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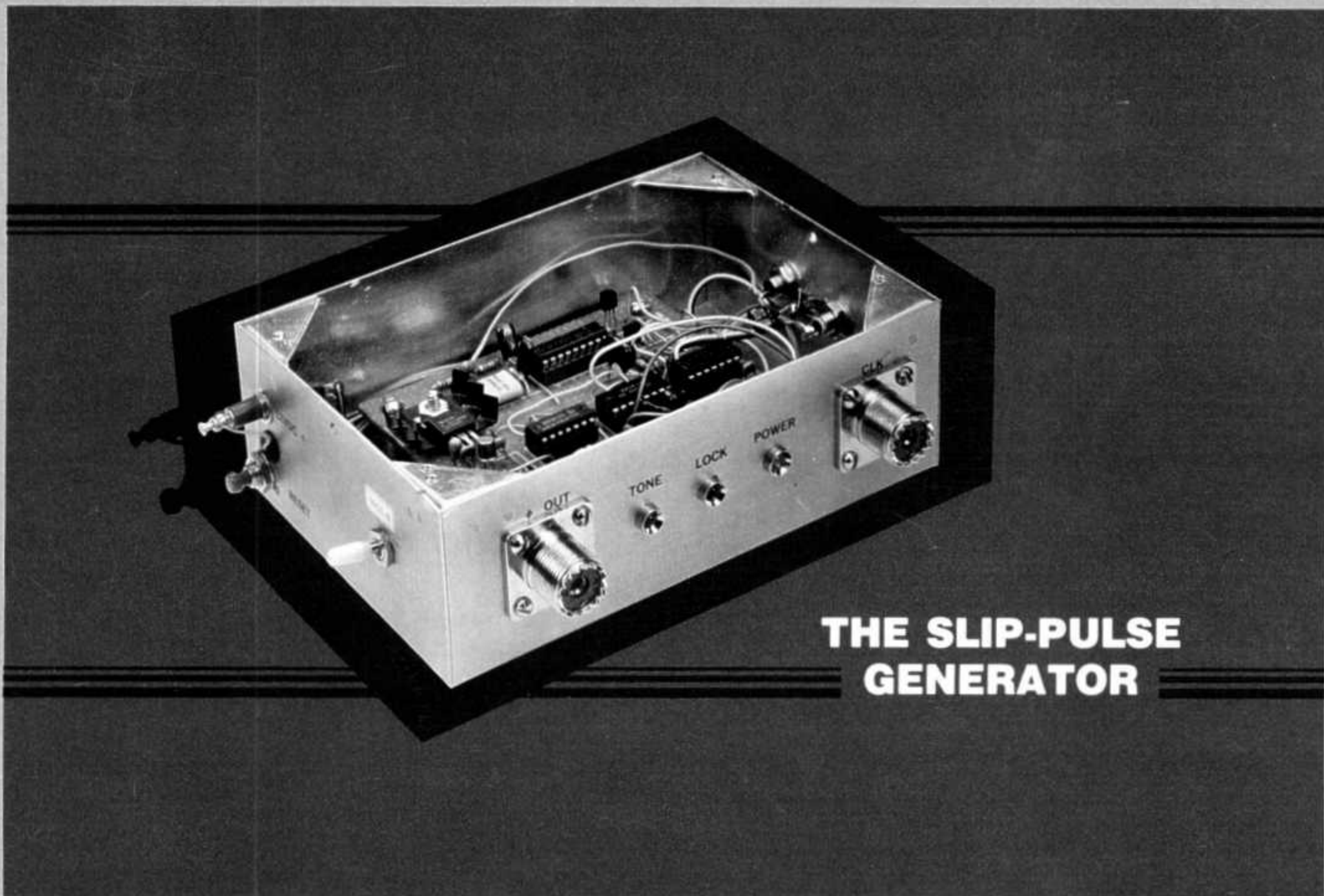

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COLUMNS

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The slip-pulse generator is an important part of a spread-spectrum communications system. The unit introduces an extra clock pulse to allow the receiver's pseudo-random noise generator to synchronize its phase to that of the transmitter.

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

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How would you like to work at your hobby (and get paid)? ARRL HQ is a good place to work and grow, with improved pay and excellent employee benefits.

Stochastic processes (and Murphy's law) have conspired to vacate three positions in the Technical Department. We've had an Assistant Technical Editor's slot open for some time, thought we had two good candidates on the line, but they slipped away. It was only February 9 that Hank Grilk, WA2CCN, came with us as Laboratory Supervisor—he's done a good job getting the lab organized but was lured back to New Jersey. More germane to *QEX* is that *QEX* Assistant Editor Maureen Thompson, KA1DYZ, will be leaving April 15 to chase happiness in Denver. Anyone who has written for *QEX* knows that Maureen has been doing the lion's share of the work it takes to put together this magazine. Also, she's been handling editor for *QST* technical articles and books. We'll miss her and wish her the very best.

So, we're looking for two journeyman-level technical editors. A technical person who can write? It's happened before, and we have some good ones here. Of these two positions, one will eventually end up editing periodicals (possibly *QEX*), the other books. It is generally desirable for technical editors to start editing periodicals because the articles are in smaller chunks, and the publication cycle is short, but that will depend on experience and projects being worked on. To qualify for the Technical Editorial Assistant position (entry-level pay $19,006) you'll need to handle the English language well, operate a word processor, have an Amateur Radio License and have some technical background. The Assistant Technical Editor job (entry-level pay $25,012) requires all that plus at least two years of formal technical education, some technical experience at work and/or as an Amateur Radio experimenter. Oh, we have a screening test to assess skills such as electronic and math theory, editing and spelling.

The Laboratory Supervisor position pay starts at $34,008. The job consists mainly of managing the lab staff of five people, who do Amateur Radio equipment testing, validation of designs offered for publication, original design work, and engineering studies. Projects of note in the lab include packet radio modems and transverters, subsystems for PACSAT in cooperation with AMSAT, transceiver remoting, integration of new radio and computer-control equipment for W1AW, and equipment for a new NASA scientific-data-collection project. We also look to the Laboratory Supervisor to work with volunteer experimenters and groups in the field on R&D at all levels. Lab engineers are encouraged to work on independent research and development (IRD) projects and to pursue continuing education.

In these days of complete disclosure, it wouldn't be nice if I didn't tell you that housing is fairly expensive in the Greater Hartford area. If you own a house in the Washington-Boston corridor, Southern California or the San Francisco Bay area, you probably have no problem moving into a house here. Locating a rental house takes a while, particularly if you're looking for one with antenna possibilities. Apartments suitable for singles or small families are available. For staffers, particularly cliff dwellers, who can't have a ham shack at home, we have a recreation station in the HQ building. We have a flexible relocation package designed to help a new employee move the family here and even help find work for the employee's spouse if needed.

So, if any of these jobs would be of interest to you, please call Technical Department Manager Chuck Hutchinson, K8CH, at 203-666-1541 during working hours.—W4RI

Correspondence

Packet Sounding Networks

I just read, with interest, W4RI's editorial in the Nov 1987 issue of *QEX* (no. 69, p 2). His ideas are well taken and, in fact, the "software" for what he proposes has been written and implemented for some time. My TNC has an "MH" command that gives a list of the last 18 or 20 stations heard with a date and time stamp. The only drawback is that it updates the time each time a station is heard, so there is no record of the number of times a particular station is heard using this method.

There are ways around this drawback, however. For instance, a dedicated "beacon frequency" could be used. (The TNC can be configured to place a date and time stamp on every packet heard.) Or, a shared frequency could be used along with "BUD" and "SUP" lists to suppress or select individual stations (most TNCs will handle up to 10 in each category).

The technology already exists off the shelf! All we must do is configure it to do what we want.—*Ronald J. Jakubowski, K2RJ, 6512 Brookland Ave, Solon, OH 44139*

Dodd/Lloyd Article Stirs Up Impedance Discussions

Several errors have crept into the program listings of the Dodd/Lloyd article, "Measurement of Antenna Impedance," (Nov 1987 *QEX*, p 6). In the TOMSMALL program of Fig 5, the variable I is not defined, so I added the following line:

205 I = E * COS(N) or I = E * M

In the first program of Fig 6, line 70 should read:

70 IF F$ = "R" THEN CHAIN "TOMFD" *not* "TOMTD"

The measurement method adjusts ER for a reading of 5 V, but B in the program is 50 (the impedance). Since all the elements are in series, the current is then 100 mA. All measured voltages must be divided by 0.1 to convert them to impedances because the vectors are *all* impedance vectors! This is not mentioned in the article.

There is no reason to force 5 V across ER. The series current is ER/R8, where R8 is nominally 50 ohms. The measured voltages can then be divided by this current to obtain the impedances (A, C, D and E) that are used in the program. Trying to set the voltage to exactly 5 is difficult and unnecessary. My modified listing is shown in Table 1.

The circuit is interesting and I plan to breadboard it to figure out what is wrong with my 40-m vertical antenna. I suspect it is more accurate than a typical R-X bridge.—*Steve Lund, WA8LLY, 10180 Mill Station Rd, Sebastopol, CA 95472*

And Another Program Change

In Fig 5 of the Dodd/Lloyd article, lines 180 and 270 should read:

180 I = E * M
270 IF X> = Ø then goto 350

With these program changes, several values in Fig 7 also change. K is now 0.857698413 and L is 0.540019972.

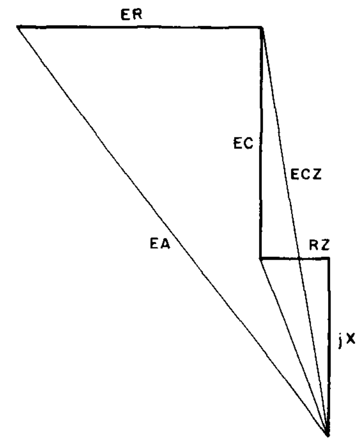
The caption in Fig 8B should read, "increasing C to 76 gives R = 76.4 + *j*48.—*Peter Dodd, G3LDO, 37 The Ridings, East Preston, West Sussex, BN16 2TW England, United Kingdom*

A Simpler Method

There is a much simpler, and more rigorous, solution to the antenna impedance problem featured in the Dodd/Lloyd article (see Fig 1 for the derivation). It is:

$$RZ = \frac{EA^2 - ECZ^2 - ER^2}{2 \times ER}$$

$$jX = \frac{ECZ^2 - EZ^2 - EC^2}{-2 \times EC}$$



1) $EA^2 = (ER + RZ)^2 + (EC + jX)^2$

2) $ECZ^2 = (EC + jX)^2 + RZ^2$

3) $EZ^2 = RZ^2 + jX^2$

4) Subtract Eq 2 from Eq 1
$EA^2 - ECZ^2 = ER^2 + 2 \times ER \times RZ$

5) Subtract Eq 3 from Eq 2
$ECZ^2 - EZ^2 = EC^2 + 2 \times jX \times EC$

From 4) $RZ = \dfrac{EA^2 - ECZ^2 - ER^2}{2 \times ER}$

From 5) $jX = \dfrac{ECZ^2 - EZ^2 - EC^2}{-2 \times EC}$

Fig 1—Derivation of the closed-form formulas used for calculating SWR values. EA^2 should always be greater than $ECZ^2 + ER^2$, or your measurements may be wrong.

Table 1
Modified Program Listing

```
1Ø CLS
2Ø Y = Ø
25 ZØ = 5Ø        ! NOMINAL VALUE OF R8
27 INPUT "B",B
29 CU = B/5Ø      ! SERIES CURRENT
3Ø B = B/CU
4Ø INPUT "A",A ! A, B, C, D, AND E ARE
45 A = A/CU       ! MEASURED WHEN ENTERED
5Ø INPUT "C",C ! AND THEN CONVERTED TO
55 C = C/CU       ! IMPEDANCES
6Ø INPUT "D",D
65 D = D/CU
7Ø INPUT "E",E
75 E = E/CU
```

By definition, this "closed form" always converges as long as EA^2 is greater than $ECZ^2 + ER^2$. If this isn't the case, something may be wrong with the measurements.

There is also no requirement to set ER to a specific value. Dodd and Lloyd seem to have misunderstood why ER was set to 50 units. In the 1965 article, ER was normalized so that a 50-Ω Smith Chart could be used to estimate SWR. Since SWR can be calculated from the data, setting ER to a normalized value is only useful if you wish to use a Smith Chart without additional arithmetic.

As for the method being "simple and the results. . .accurate," accuracy has little to do with the method (as long as it is technically correct) and everything to do with the instrumentation used. A better term for this method might be first order rigorous.

Table 2 shows the results obtained using the closed form, as well as the results from the article. Table 3 is the BASIC program I used to calculate my data. You can also derive SWR values from the data; this is shown in the last column of the table. Not knowing what the feed-line impedance was for the tests, I have calculated the SWR for both 52- and 75-Ω lines. It would be interesting to compare these values with those measured during the tests.—*D. S. Jenkins, WA6OGH, 5045 Donna Ave, Tarzana, CA 91356*

Table 2

Assumed Feed-Line Impedance For 75-Ω Line

Freq (MHz)	*EA*	*ECZ*	*EC*	*EZ*	*RES*	*ANTO*	*jX*	*ANTJ*	*SWR*
3.55	120	105	52	53	20.10	8.75	−39.10	−53.00	12.89
3.56	115	98	52	47	23.30	11.21	−34.20	−45.11	9.15
3.57	109	90	52	40	22.40	12.81	−28.60	−36.50	7.27
3.58	102	80	51	31	18.10	15.04	−24.30	−27.82	5.70
3.59	95	70	51	24	17.80	16.25	−15.70	−16.89	4.86
3.60	89	57	51	22	21.90	21.72	−1.60	−1.61	3.45
3.61	87	45	52	34	29.70	30.44	17.80	17.64	2.63
3.62	94	53	51	61	43.10	35.27	29.50	34.44	2.67
3.63	118	70	50	102	64.30	65.24	78.20	80.04	2.99
14.00	183	161	47	117	68.80	50.68	−86.20	−106.63	4.94
14.05	169	145	47	102	63.40	50.36	−75.00	−89.49	4.03
14.10	137	113	47	71	47.10	35.00	−48.20	−58.71	3.65
14.13	129	99	47	63	50.80	43.40	−34.10	−38.54	2.33
14.15	128	90	48	60	55.50	57.84	−23.80	−22.88	1.54
14.17	135	90	48	79	77.90	76.25	4.40	4.64	1.07
14.20	188	141	48	149	136.10	129.63	69.40	48.17	2.06
14.00	119	101	70	34	19.40	14.60	−26.80	−29.61	5.96
14.04	114	93	71	28	20.00	18.47	−19.20	−19.89	4.36
14.10	109	82	70	28	26.90	26.57	−7.40	−7.43	2.85
14.13	110	77	71	38	37.40	36.71	4.00	3.92	2.05
14.15	113	75	71	50	47.50	46.44	13.40	13.49	1.70
14.20	159	110	71	106	103.80	106.81	33.50	29.42	1.62

Assumed Feed-Line Impedance For 50-Ω Line

Freq (MHz)	*EA*	*ECZ*	*EC*	*EZ*	*RES*	*ANTO*	*jX*	*ANTJ*	*SWR*
3.55	120	105	52	53	20.10	8.75	−39.10	−53.00	12.20
3.56	115	98	52	47	23.30	11.21	−34.20	−45.11	8.22
3.57	109	90	52	40	22.40	12.81	−28.60	−36.50	6.14
3.58	102	80	51	31	18.10	15.04	−24.30	−27.82	4.52
3.59	95	70	51	24	17.80	16.25	−15.70	−16.89	3.57
3.60	89	57	51	22	21.90	21.72	−1.60	−1.61	2.40
3.61	87	45	52	34	29.70	30.44	17.80	17.64	1.99
3.62	94	53	51	61	43.10	35.27	29.50	34.44	2.38
3.63	118	70	50	102	64.30	65.24	78.20	80.04	3.67
14.00	183	161	47	117	68.80	50.68	−86.20	−106.63	6.15
14.05	169	145	47	102	63.40	50.36	−75.00	−89.49	4.85
14.10	137	113	47	71	47.10	35.00	−48.20	−58.71	3.79
14.13	129	99	47	63	50.80	43.40	−34.10	−38.54	2.25
14.15	128	90	48	60	55.50	57.84	−23.80	−22.88	1.53
14.17	135	90	48	79	77.90	76.25	4.40	4.64	1.48
14.20	188	141	48	149	136.10	129.63	69.40	48.17	2.89
14.00	119	101	70	34	19.40	14.60	−26.80	−29.61	4.79
14.04	114	93	71	28	20.00	18.47	−19.20	−19.89	3.28
14.10	109	82	70	28	26.90	26.57	−7.40	−7.43	2.01
14.13	110	77	71	38	37.40	36.71	4.00	3.92	1.43
14.15	113	75	71	50	47.50	46.44	13.40	13.49	1.34
14.20	159	110	71	106	103.80	106.81	33.50	29.42	2.25

Table 3
BASIC Program For Calculating Table 1 Values

```
PI = 3.1416
C2 = Ø.ØØ1
R8 = 51
ER = 5Ø
ZØ = 52
LPRINT "ASSUMED FEED-LINE
  IMPEDANCE ";ZØ;" OHMS"
PRINT
FOR X = 1 TO 22
READ F,EA,ECZ,EC,EZ,RES,JX
RSUM = ECZ^2 + RREF^2
IF RSUM > EA^2 THEN SKIP
ANTO = (EA^2 - ECZ^2 - ER^2)/(2 * ER)
ANTJ = (ECZ^2 - EZ^2 - EC^2)/(-2 * EC)
REF1 = (ANTO - ZØ)^2 + ANTJ^2
REF2 = (ANTO + ZØ)^2 + ANTJ^2
REFØ = SQR(REF1/REF2)
SWR = ABS((REFØ + 1)/(REFØ - 1)
LPRINT USING "###.## "; F,EA,ECZ,EC,EZ,
  RES,ANTO,JX,ANTJ,SWR
SKIP: NEXT X
DATA 3.55,12Ø,1Ø5,52,53,2Ø.1,-39.1
DATA 3.56,115,98,52,47,23.3,-34.2
DATA 3.57,1Ø9,9Ø,52,4Ø,22.4,-28.6
DATA 3.58,1Ø2,8Ø,51,31,18.1,-24.3
DATA 3.59,95,7Ø,51,24,17.8,-15.7
DATA 3.6,89,57,51,22,21.9,-1.6
DATA 3.61,87,45,52,34,29.7,17.8
DATA 3.62,94,53,51,61,43.1,29.5
DATA 3.63,118,7Ø,5Ø,1Ø2,64.3,78.2
DATA 14,183,161,47,117,68.8,-86.2
DATA 14.Ø5,169,145,47,1Ø2,63.4,-75
DATA 14.1,137,113,47,71,47.1,-48.2
DATA 14.13,129,99,47,63,5Ø.8,-34.1
DATA 14.15,128,9Ø,48,6Ø,55.5,-23.8
DATA 14.17,135,9Ø,48,79,77.9,4.4
DATA 14.2,188,141,48,149,136.1,69.4
DATA 14,119,1Ø1,7Ø,34,19.4,-26.8
DATA 14.Ø4,114,93,71,28,2Ø,-19.2
DATA 14.1,1Ø9,82,7Ø,28,26.9,-7.4
DATA 14.13,11Ø,77,71,38,37.4,4
DATA 14.15,113,75,71,5Ø,47.5,13.4
DATA 14.2,159,11Ø,71,1Ø6,1Ø3.8,335
```

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- Worst-Case Circuit Analysis: Philadelphia, Jun 20-22
- Printed Circuit Board and Wiring Design for EMI and ESD Control: Boston, Jun 13-14

For registration information contact R & B Enterprises, 20 Clipper Rd, W Conshohocken, PA 19428, tel 215-825-1966.—*Maureen Thompson, KA1DYZ*

Dynascan Corp Buys Lloyd's Electronics

A Chicago-based consumer electronics corporation, Dynascan Corp, has agreed in principle to purchase the assets of Lloyd's Electronics in Edison, NJ. Lloyd's Electronics distributes a broad range of consumer electronics products in the US, South America and in Asia under the brand name of Lloyd's. Dynascan Corp is known for its Cobra brand name corded and cordless telephones, answering machines, CB radios, radar detectors, scanners and facsimile machines. Dynascan's Marantz subsidiary markets quality high fidelity audio and video systems and components.—*Maureen Thompson, KA1DYZ*

Practical Spread Spectrum: Achieving Synchronization with the Slip-Pulse Generator

By Andre Kesteloot, N4ICK
ARRL Technical Advisor
6800 Fleetwood Road
McLean, VA 22101

A two-way, spread-spectrum contact was successfully conducted in January 1988 as part of a continuing series of experiments in spread-spectrum communications administered by the Amateur Radio Research and Development Corporation (AMRAD). This article describes the experiment and the equipment used. Before detailing the experiment, a brief introduction to direct-sequence spread-spectrum (DSSS) communications is in order.

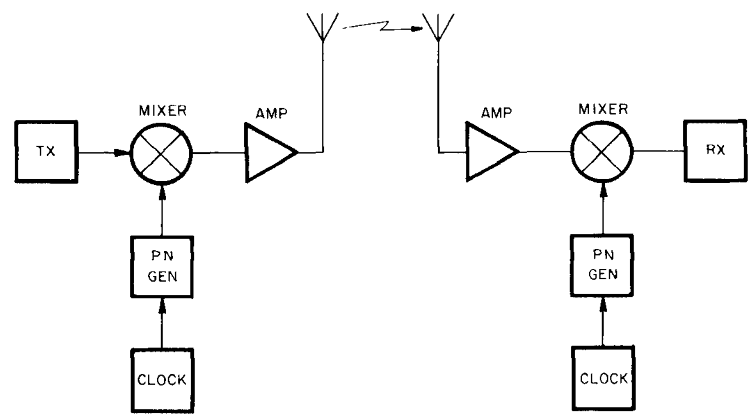


Fig 1—Block diagram of a direct-sequence spread-spectrum communications system.

Direct-Sequence Spread-Spectrum Communications

The RF bandwidth of conventional Amateur Radio modulations (AM, NBFM, FSK) is usually proportional to the amount of information that is transmitted. The bandwidth of the RF signal is kept as narrow as possible (you generally think of narrow contiguous channels), possibly a few kilohertz wide, with each channel containing separate information. When a conventional channel is in use, it is impossible for other users to transmit on the same channel without creating interference.

Using DSSS techniques, you can have a wide RF bandwidth—possibly several megahertz wide—and different networks can transmit on the same frequencies without interfering with one another. Also, the bandwidth of the signal is not related to the information rate.

At the DSSS transmitting site, a carrier is mixed with the output of a pseudo-random noise (PN) generator to spread the information over a wide bandwidth. The same pseudo-random sequence is reintroduced in the receiver (see Fig 1). If the original carrier is frequency modulated (to transmit voice, for instance), the original modulation information can be extracted at the receiver. If data other than voice is sent, this data can be logically XORed with the PN sequence at the transmitter, and similarly recovered at the receiver.[1] In either case, only the signals corresponding to the original PN sequence are correlated at the receiving site. The noise is spread over a wide bandwidth and filtered out. The signal-to-noise improvement thus derived, called *process gain*, is an important derivative of the DSSS process.

The Doubly Balanced Mixer

Fig 2 shows the schematic of a doubly balanced mixer used to mix the carrier signal and the output of the PN generator. Note that D1 through D4 are connected as a ring-modulator. The RF carrier is fed to the input transformer, while the PN signal is fed to the center tap of the output transformer. The PN generator uses TTL ICs, and by connecting it via a capacitor, the center tap of the output transformer swings from +2 V to -2 V, with respect to the center tap of the input transformer. When +2 V is applied, D1 and D3 conduct, and D2 and D4 are reverse-biased, connecting point A to C, and point B to D. When the output of the TTL goes low, the center tap of the output transformer goes to -2 V. At this instant, D2 and D4 conduct, connecting point A to D and point B to C. This produces a succession of 180-degree phase reversals of the RF carrier at the output, effectively canceling the carrier. The PN signal, which is fed in opposing phase to the input and output windings, also cancels. On the other hand, sidebands are created because of the heterodyne process. Since the output of our PN generator is a square wave in the time domain, its Fourier transform in the frequency domain is a (sin x/x) waveform. Because we are dealing with a signal proportional to the output *power* of the mixer stage, what we see on a spectrum analyzer connected to the output is proportional to the *square* of the voltage, or $(\sin x/x)^2$ (see Fig 3). As shown in the waveform of Fig 3, the original 445-MHz signal is spread over almost 10 MHz. In practice, a suitable filter shapes the transmitter output to allow radiation of only the main lobe. (The main lobe contains approximately 95% of the total power.)

At the receiving site, a similar doubly balanced mixer arrangement recreates, or "despreads," the original carrier (Fig 4). To maximize isolation between ports, each port must be properly terminated in 50-ohm loads. The setup used at both the transmitting and receiving ends is shown in Fig 5.

Synchronization

To despread a signal, the PN sequence

[1]Notes appear on page 11.

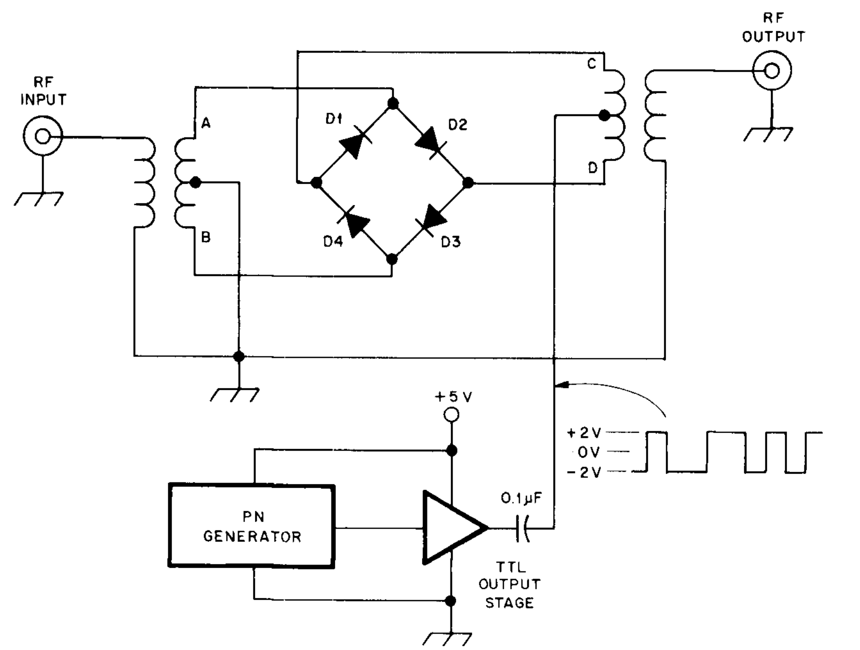


Fig 2—The doubly balanced mixer. Its function is to combine the carrier signal with the output of the PN generator.

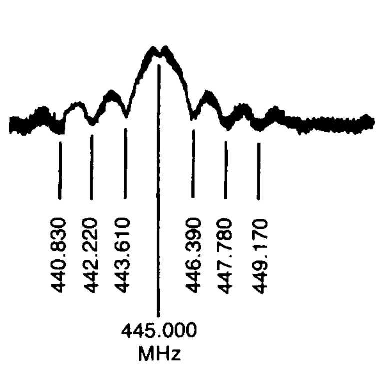


Fig 3—Output of the doubly balanced mixer at the transmitting site. The signal is spread over several megahertz, where $F_{carrier}$ = 445 MHz and F_{clock} = 1.390 MHz.

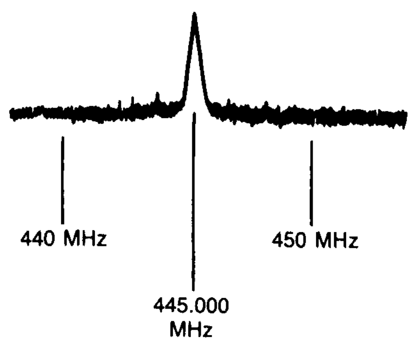


Fig 4—Output of the doubly balanced mixer at the receiving site. The signal is "despread."

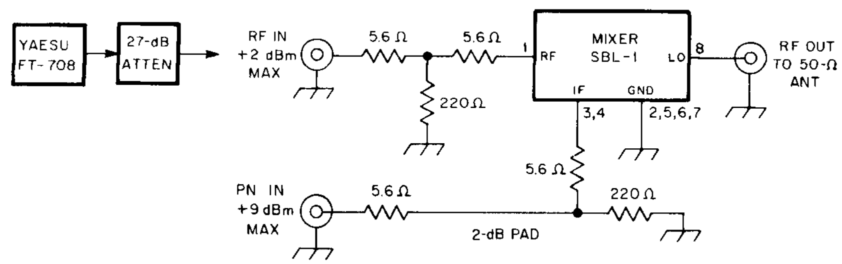


Fig 5—Schematic of the doubly balanced mixer port termination at the transmitting site.

at the receiver must be synchronized in both frequency and phase with that of the transmitter. Synchronization is considered to be the most difficult problem to solve in spread-spectrum applications.[2]

There are several ways to achieve synchronization: (1) recover the transmitter clock at the receiving site (a considerable undertaking at 445 MHz), (2) transmit the clock separately, (3) transmit the clock as part of the signal, or (4) synchronize both transmitter and receiver to an external reference, an approach suggested by William Sabin, WØIYH.[3]

The latter technique is used in this experiment, and the equipment required to extract a stable clock pulse from an AM radio station is described in an earlier article.[4] Thus, the PN generator for both the transmitter and receiver are clocked at the same frequency. The remaining problem is for the two PN sequences to operate in phase.

The station setup used for the successful spread-spectrum QSO is shown in Fig 6. Although the actual QSO was conducted at 445 MHz, some of the preliminary work was performed at 146 MHz. At 146 MHz, the output of the transmitting equipment was connected via a cable to the receiving gear. At 445 MHz, actual antennas were used at both transmitting and receiving sites. Since the equipment designed for this experiment is not frequency specific, the operation of the synchronization arrangement was exactly the same in both cases.

Referring to Fig 6, the transmitter used was a Yaesu hand-held FT-208R for 146 MHz and an FT-708R for 445 MHz. At the receiving site, I used a Yaesu FT-23R for 146 MHz and an ICOM IC-4AT for 445 MHz. The output of the transmitter and the output of the PN generator are fed to a doubly balanced mixer, as explained earlier, and the signal at the output of the mixer looks like that of

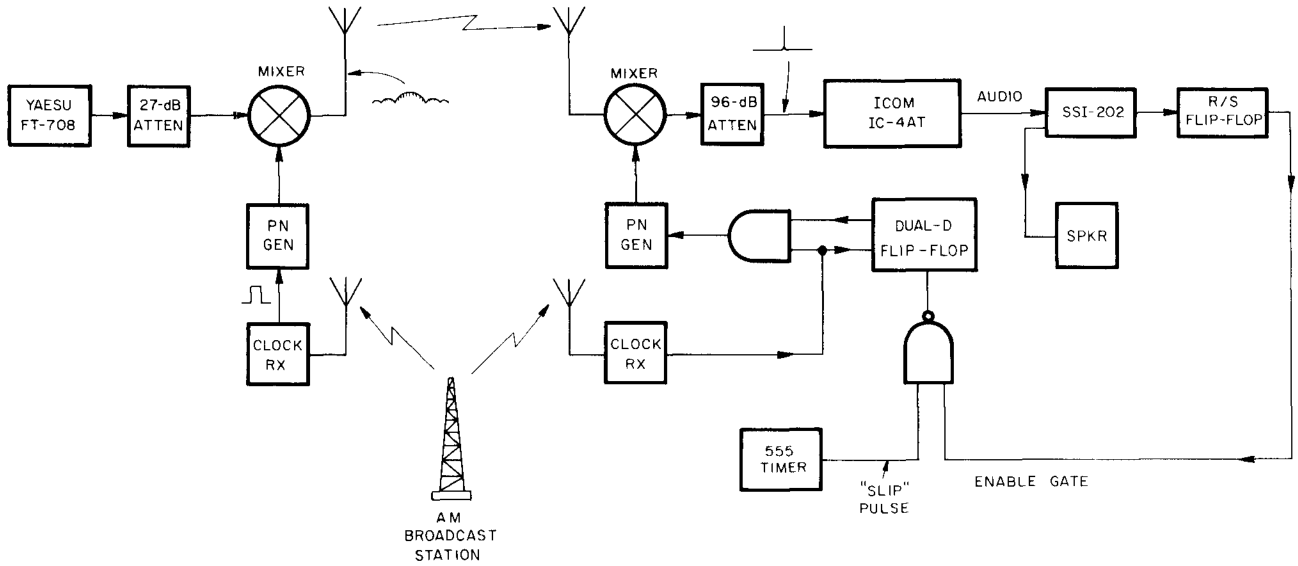


Fig 6—General equipment arrangement for the DSSS experiment.

Fig 3. To decode the original FM modulation at the receiving site, the signal must be despread. (This happens only when the two PN generators are in phase.)

To understand how the phase-synchronization process works, imagine that the two PN generators are receiving the same 1.390-MHz external clock pulse. These pulses, at both sites, are used to clock identical 7-stage PN generators (Fig 7).[5] Although the two PN generators are identical, their sequences are not necessarily in phase because they may have been started or reset at different times. To obtain phase synchronization, it is necessary to shift the phase of the receiver's PN generator, with respect to that of the transmitter's, until the two sequences coincide. To that effect, the *slip-pulse generator* occasionally introduces an extra pulse in the 1.390-MHz clock stream. Thus, the receiver's PN generator slowly "catches up" with the transmitter's until coincidence is achieved. When this happens despreading takes place, the receiver's squelch opens, and the slip-pulse generator inhibits the introduction of additional slip pulses.

Circuit Description

To achieve phase synchronization, I used the built-in DTMF (dual-tone multi-frequency) generator of the Yaesu FT-208R/708R to send a "1." I also prewired U1, an SSI-202 DTMF decoder, in the slip-pulse generator to accept a "1" as a valid output (see Fig 8).

Let's assume that the two PN sequences are not in phase. There is no despreading and thus no output from the receiver. The slip-pulse generator continues to insert pulses at a very slow rate. Since a 7-stage PN register is 127 steps long, after a maximum of 126 additional slip pulses, there will be a moment when the two PN sequences are in phase. At that moment, despreading takes place, the squelch opens, and the receiver generates a "1" that is recognized by U1. The green LED illuminates and the reset-set (R-S) flip-flop (U2A and U2B) latches, turning on the red LED and grounding pin 4 of U4B.

U3, a 555, free-runs at about 200 Hz. If U4B is enabled (ie, when there is an absence of a valid DTMF signal), U3's output pulses are channeled to U5. U5 is a 7474 connected in a one-and-only-one configuration.

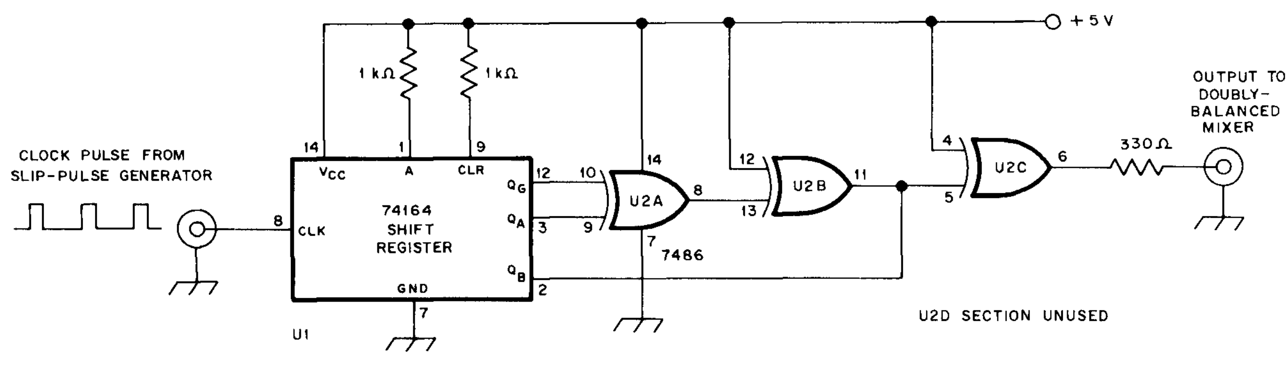
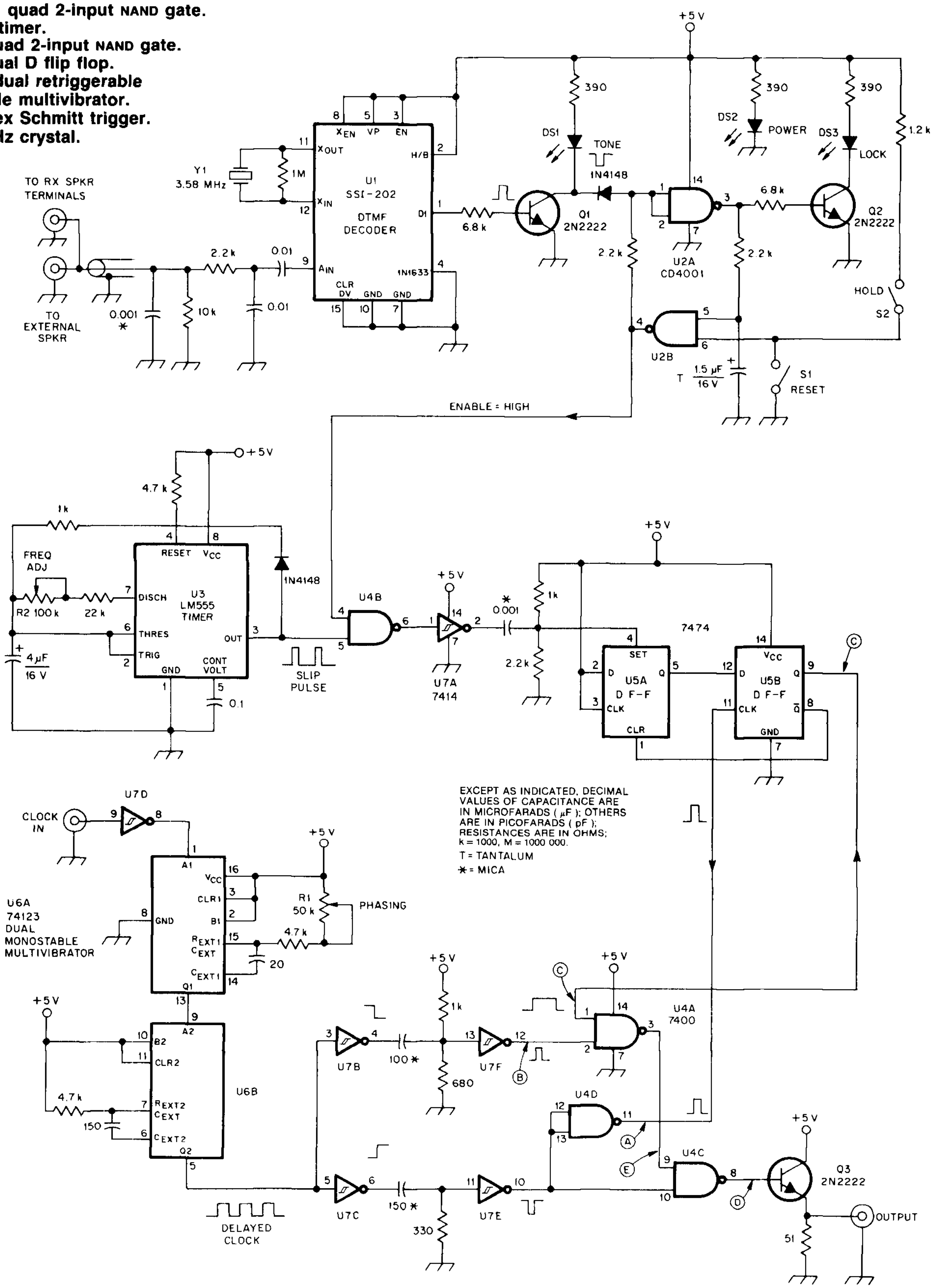


Fig 7—Schematic of the pseudo-random noise generator.
U1—74164 8-bit shift register.
U2—7486 quad 2-input XOR gate.

Fig 8—Schematic of the slip-pulse generator. Circled letters reference the waveforms shown in Fig 9.

Q1, Q2, Q3—2N2222 NPN transistor.
S1, S2—SPST switch.
U1—SSI-202 DTMF decoder.
U2—CD4011 quad 2-input NAND gate.
U3—LM555 timer.
U4—7400 quad 2-input NAND gate.
U5—7474 dual D flip flop.
U6—74123 dual retriggerable monostable multivibrator.
U7—7414 hex Schmitt trigger.
Y1—3.58-MHz crystal.



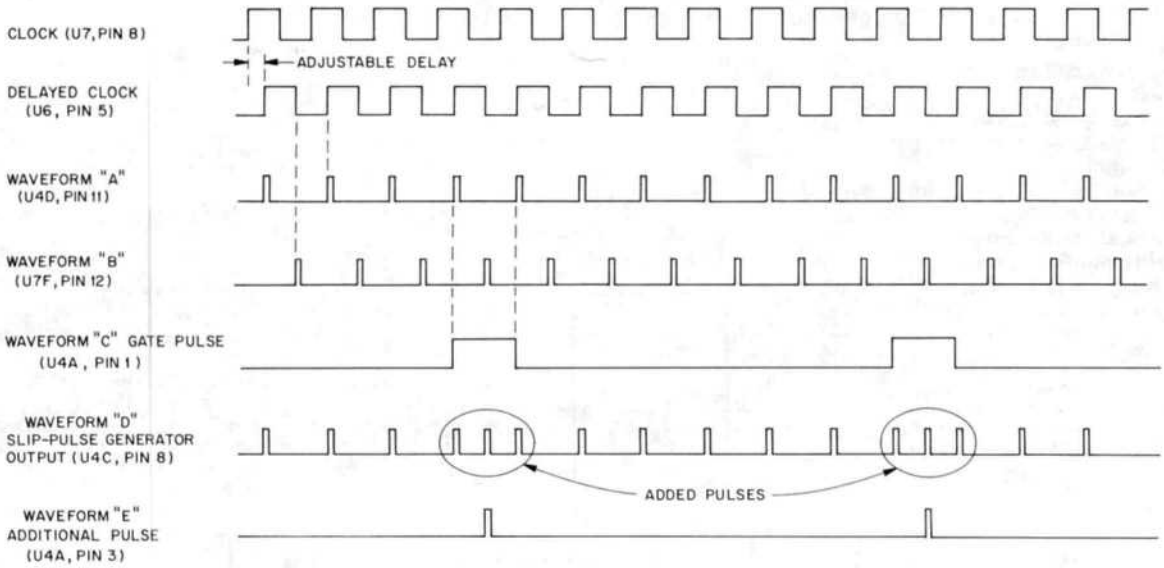


Fig 9—Slip-pulse generator timing sequence.

The external clock signal is applied to U6A and U6B, two monostable multivibrators connected in series. This circuit provides an adjustable delay to compensate for TTL gate propagation delays, different cable lengths, and so on. The delayed clock signal is then buffered by U7B and U7C, and fed to U7F and U7E (connected as two "half-monostable" sections that produce a short pulse on the trailing and leading edges of the clock pulse, respectively). The signal from U7E is also used to clock U5. U5 produces a gating signal, the length of which is the interval between two clock pulses, only after it receives a pulse from U3. Whenever this gating pulse occurs, U4A is enabled and allows a pulse synchronized on the trailing edge of the clock (ie, half-way between two regular clock pulses) to reach U4D, where it is added to the normal stream of 1.390-MHz pulses. (A pulse timing diagram is shown in Fig 9.) We have thus "slipped-in" an extra clock pulse for every pulse created by U3 (hence my name for the slip-pulse generator), and the effective clock pulse at the receiver end is 1,390,200.

This process continues as long as U4B remains enabled. A valid DTMF signal inhibits U4B. To prevent further tones or audio noise from adding extra slip pulses once phase synchronization has been achieved, the output of the DTMF decoder is connected to U2A and U2B, an R-S latch, that can only be reset by grounding pin 6 of U2B via S1. Similarly, S2 can simulate the reception of a valid tone by stopping the slip-pulse generator, a help during testing.

The Experiment

To facilitate experimentation, various parts of the equipment were built in separate shielded boxes and connected as shown in Fig 6. Fig 10 is a photograph of the slip-pulse generator. I found that at slip-pulse frequencies above 250 Hz, the system cannot achieve lock. This upper limit occurs because it takes a finite time for the receiver's squelch to open and the DTMF decoder to recognize a valid tone. By adjusting R2 so that the output frequency of U3 is about 200 Hz, it took a maximum of 42 seconds for locking to occur reliably. (Depending on the respective position of the two PN sequences, it can take less time; 42 seconds was the maximum time required when the two PN sequences were 126 steps apart.) Once synchronization is achieved, however, both transmitter and receiver clocks remain in phase, whether communication takes place or not. In a regular spread-spectrum system, synchronization must be achieved each time the push-to-talk switch is engaged. In this experiment the transmitting site and the receiving site were symmetrical (no amplifier was connected between either antenna and doubly balanced mixer), and it was possible to establish a two-way contact between the two units. Because

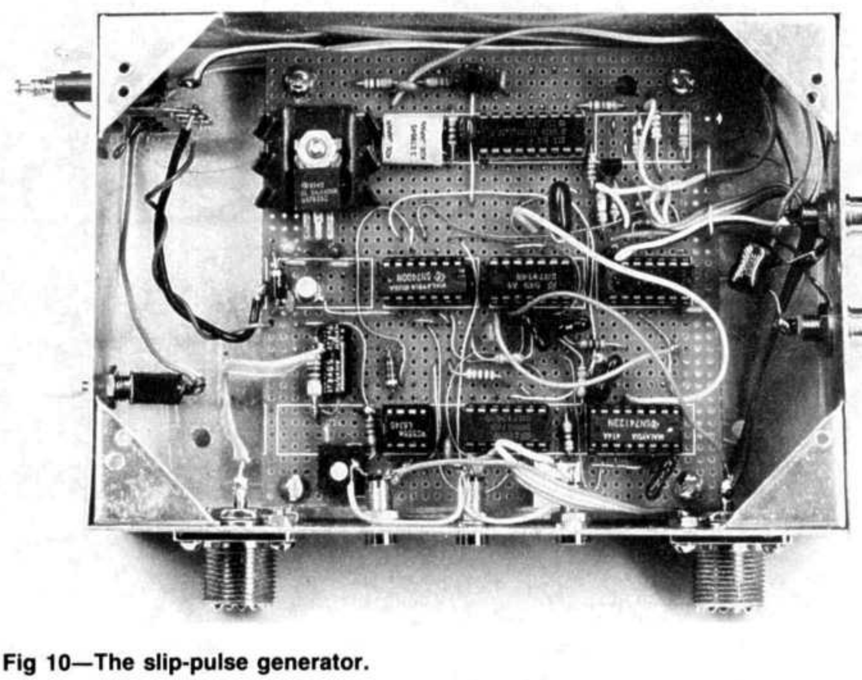


Fig 10—The slip-pulse generator.

of the extremely low power levels used (less than 1 mW), the maximum distance between the transmit and receive antennas was less than 12 inches! (The actual transmitter and receiver were located about 50 feet apart to avoid direct feedthrough from transmitter to receiver.)

Once the two PN sequences are properly phased, adjust the relative phase of the clock pulses. By properly adjusting R1 in the slip-pulse generator, it is possible to match the phase of the transmitter and receiver clock pulses. Fig 4 shows the output of the despread signal when the two clocks are perfectly in phase. Fig 11 shows the same despread signal with the receive clock about 10 degrees out-of-phase.

All parts used for this experiment are readily available. The SSI-202 touch-tone decoder chip is available from Radio Shack® (RS 276-1303), and the doubly balanced mixers were purchased from Mini-Circuit Labs (type SBL-1(6)).[6]

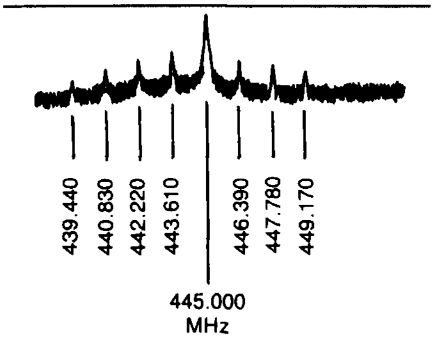


Fig 11—Output of the doubly balanced mixer at the receiving site. The two clocks are out of phase by about 10 degrees. Here, $F_{carrier}$ = 445 MHz and F_{clock} = 1.390 MHz

Conclusion

Although this experiment worked faultlessly, the equipment is fairly slow to reach synchronization and somewhat cumbersome, because it relies on the presence of a separate unit to provide synchronizing pulses. This experiment was designed to prove that it is possible to use such a synchronization approach.

My thanks go to the core group of AMRAD, particularly Chuck Phillips, N4EZV, Lawrence Kesteloot, N4NTL, and Mike O'Dell, N4NLN.

Notes

[1]A. Kesteloot, "Experimenting with Direct Sequence Spread Spectrum," *QEX*, Dec 1986, pp 5-9.

[2]R. Dixon, *Spread Spectrum Systems*, New York, Wiley 2nd ed, 1984, p 214.

[3]W. Sabin, "Spread Spectrum Applications in Amateur Radio," *QST*, Jul 1983, p 16.

[4]A. Kesteloot, "Extracting Stable Clock Signals from AM Broadcast Carriers for Amateur Spread Spectrum Applications," *QEX*, Oct 1987, pp 5-9.

[5]A. Kesteloot, "Practical Spread Spectrum: A Simple Clock Synchronization Scheme," *QEX*, Oct 1986, pp 4-7.

[6]Mini-Circuit Labs, PO Box 166, Brooklyn, NY 11235, tel 718-934-4500

Bits

Teledyne Semiconductor Markets New Components

The TSC9405 is a serial input, 16-bit parallel output shift register. It features high-current on-chip data latches and power MOS output drive devices. The part is an ideal serial interface component between microprocessor I/O ports and high-current/voltage peripheral devices, such as LEDs, thermal print heads and relays. The TSC9405 is available in a 24-pin plastic DIP and a 24-pin CerDIP package. Prices start at $4.75 each for quantities of 100 or more.

The TSC500A is an integrating converter analog processor designed for applications that require high accuracy and low noise, with the added flexibility of microprocessor control. The TSC500A is unique because all measurement functions are microprocessor controlled through two logic input signals.

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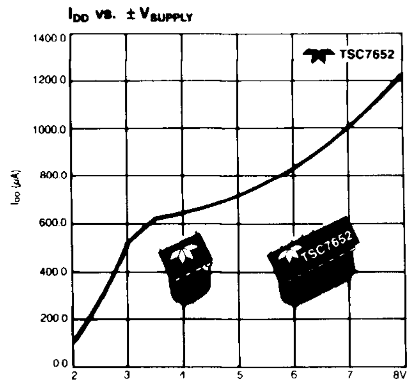

±5 V and has a power dissipation of 10 mW. This device is available in a 16-pin plastic DIP and a 16-pin CerDIP. Surface mount packages are available. Prices in quantities of 100 or more start at $6.85 each ($8.22 each for surface mount devices).

The TSC7652 is a low-noise chopper-stabilized op amp that improves noise performance and features an extremely low input-offset voltage for precision in-

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strumentation, temperature measuring and medical electronics applications. Input-offset voltage is typically ±10 μV over the device's entire temperature range. The TSC7652 is available in a 14-pin plastic package, an 8-pin plastic package, a 14-pin CerDIP package and an 8-pin CerDIP package. Prices in quantities of 100 or more start ar $4.40 each. For more information on any of these Teledyne Semiconductor devices contact the manufacturer at 1300 Terra Bella Ave, Mountain View, CA 94039-7267, tel 415-968-9241.—*Maureen Thompson, KA1DYZ*

A 759-MHz Local Oscillator

By Dave Mascaro, WA3JUF
RD 1, Box 467
Ottsville, PA 18942

The local oscillator (LO) chain is the heart of a transverter. The LO is the stage or stages that produce the signal that is mixed with the IF signal to produce the transmitted signal in an upconverter or mixed with the received signal to produce the IF signal in a downconverter. The frequency stability of a transverter depends almost totally on the LO, since the frequency stability of the IF transceiver is usually very good. (Normally a state-of-the-art HF transceiver or 2-meter transceiver is used for the IF rig.)

My design goal was a local oscillator that could be duplicated easily and could be used for many different amateur bands. The LO chain described here is the easiest to duplicate of any I have tried. Although designed for a 903-MHz transverter, the same basic design can be used for other bands.

Circuit Description

Fig 1 is a schematic of my design. Q1, the 94.875-MHz crystal oscillator, is a version of an oscillator described by Joe Reisert, W1JR, in March 1984 *ham radio magazine*. This is one of the better designs published in the amateur literature. It is relatively easy to build and tune up. Q2 is a 94-MHz amplifier with sufficient gain and power output to drive the multiplier stage.

The crystal frequency chosen, 94.875 MHz, results in an LO frequency of 759 MHz. This frequency makes the calling frequency of 903.1 MHz correspond to 144.1 MHz on the IF transceiver (± crystal oscillator and IF-rig calibration error). Changing the crystal frequency to 94.75 MHz results in an LO frequency of 758 MHz; in this case, 903.1 MHz corresponds to 145.1 MHz on the IF radio.

Very few crystal oscillators come out exactly on frequency. More than the "cut" of the crystal is involved here. The circuit design and components contribute most of the frequency error of a crystal oscillator. After multiplication to the final frequency, a small crystal-oscillator error can contribute to an LO error of many kilohertz.

Trying to "pull" the crystal usually results in an unstable oscillator, which is worse than not being able to read the frequency directly on the IF radio's display. Measure your local oscillator frequency and note the frequency error. Add or subtract the error from the IF rig display when operating. An alternative is to replace the crystal with one cut for a frequency higher or lower than the original to correct for frequency error.

Frequency Multipliers

Before going into the specifics of my multiplier design, a few general words about frequency multipliers are in order. Over the years I have tried all kinds of frequency multipliers—active and passive, tube type and solid state. I found it was harder to get the multipliers running as I added more stages of multiplication. Even with a spectrum analyzer, most solid-state multiplier-chain designs are hard to duplicate.

Active multipliers present a real challenge. Transistor beta varies from device to device, so bias characteristics vary among devices. In the case of a dc-biased multiplier stage, you must carefully adjust the bias circuit or the transistor may act like a linear amplifier instead of a frequency multiplier. This means you have to monitor and adjust the collector current of each stage for best performance as a multiplier. Many hours can be spent in front of a spectrum analyzer, and still the results can be unfavorable. Of course, when more drive is available, class-C multiplier stages are easier to use.

Passive multipliers using varactor diodes and step recovery diodes (SRDs) work very well, but their tune-up and mechanical mounting is time consuming and the results are variable. The price and availability of such diodes is also a limiting factor for amateur experimenters.

A Simple Multiplier

The x8 frequency multiplier described here uses an inexpensive 1N914 diode (D1) and a two-pole microstrip filter (C13, L5, C14, L6, C15). The crystal-oscillator frequency (94.875 MHz) is multiplied eight times directly in one stage. The two-pole filter selects the desired harmonic—759 MHz. All other harmonically related signals are down more than 25 dB. Non-harmonically related spurious signals are down more than 60 dB.

759-MHz Amplifier

The 759-MHz RF signal at the filter output is of relatively low power (-10 dBm or less) because of the low efficiency of the 1N914 multiplier diode. I tried a Schottky diode at D1, but it made little difference in power output. I suppose other diodes and matching could be used to increase the efficiency of the multiplier stage, but that would hardly be worth the effort. Any signal available at the output of the filter can be amplified to a usable level with monolithic microwave integrated circuits (MMICs). These amplifiers are truly an RF experimenter's dream-come-true.

The 759-MHz signal is coupled from L6 through a dc blocking capacitor, C16. Following the filter is a two-stage MMIC amplifier consisting of U2 and U3. Gain is 20-25 dB depending on the MMICs themselves and the values of bias resistors R8 and R9. Power output in excess of +10 dBm (10 mW) is easily obtainable. There is power output to spare, so the two-pole filter can be tuned with a little added insertion loss to further remove the unwanted harmonics. The 10-mW output can be used with a two-way power divider to supply an LO to two mixers in a transverter.

Construction

Initially I built the different stages of this project on multiple breadboards, as I do with most of my RF designs. I used separate boards for the crystal oscillator, 94-MHz amplifier, x8 multiplier and filter, and MMIC amplifiers. I scratched out the circuit patterns on G-10 PC-board material with a razor knife and a soldering iron. After each stage was checked out, I made a single PC board.

Fig 2 shows the final PC board design. The LO is built on 1/16-inch G-10 fiberglass double-sided PC board material. Double-sided PC board is used because of the microstrip circuitry and to allow for easy ground connections. The board can be etched using the full-size artwork supplied.

After drilling the component-mounting holes, you must clear away copper on the unetched side for all connections that are not grounded. This is necessary so that components do not short to the ground plane. This procedure is easily accomplished as follows: 1) Drill all non-ground mounting holes. 2) Clear these holes on the ground-plane side with an oversized drill bit. Be careful not to drill out the holes; you just want to clear away the copper. Set the drill-press depth so that the bit can only drill through half the thickness of the board. 3) Drill all the ground holes. Do this last to prevent

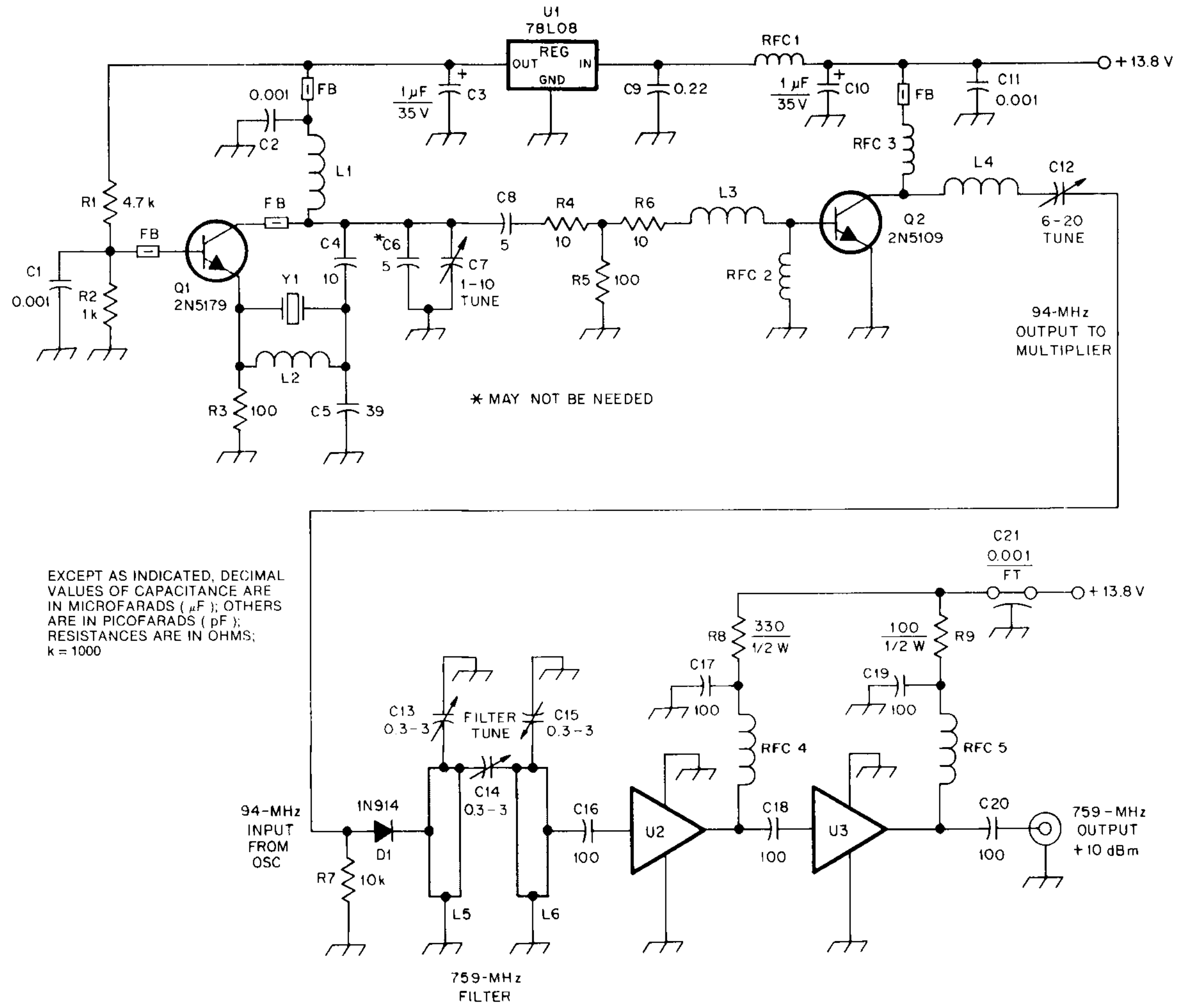

Fig 1—Schematic diagram of the 759-MHz local oscillator. Capacitors are 50-V disc-ceramic types and resistors are ¼-W types unless otherwise noted.

C3, C10—1-µF, 35-V tantalum.
C6—5-pF ceramic; may not be needed.
C7—1-10-pF ceramic trimmer.
C12—6-20-pF ceramic trimmer.
C13, C14, C15—0.3-3-pF Johanson piston trimmer.
C16-C20—100-pF chip capacitor, 100 mil square.
C21—0.001-µF feedthrough.
D1—1N914/1N4148.
FB—ferrite bead.
L1—10 ts no. 24 enam wire, 0.1-in. ID, close wound.
L2—0.39-µH molded choke.
L3, L4—10 ts no. 24 enam wire, 0.1-in. ID, close wound.
L5, L6—microstrip filter poles. See artwork.
Q1—2N5179.
Q2—2N3866 or 2N5109.
RFC1, RFC2—0.39-µH molded choke.
RFC3—10 ts no. 22 enam wire, 1/8-in. ID, close wound.
RFC4, RFC5—12 ts no. 24 enam wire, 0.1-in. ID, close wound.
U1—78L08 8-V, 100-mA 3-terminal regulator.
U2—MMIC, Avantek MSA-0104, Mini-Circuits MAR1 or equiv.
U3—MMIC, Avantek MSA-0404, Mini-Circuits MAR4 or equiv.
Y1—94.875-MHz series-resonant crystal. See text.

confusion about which holes get cleared.

After the board is drilled and cleared, drill two holes for C13 and C15. These holes are drilled on an angle to accommodate the length of the piston trimmer between the filter poles and ground on the reverse side of the board. See Fig 3. Drill the holes for the feedthrough capacitor (C21), the rivets (6 places) and the output SMA connector (if used). The rivets are pieces of wire that connect the top and bottom ground planes at critical RF grounding points on L5, L6, U2 and U3. A small piece of 50 Ω Teflon®-dielectric coax could be used between the LO and the power divider or next stage.

You can drill small divets into the PC board where the MMICs mount. Recessing the MMICs into the board slightly allows the two ground leads to be soldered directly to ground without the added inductances of the leads. Also the input and output leads will seat flush to the microstrip. This is a good practice to get into, especially at 1296 and 2304 MHz, where the few tenths of dB of extra gain add up fast.

All the components can be mounted next. Note that some components mount on the ground-plane side of the board and some mount on the etched side. D1, C14 and the 759-MHz amplifier parts all mount on the etched side of the board.

Insert the crystal last so you don't accidentally damage it. Place a piece of mica insulating material between the case (collector) of Q2 and the PC board to prevent shorting the collector supply to ground. Don't forget the hookup wire

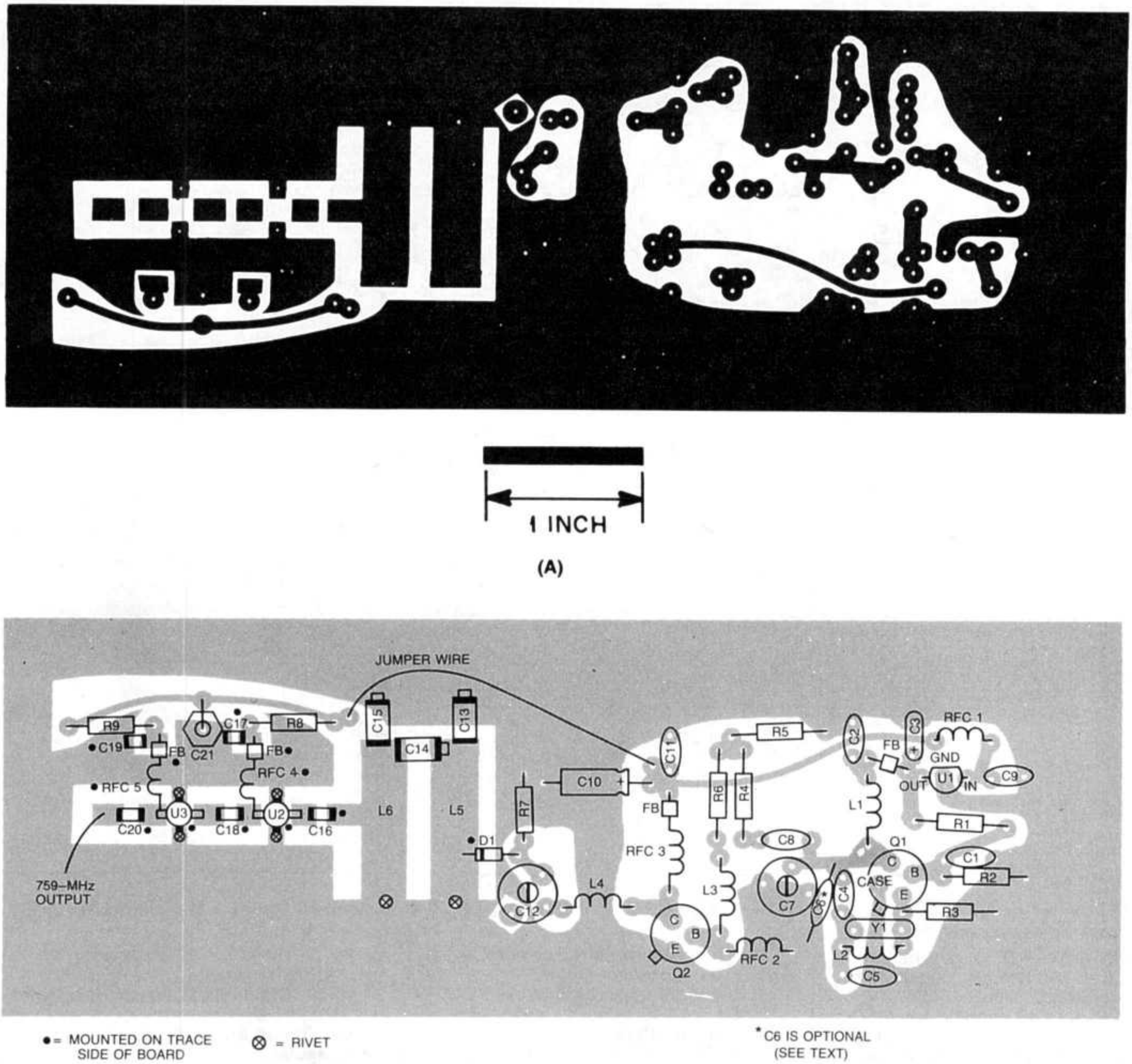

Fig 2—Circuit-board etching pattern (A) and parts-placement diagram (B) for the 759-MHz LO. The etching pattern is shown full size from the etched side of the board. Black areas represent unetched copper foil. Board material is 1/16-in. G-10, double sided. The other side of the board is unetched to form a ground plane. The shaded area in B represents an X-ray view of the copper pattern. Some components are mounted on the etched side of the board; see text.

between R8 and C10, as connection is not made on the PC board. Solder all grounds on both sides of the PC board. This also ensures that the unetched side of the PC board will be at RF ground.

Depending on your choice, many different enclosures can be used. I used the PC board itself as the top cover of the unit. Unetched PC-board scraps can be soldered to the LO board to create side walls, and a PC-board bottom cover can be added. Alternatively, all components could be mounted on the etched side of the PC board. Only U1, Q1 and Q2 require different orientation for etched-side mounting. The board could then be mounted inside a die-cast box, a milled housing or an aluminum box.

Tune-Up

After mounting all components, install the crystal. Apply 13.8-V dc (12 to 14.5 V dc usable). Check to see that regulator U1 is working properly at 8-V dc output. Connect a spectrum analyzer and power meter to the output connector and adjust C7 for crystal oscillation and maximum amplitude on the analyzer at 759 MHz. Note that you may not need C6, depending on your crystal. If the oscillator starts reliably without C6, leave it out. Tune C12 for maximum output. Adjust the tap on L5 for maximum signal at 759 MHz, while tuning C12. Then tune C13, C14 and C15 for the cleanest signal and maximum power output. C14 will be near minimum capacitance for lowest harmonic content at 10-mW output.

Go back over all adjustments as needed to get the cleanest output and the required output level for your project. At 10 to 12 mW output, all harmonics are down at least 25 dB on most of my LOs.

If a cleaner LO signal is desired (harmonics down more than 25 dB), move the D1 tap a lot closer to ground on L5. Power output at 759 MHz will be about 10-dB lower, so add another MSA-0404 MMIC after U3. Retune the entire oscillator chain. All unwanted signals will be down 35 to 40 dB, and power output will be +10 dBm.

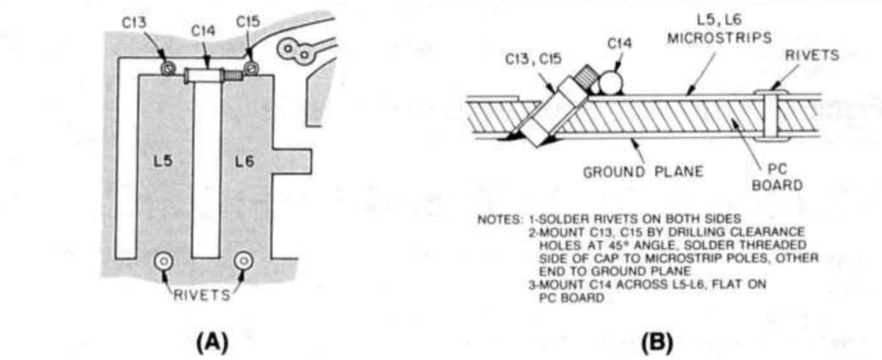

Fig 3—Construction details for the two-pole filter.

The completed 759-MHz oscillator shown from the ground-plane side (A) and the etched trace side (B).

Using the Oscillator Chain

There are other applications for this LO board. It can be used as an exciter for a 903-MHz beacon transmitter or as a crystal-controlled 903-MHz portable transmitter by installing the right crystal and shortening L5 and L6 by a few tenths of an inch. This is done by moving the rivet grounds up the line, which shortens the poles of the filter. The power level can be increased by adjusting the power supply up to 14.5 V dc, or by adding more amplifier stages. Key the MMIC dc supply with a 2N3906 PNP transistor for CW operation.

By changing the crystal frequency, this oscillator board can be used as the low-frequency portion of a microwave frequency multiplier chain. An output of 828 MHz could be used to drive an x4 frequency multiplier to 3312 MHz, the LO frequency for a 3456-MHz transverter with a 144-MHz IF.

Conclusion

This LO design has been duplicated by several hams in the local area and used in their 903-MHz transverters. Dave Hackford, N3CX, uses this design as the exciter for his 903-MHz beacon. My 3456-MHz transverter uses an LO like this to generate an 828-MHz signal, which drives an interdigital x4 multiplier to 3312 MHz. Generating stable low-level RF at frequencies up to 1 GHz is easy with this design. The use of MMIC amplifiers makes it that much easier.

Bits

RSGB News—What's Happening Across the Pond?

RSGB Holds First Data Symposium

As part of the 75th anniversary of the Radio Society of Great Britain, the RSGB will holds its first Data Symposium at the Harrow School near London on July 22-23, 1988. The program will cover a wide range of Amateur Radio data-oriented topics such as different modes, operating procedures and band usage. The RSGB is interested in hearing from all US amateurs who would like to participate in lecturing on data communications or who would like to visit and enjoy the event. Further details about the Symposium is available by contacting David Gough, G6EFQ, RSGB, Lambda House, Cranborne Rd, Potters Bar, Hertfordshire EN63JE. An AMSAT-UK Satellite Colloquium at the University of Surrey, Guildford is also planned for the same dates.—*Maureen Thompson, KA1DYZ*

Call For Papers

The University of Leicester, UK will host the Second Frequency Control and Synthesis Conference during April 10-14, 1989. This topic is important to modern electronic equipment design because frequency synthesizers are used in consumer radios, television, high-performance communications equipment, radar and precision test equipment. The Conference theme will be on advances in this field. Original technical papers and discussions are needed in areas such as devices and materials, control, measurements, frequency synthesis techniques, applications and hardware. Papers must be submitted for evaluation by the Organizing Committee on or before September 1, 1988. Complete Conference information and registration forms are available from The Conference Secretariat, Institution of Electronic and Radio Engineers, Savoy Hill House, Savoy Hill, London WC2R 0JD.—*Maureen Thompson, KA1DYZ*

International Conference on Communications

The Wyndham Franklin Plaza Hotel in Philadelphia, PA will host the 1988 IEEE International Conference on Communications during June 12-15. A technical session featuring recent developments in Amateur Radio is scheduled for Wednesday, June 15, from 12:15 PM to 1:45 PM. Attendance is limited to 50 and preregistration is necessary. The general theme for the Conference is "Digital Technology Spanning the Universe." For further information write ICC'88, c/o ATT Network Systems, 1800 John F. Kennedy Blvd, Suite 1300, Philadelphia, PA 19103. Contact Kay Craigie, KC3LM, at 215-688-5045 for information on the Amateur Radio technical session.—*Maureen Thompson, KA1DYZ*

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RF Filters for VHF and UHF: Part 2, Practical Designs

Last month, I discussed basic RF filter terminology and filter applications in Amateur equipment. This month I will show some practical filter designs. Since literally hundreds of volumes have been written on filter design and theory, this discussion is not meant to be exhaustive. Rather, I will share a few practical ideas and a place to start to find more information. The reader is urged to consult the reference material for additional details.

As mentioned last time, filters take four basic forms: low pass, high pass, band stop and band pass. In the VHF and UHF amateur world we usually use band-pass filters and an occasional notch (band-stop) filter. This is true because our operation is narrowband, and our primary filter function is to pass one narrow band of frequencies and reject all others.

Low-pass and high-pass filters pass all frequencies either below or above specified "cutoff" frequencies. They are useful in applications such as low-pass TVI filters after multiband HF transceivers, where you want to pass a whole range of frequencies below 30 MHz and reject any harmonics or spurious outputs above that frequency. At VHF and above, the situation is different. Most transmitters are dedicated to one band, rather than to a wide range of frequencies. Here, a high-Q band-pass filter is usually used in the output to attenuate harmonics and spurious emissions above and below the desired frequency.

Now that we are seeing multiband transceivers with HF *and* 6-meter capability, a low-pass filter with a cutoff frequency just above 50 MHz is probably a useful device. The design of such a filter is covered in *The ARRL Handbook*. In just about every other instance, VHFers can use band-pass filters—the main subject of this month's discussion.

Band-Pass Filters

Band-pass filters take many forms. In choosing the right one for your intended application, you must consider the operating frequency, power-handling requirements and filtering requirements. Basically, filters rely on the reactance of inductive and capacitive elements to define their performance.

The simplest band-pass filter is the parallel resonant LC network shown in Fig 1A. This circuit has a very high impedance at resonance, and the impedance gradually decreases above and below resonance. Increased selectivity is obtained by cascading two or more tuned circuits and by decreasing the loading on these networks by tapping the input and output loads down on the tuned circuit. See Fig 1B. Coupling between the two cascaded tuned circuits is usually accomplished by mutual coupling between the inductors, but they can be link coupled or capacitively coupled also. Filter tuning is accomplished by means of variable capacitors.

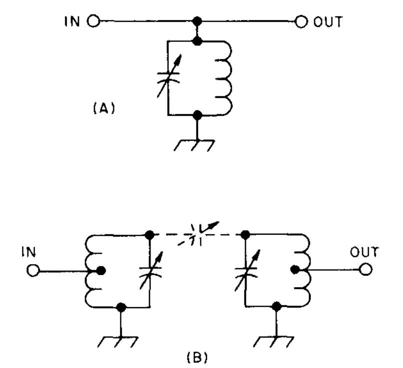

Fig 1—The most basic RF filter is the parallel LC type, shown at A. Networks can be cascaded for more selectivity, as shown at B.

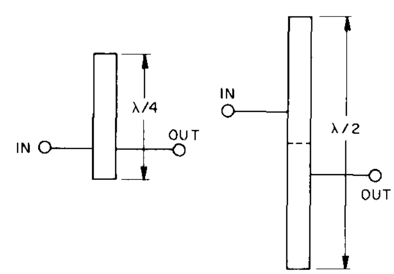

Fig 2—Transmission-line resonators—either ¼ or ½ wavelength long—can be used as filter elements. Transmission lines can be made from sections of coaxial cable, or they may be made as striplines or microstriplines. The quarter- and half-wavelength structures are equivalent, except the half-wavelength elements require no ground connection.

The transmission-line equivalent of the parallel tuned circuit is a resonant quarter-wavelength line with one end grounded (see Fig 2A), or a half-wavelength ungrounded line (see Fig 2B). The transmission-line elements in these filters can be sections of coaxial line, stripline or microstrip. The right element type for the application depends on the space available, frequency of operation, desired rejection characteristics and insertion loss. Microstrip tends to be a bit more lossy than air stripline or coaxial line, but more compact. "Lossy" here refers to insertion loss as well as to broader response. A special case of the quarter-wave coaxial line is the helical filter where the line is coiled up into a more compact form.

For increased selectivity, transmission-line elements can be cascaded. Many standard filters we see in equipment are variations on this theme. Examples include comb-line and interdigital filters, as well as parallel-coupled half-wave line and hairpin-line filters. See Fig 3.

As with the parallel LC filters of Fig 1, transmission-line filters are often made tunable by shortening the lines a bit and end loading them with a variable capacitor. By having a tunable element, the need for extremely precise construction tolerances is lessened. In addition, provision for tuning allows one filter to be used over a limited frequency range.

Tuned versus Fixed-Frequency Filters

Sometimes we need a filter for just one application at one frequency (for example, to clean up the output of a local oscillator). The filter doesn't need to be tunable, so why bother? There are a number of good reasons to build filters with no tunable elements. One reason is that the tuning capacitor will degrade the performance of the filter by introducing additional losses. Another is the expense of the extra component or components. If we were building large quantities of the same circuit, the cost factor would become significant, and the time spent optimizing a no-tune filter would be justified.

Today, computer programs can design fixed-tuned filter structures that perform almost exactly as the theory dictates—if the mechanical layout is adhered to exactly. In the home "hammer and hacksaw" shop, this is not always possi-

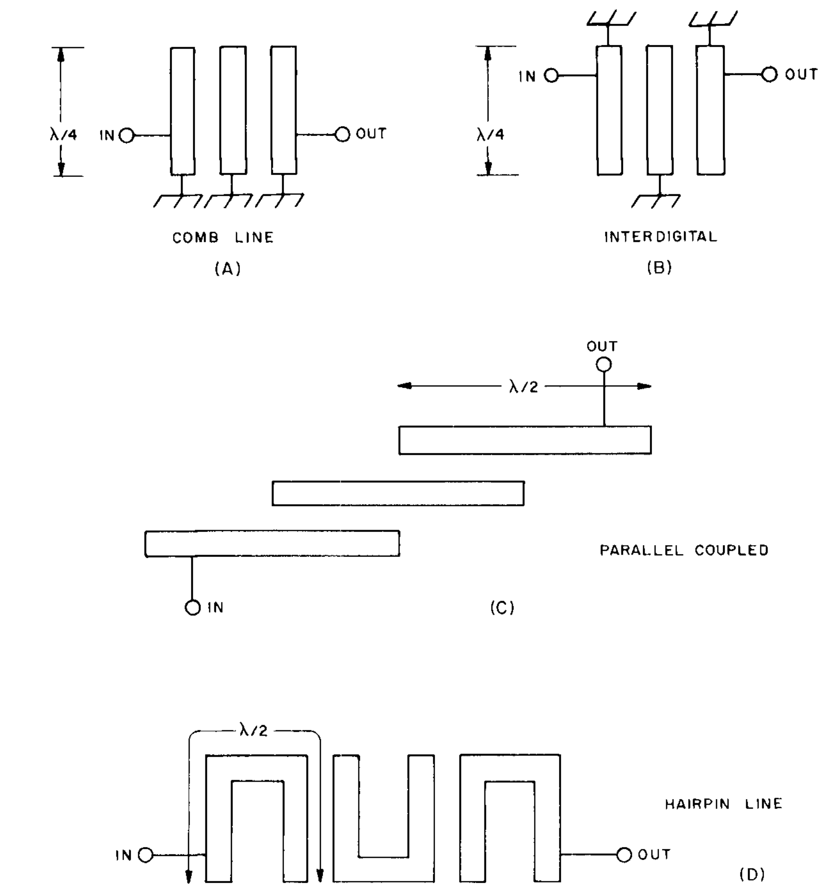

Fig 3—Cascaded transmission-line filters may take several forms. Note that the filters at B, C and D are equivalent structures.

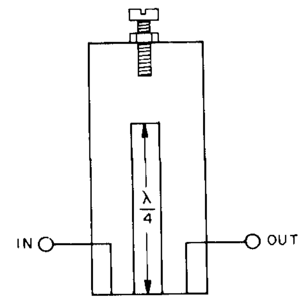

Fig 4—The single-resonator coaxial-line filter with inductive input and output coupling is a simple but effective design.

ble. A lot of us don't have access to accurate PC-board layout and etching services, so building in a tunable element is the only way to get the filter to work! Of course, now you have to have some way to tune the bloody thing up . . . trade-offs again.

Bandwidth Considerations

Sometimes, even in amateur narrow-band systems, we want a filter with a wider passband than can be obtained by just peaking every adjustment at one frequency, yet we also want high attenuation just outside the passband. One application that comes to mind is in a transverter designed to operate on 1296 MHz for terrestrial work *and* 1269 MHz for OSCAR Mode L. Filters in the LO, transmitter and receiver stages must be able to pass signals over a 27-MHz bandwidth, presumably with equal attenuation, if flat across-the-band performance is to be maintained.

We usually make wide-bandwidth band-pass filters by taking a multi-element filter and "stagger tuning" the resonators across the desired band. In-band ripple may be a problem, but with patience and practice, a multi-resonator filter can be tuned to pass a fairly wide band of frequencies.

Many computer programs will design filters with a specified in-band ripple and bandwidth. Such computer-designed filters have elements that resonate at slightly different frequencies; that's what gives them wider bandwidths. Such filters have to be built as described to work properly.

Usually, a computer-generated filter design must be built and tested to verify performance. If the passband is offset by some amount, the design is then rescaled and the process is repeated until the desired performance is obtained. Another reason to build some bandwidth into an untuned filter structure is to compensate for manufacturing variations.

Recommendations

My recommendation is to stick with tunable filters that can be built with hand tools. I prefer to use quarter-wave coaxial filters (see Fig 4) and interdigital designs above 900 MHz, and parallel LC networks below about 400 MHz. Between these frequencies, an end-tuned quarter-wave line and a parallel LC circuit start to look curiously similar.

A very useful coupled transmission-line filter design is described in Chapter 32 of *The ARRL Handbook.* It uses short lengths of 0.141-inch semirigid coaxial cable for the line elements. A portion of the center conductor inserted into the dielectric serves as the tuning capacitor. This technique has been used at frequencies from 900 MHz to 3456 MHz in the straight coupled comb-line filter configuration or in the interdigitated configuration. Table 1 shows quarter wavelengths in air for frequencies from 900 to 3456 MHz. In real-world tunable filters, these lengths are made about 10% shorter and the resonators are capacitively tuned. Single-resonator capacitively tuned coaxial filters are described in the literature, and the lengths for a quarter wavelength are the same as shown in Table 1. On microstrip, remember to take into account the shortening caused by the dielectric constant of the board material.

One note about resonant transmission-

Table 1

Lengths for Quarter-Wavelength Resonators in Air

Frequency (MHz)	*Length (Inches)*
902	3.274
1152	2.563
1269	2.327
1296	2.280
2160	1.367
2304	1.280
3312	0.892
3456	0.850

Note: Stray capacitance tends to lower the resonant frequency of filter elements. For a tunable filter, start with a line length 10-15% shorter than specified and tune to resonance.

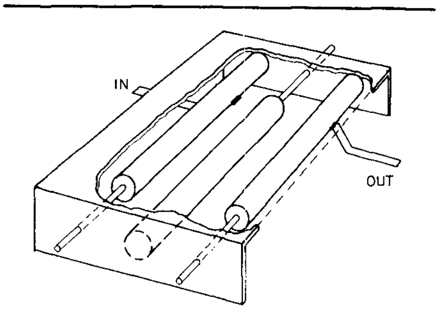

Fig 5—Another easy-to-build design is the interdigital filter using quarter-wavelength pieces of 0.141-inch semirigid cable for resonators. The filter may be tuned by varying the amount of center conductor inserted in the dielectric.

line filters: They are also resonant at other frequencies. For example, a 432-MHz quarter-wavelength filter will also pass 1296 MHz (as a ¾-wavelength filter) and 2160 MHz (as a 5/4-wavelength filter). If this is a problem, coupling into the filter is usually made frequency selective in some way.

When designing a multi-element resonant line filter (such as an interdigital filter), the actual impedance of the line has little effect on the design as long as the Q is high and the impedance is physically realizable. Most filter designs of this type use line impedances between 50 and 75 ohms. (The impedance of a round line centered between two ground planes, where the diameter of the round line is ½ the spacing between the ground planes, is around 58 ohms.) This is a good trade-off between performance and construction ease. A 1:3 ratio results in a line impedance of 80 ohms, also an acceptable value. Spacing of the lines obviously determines the coupling factor and consequently affects the insertion loss as well as selectivity. There is lots of stuff to play around with here!

Conclusions

There is nothing magic about filters if you understand the basics, and there is plenty of good design material in the amateur literature and elsewhere. Therefore, there's no reason not to have birdie-free receivers and ultra-clean transmitter outputs! Go for it.

UPCOMING EVENTS

Whoa! It's conference time again! At this writing there's still plenty of snow on the ground in Maine, but it's time for the Spring gatherings again. The 14th Annual Eastern VHF/UHF/SHF Conference is again being held at Rivier College in Nashua, New Hampshire. Noise-figure and antenna-measuring contests are offered, as well as a slate of guest speakers and informal discussions. The date to remember is May 20-22. Contact Lew Collins, W1GXT, 10 Marshall Terrace, Wayland, MA 01778 for more information.

Another popular event, the West Coast VHF Conference, will be held May 14-15 at the Buellton Holiday Inn near Santa Barbara, California. Additional details are available from Al Soenke, WA6VNN, tel 805-968-0873.

References

R. Campbell, "9 and 13 cm Transverters," *Proceedings of Microwave Update '87*, pp 12-21. (Available from ARRL.)

J. Davey, "Microstrip Bandpass Filters," *Proceedings of Microwave Update '87*, pp 42-53. (Available from ARRL.)

J. Hinshaw and S. Menemzadeh, "Computer-Aided Interdigital Bandpass Filter Design," *ham radio*, Jan 1985, pp 12-26.

The 1988 ARRL Handbook, Chapter 32, pp 32-17, 32-23 and 32-30.