is adjusted anyway. Table 1 shows the nearest-standard component values for the amateur bands from 160 through 10 meters.

A PSPICE analysis of the network in figure 1 on 20 meters shows a maximum phase error of 0.2 degrees from 13 MHz to 16 MHz, and an amplitude variation of plus or minus 1 dB over the same range. The network may therefore be adjusted only once, anywhere in the amateur band, and then left alone until the radio is cannibalized for parts years later.

This LO phase-shift network has the desirable qualities of being completely passive, providing 90 degree phase shift over an entire ham band, adjusting easily and operating gracefully in a 50 ohm system. In addition, the 1/8th wavelength lumped-element transmission lines are lowpass structures, which reduces the effects of LO harmonics reflecting back and forth on the transmission lines from the LO to the mixers.

Since the phase errors in a phasing transmitter are likely to be different than the phase errors in a companion phasing receiver, it is useful to be able to independently tweak the transmitter and receiver LO phase shifts. The network in figure 2 is one I used in a 20m transceiver incorporating an R2 receiver board and a T2 transmitter board. My junk box had a two-hole ferrite core for the in-phase splitter, but a pair of small ferrite toroids as specified will work just as well.

Table 1 Component Values for Amateur Bands						
fo MHz	T1 uH	# Turns	Core	L1, L2 uH	# Turns	C1-C6 pF
1.8	4.2	29	T50-2	2.1	21	820
3.7	2.2	21	T50-2	1.1	15	390
7.1	1.1	17	T37-2	0.55	12	220
10.1	0.79	14	T37-2	0.40	10	150
14.1	0.56	14	T37-6	0.28	10	100
18.1	0.44	12	T37-6	0.22	9	82
21.1	0.38	11	T37-6	0.19	8	68
24.9	0.32	10	T37-6	0.16	7	56
28.5	0.28	10	T25-6	0.14	7	56
50.1	0.16	8	T25-6	0.08	5	27

Notes:

Wind T1 with a pair of enamelled wires side-by-side. The “# Turns” for T1 is the number of times the pair is wound through the center of the core.

The same core is used for T1, L1, and L2.

Capacitors are the nearest lower standard value, to help compensate for the capacitance between the windings of T1.

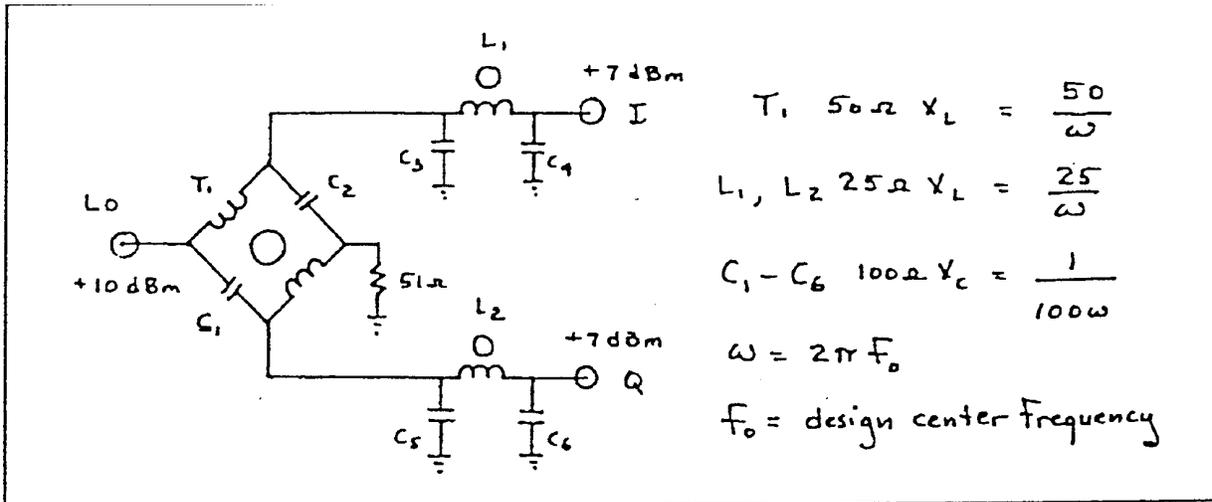


Figure 1 - A "Whole Band" Phase Shift Network

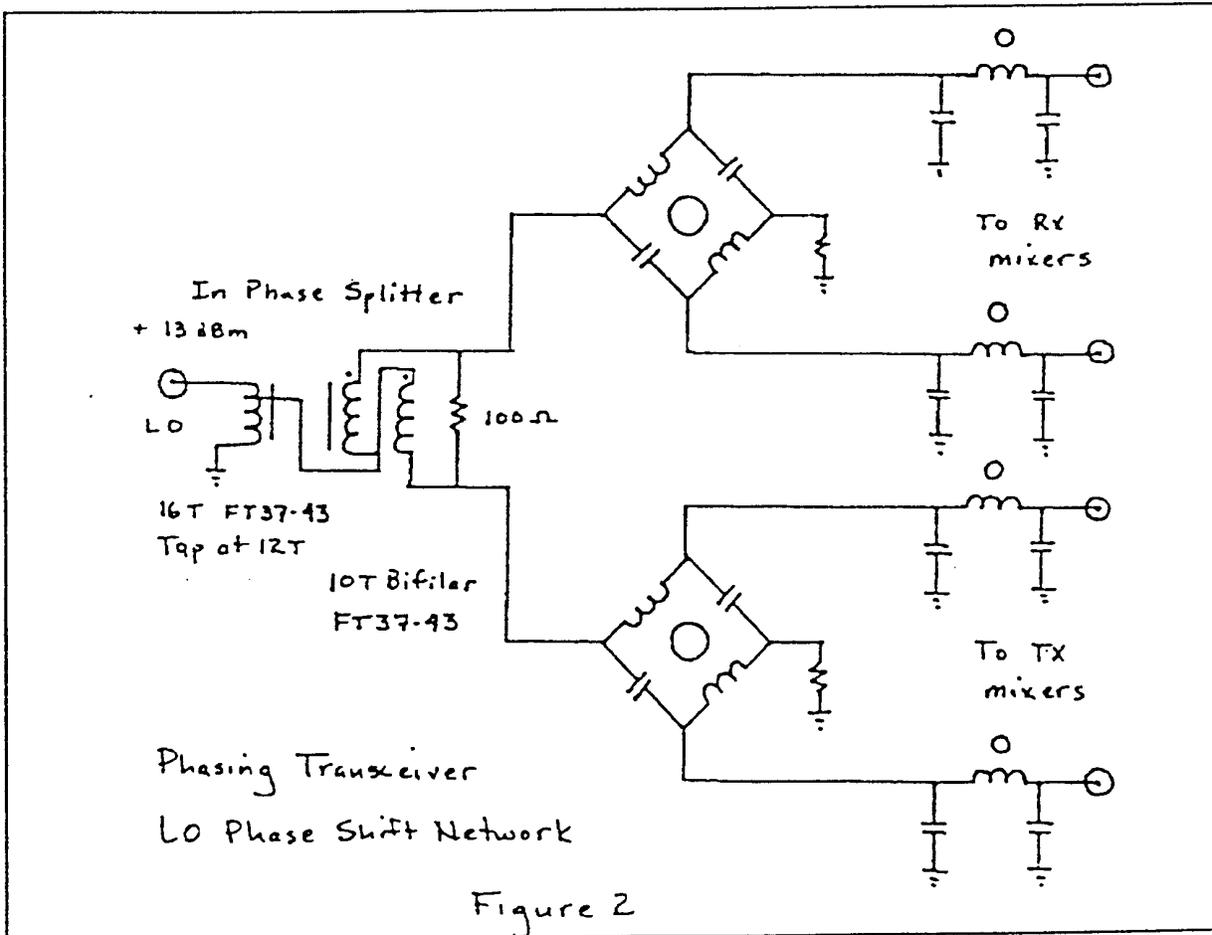


Figure 2 - Phasing Transceiver LO Phase Shift Network

## KK7B Sprat Technical Cartoon #2

### A “Minimum Component” Phasing Receiver Front End

Rick Campbell - KK7B

Published in SPRAT #82, pp 14,15, reprinted with permission

Sprat Technical Cartoon #1 (hereafter STC1) describes a nice LO phase-shift network that works well with diode ring mixers such as the ones in the R2, T2, and miniR2 boards. The QRP spirit requires that we examine our radio circuits to see if there is a way to do the same job with fewer parts. (I suspect the spirit is a Scot.)

A phasing receiver front end consists of an RF filter, and in-phase splitter driving the ports of two mixers, a quadrature LO splitter driving the LO ports of the mixers, and a means of tweaking the phase shift for maximum opposite sideband suppression. A minimum component phasing front end is shown in figure 1. The quadrature hybrid is used to drive the two ports directly. The RF ports are driven by a lumped element Wilkenson two-way splitter. Squeezing the turns of one of the toroids in the Wilkenson splitter will provide a small phase adjustment. Since the lumped element Wilkenson splitter is a low pass structure, it serves as the low pass filter needed at the input of a direct conversion receiver.

The phasing receiver front end may be used with the R2 and miniR2 circuit boards by leaving the TOKO in-phase RF splitter off the board. Connections from the circuit in figure 1 are made directly to the mixer LO and RF ports. This may be a desirable option for frequencies below 20 MHz, as the on-board TOKO splitter becomes increasingly lossy as the frequency goes down.

Included in Figure 1 are the element value equations. Table 1 shows the element values for the 160 through 6 meter amateur bands.

The QRP spirit can be kept particularly happy if the R2 or miniR2 circuit board is viewed as a “component” rather than a “receiver with a bunch of parts missing”. Think of it as a big, expensive, IC. Figure 2 is the complete circuit diagram for a minimum parts count receiver using the miniR2 as a component. The VFO circuit is from Wes and Roger Haywards “Ugly Weekender” with a U310 to get the output level up a bit. The VFO shielding is a necessary part of the circuit. One reason that commercial rigs use superhet receivers with all their images and birdies is that they can get away with an unshielded VFO. Shielding is expensive if you are building a thousand radios, but free if you are building only one.

One final comment: if the TOKO RF splitter is already soldered on the R2 or miniR2 board, it is easier to use the LO phase shift circuit in STC1 and an RF tuned circuit than to remove the 6 pin TOKO Rf splitter from those plated thru holes. The QRP spirit awards no extra points for making lots of extra work just to reduce the parts count.

Mhz	T1 uh	# turns	core	L1,2 uh	# turns	C1,2 pf	C3 pf	C4,5 pf
1.8	4.2	29	T50-2	5.9	35	820	2200	120
3.7	2.2	21	T50-2	3.0	25	390	1200	560
7.1	1.1	17	T37-2	1.6	20	220	680	330
10.1	.79	14	T37-2	1.1	17	150	470	220
14.1	.56	14	T37-6	.79	16	100	330	150
18.1	.44	12	T37-6	.62	14	82	270	120
21.1	.38	11	T37-6	.53	13	68	220	100
24.9	.32	10	T37-6	.45	12	56	180	100
28.5	.28	10	T25-6	.39	12	56	150	82
50.1	.16	8	T25-6	.22	9	27	100	47

Table 1 - Components for Amateur Bands

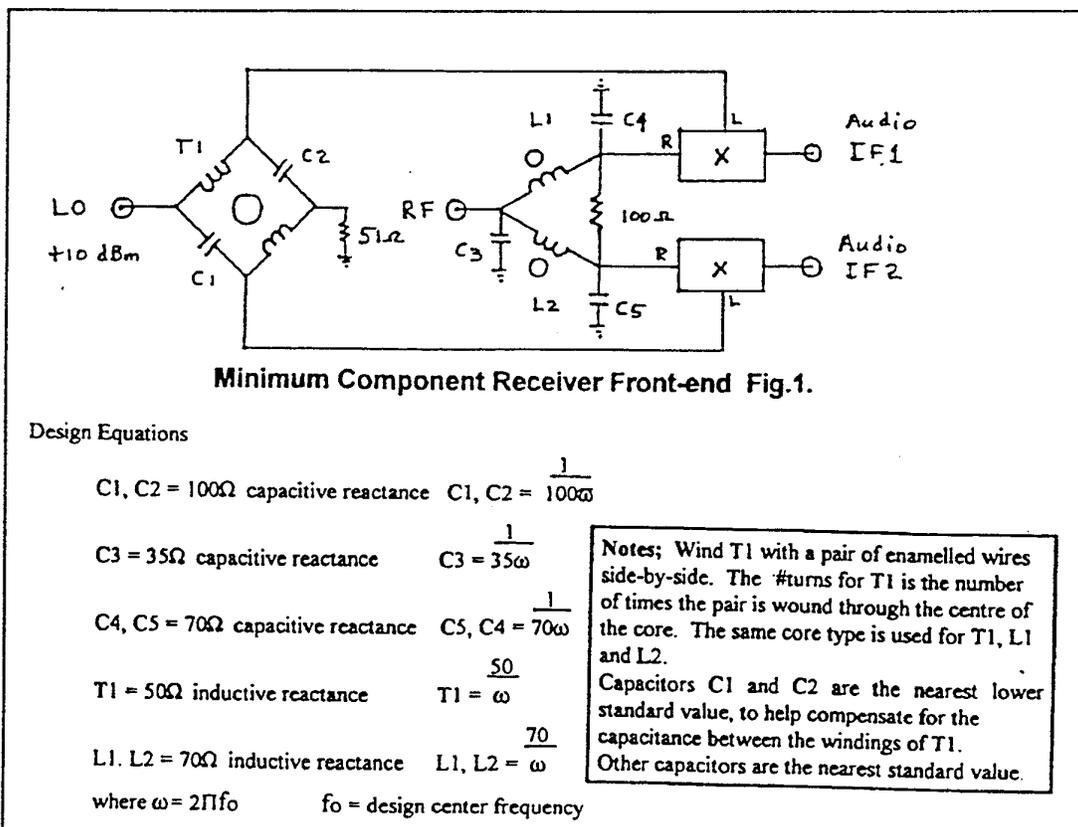


Figure 1 - Minimum Component Count RX Front-end with Design Formulas

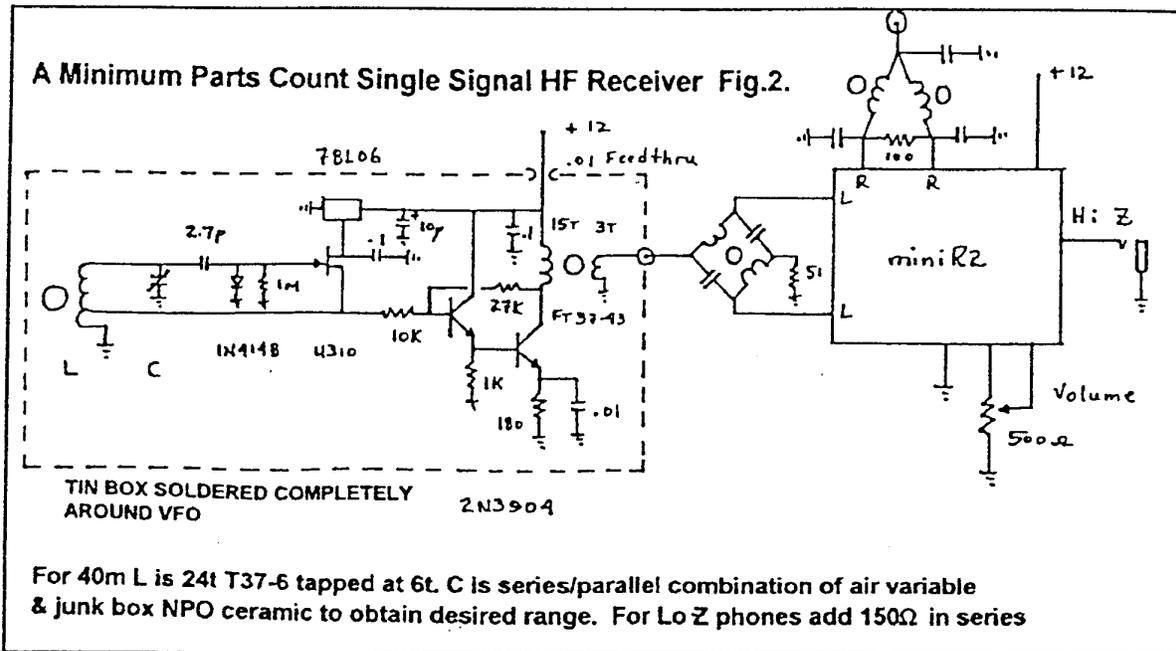


Figure 2 - A Minimum Parts Count Single Signal Receiver



