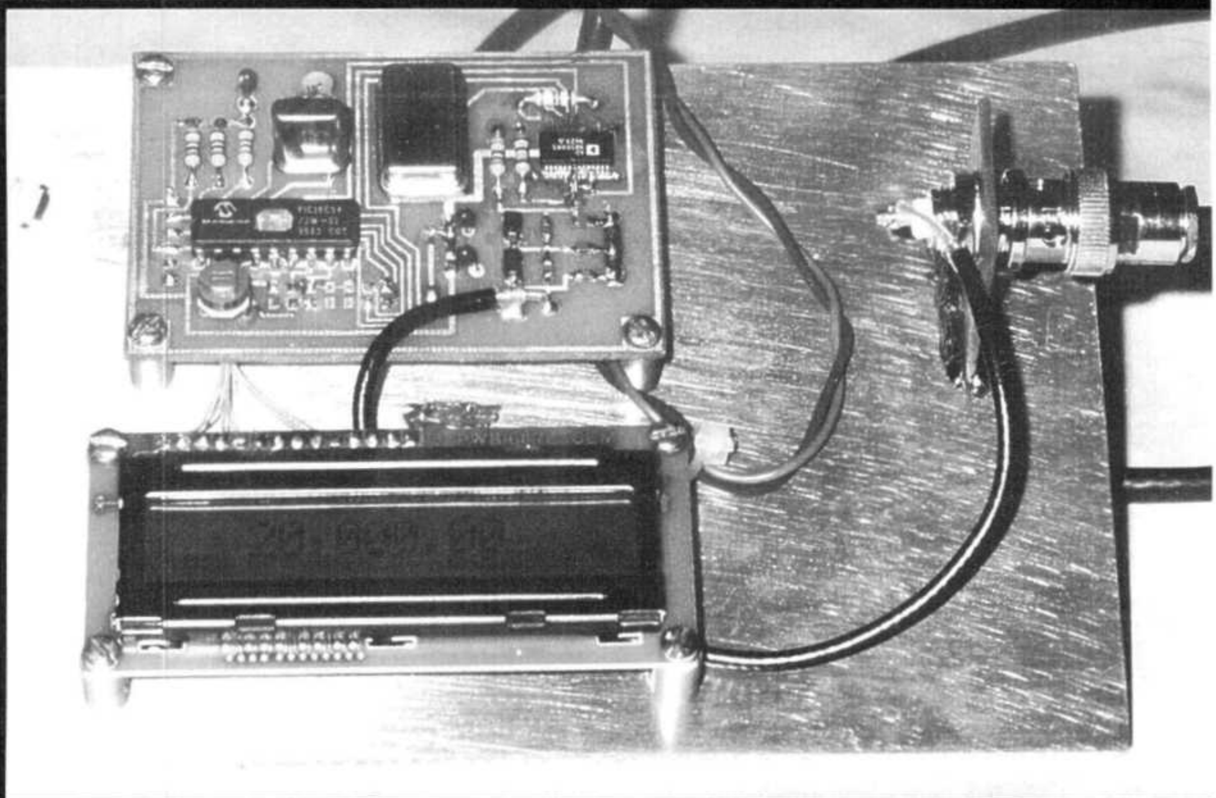
QEX

$1.75

ARRL Experimenter's Exchange

July 1997



Get Started with DDS

QEX: The ARRL
Experimenter's Exchange
American Radio Relay League
225 Main Street
Newington, CT USA 06111

QEX

QEX (ISSN: 0886-8093) is published monthly by the American Radio Relay League, 225 Main Street, Newington CT 06111-1494. Subscription rate for 12 issues to ARRL members is $15; nonmembers $27. Other rates are listed below. Periodicals postage paid at Hartford CT and at additional mailing offices.

POSTMASTER: Form 3579 requested. Send address changes to: QEX, 225 Main St, Newington CT, 06111-1494

Issue No. 180

Offices

225 Main St, Newington, CT 06111-1494 USA
Telephone: 860-594-0200
Telex: 650215-5052 MCI
Fax: 860-594-0259 (24 hour direct line)
Electronic Mail: MCIMAILID: 215-5052
Internet:qex@arrl.org

Subscription rate for 12 issues:

In the US: ARRL Member $15, nonmember $27;

US, Canada and Mexico by First Class Mail: ARRL Member $28, nonmember $40;

Elsewhere by Surface Mail (4-8 week delivery): ARRL Member $20, nonmember $32;

Elsewhere by Airmail: ARRL Member $48, nonmember $60.

Members are asked to include their membership control number or a label from their *QST* wrapper when applying.

In order to insure prompt delivery, we ask that you periodically check the address information on your mailing label. If you find any inaccuracies, please contact the Circulation Department immediately. Thank you for your assistance.

About the Cover

Want a DDS project to get started with? This 0.1- to 20-MHz VFO from WB2V may be just the thing.

Features

Columns

July 1997 QEX Advertising Index

THE AMERICAN RADIO RELAY LEAGUE

The American Radio Relay League, Inc, is a noncommercial association of radio amateurs, organized for the promotion of interests in Amateur Radio communication and experimentation, for the establishment of networks to provide communications in the event of disasters or other emergencies, for the advancement of radio art and of the public welfare, for the representation of the radio amateur in legislative matters, and for the maintenance of fraternalism and a high standard of conduct.

ARRL is an incorporated association without capital stock chartered under the laws of the state of Connecticut, and is an exempt organization under Section 501(c)(3) of the Internal Revenue Code of 1986. Its affairs are governed by a Board of Directors, whose voting members are elected every two years by the general membership. The officers are elected or appointed by the Directors. The League is noncommercial, and no one who could gain financially from the shaping of its affairs is eligible for membership on its Board.

"Of, by, and for the radio amateur, "ARRL numbers within its ranks the vast majority of active amateurs in the nation and has a proud history of achievement as the standard-bearer in amateur affairs.

A bona fide interest in Amateur Radio is the only essential qualification of membership; an Amateur Radio license is not a prerequisite, although full voting membership is granted only to licensed amateurs in the US.

Membership inquiries and general correspondence should be addressed to the administrative headquarters at 225 Main Street, Newington, CT 06111 USA.

Telephone: 860-594-0200
Telex: 650215-5052 MCI
MCIMAIL (electronic mail system) ID: 215-5052
FAX: 860-594-0259 (24-hour direct line)

Officers

President: RODNEY STAFFORD, KB6ZV
5155 Shadow Estates, San Jose, CA 95135

Executive Vice President: DAVID SUMNER, K1ZZ

Purpose of *QEX*:

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

All correspondence concerning *QEX* should be addressed to the American Radio Relay League, 225 Main Street, Newington, CT 06111 USA. Envelopes containing manuscripts and correspondence for publication in *QEX* should be marked: Editor, *QEX*.

Both theoretical and practical technical articles are welcomed. Manuscripts should be typed and doubled spaced. Please use the standard ARRL abbreviations found in recent editions of *The ARRL Handbook*. Photos should be glossy, black and white positive prints of good definition and contrast, and should be the same size or larger than the size that is to appear in *QEX*.

Any opinions expressed in *QEX* are those of the authors, not necessarily those of the editor or the League. While we attempt to ensure that all articles are technically valid, authors are expected to defend their own material. Products mentioned in the text are included for your information; no endorsement is implied. The information is believed to be correct, but readers are cautioned to verify availability of the product before sending money to the vendor.

Empirically Speaking

We're Back

Did you miss us? The last issue of *QEX* published prior to the one you hold in your hands was the December, 1996 issue. So, what happened to the January through June 1997 issues? The short answer is: there weren't any.

The longer answer is that due to a combination of circumstances, we suspended publication of *QEX* for the first six months of 1997. The reasons why *QEX* failed to appear during that period are several. Foremost among them is that staff resources to get *QEX* out the door suddenly became scarce due to staff departures and the shifting of duties of other staff to focus on income-generating projects in light of the League's 1996 financial performance. (Nobody here works full-time on *QEX*. We all have other duties as well.)

At the same time, incoming article material suitable for use in *QEX* slowed to a trickle. It seemed expedient, therefore, to defer working on *QEX* for a while. Our original idea was that we would simply be a bit late on a few issues. But as other projects consumed more-than-expected time, *QEX* kept slipping further behind. Finally, we decided the best thing to do is to simply start afresh, with this July issue, and just not publish the January through June issues.

Each *QEX* subscriber will receive all of the issues paid for; your subscription will simply be extended by six months to account for the six "missing" issues.

We regret the annoyance this has caused some subscribers. We're confident we have implemented solutions to the problems of getting *QEX* to you, and we don't expect any repetition of this suspension of publication.

This Month in *QEX*

Direct digital synthesis (DDS) is a standard way of generating signals these days—nothing new or mysterious about it. But you may still be looking for a DDS project to get started using and understanding the technology. "Building a Direct Digital Synthesis VFO," by Curtis W. Preuss, WB2V, provides such a project.

Then again, maybe you feel that the analog VFO is more your style. In that case, Mark L. Meyer, WUØL, invites you to "Meet the Vacker: The Simple, Stable VFO You've Been Looking For."

The balun: is there any component of the average amateur station quite so misunderstood? "Exploring the 1:1 Current (Choke) Balun," by William E. Sabin, WØIYH, is a thorough look at how these devices work and what they do.

Finally, everybody needs a high-performance 10-GHz band-pass filter, right? Okay, maybe not everyone needs one, but if *you* need one, check out "RF," by Zack Lau, W1VT, where you'll find complete construction details.—*KE3Z, email:* ***jbloom@arrl .org***.

Building a Direct Digital Synthesis VFO

This DDS-based VFO requires few parts and is a good starting DDS project.

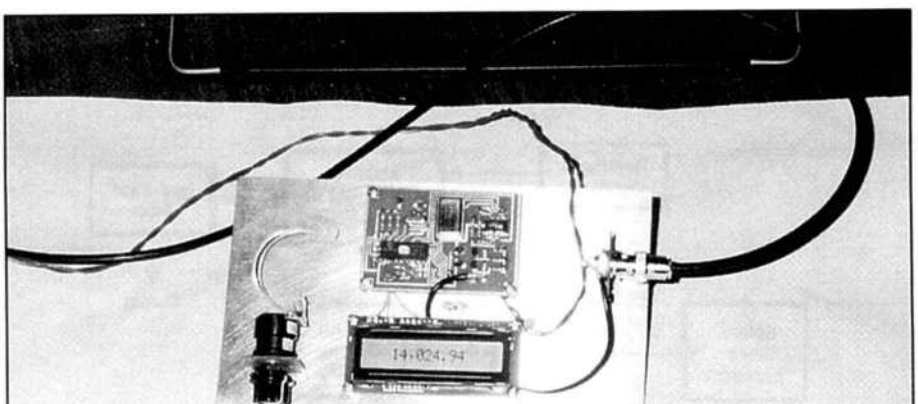
By Curtis W. Preuss, WB2V

Since direct digital synthesis, (DDS) was invented[1] in 1970, it has become more and more prevalent in the communications world—including Amateur Radio. Some interesting articles have been published in amateur-radio magazines explaining how DDS works.[2,3] However, if you have an interest in building something with DDS, finding a related construction project is a problem.

The purpose of this project was to try out DDS by building a DDS-based VFO. The project was deliberately kept as basic as I could make it. It was very tempting to add a few bells and whistles since DDS has many capabilities, but there were enough problems to solve as it was. Of course, the first problem was to get hold of a DDS integrated circuit. Some Web surfing revealed that a number of companies are supplying DDS chips or hybrids. These range from commodity CMOS devices that have clock rates up to

[1]Notes appear on page 7.

5150 Timberidge Ct SE
Rochester, MN 55904



The DDS VFO. The rotary shaft encoder is at the lower left.

125 MHz to specialty GaAs devices with clock rates over 1 GHz. Corresponding prices range from about ten dollars (in quantity) to several thousand dollars. Several of these companies also supply evaluation boards for their devices. These would be fine for a technical evaluation, but I wanted to apply DDS to an amateur-radio project. I decided to base the VFO on a CMOS DDS chip, the AD9850, that was announced by Analog Devices Inc in April of 1996. An AD9850 and a few other components were used to build a VFO that can tune from 100 kHz to 20 MHz in 10-Hz steps.

Project Description

DDS chips such as the AD9850 convert a reference oscillator input into a sine wave output at a frequency selected by the user. Basically, this project provides a mechanism to set the AD9850 for a desired frequency. See Fig 1 for a block diagram of the project. A complete schematic is shown in Fig 2. The rotary shaft encoder shown is for dialing in a frequency. A microcontroller monitors the outputs from the rotary encoder. The microcontroller then translates the rotary encoder signals into frequency control data, which is loaded into the DDS chip. Likewise, the microcontroller translates the selected frequency into data for display on the liquid-crystal display. The reference oscillator provides a digital clock to the DDS chip. The purpose of the low-pass filter is to smooth the digitized sine wave output of the DDS chip.

Microcontroller

A quick look at the schematic shows it doesn't take many parts to use the AD9850. The control program stored on the microcontroller contains most of the complexity involved with this project. The nice thing about programs is that they are easy to replicate. A complete listing of the source code for this project is available at **ftp://ftp.arrl.org/pub/qex**. The microcontroller is a low-cost 8-bit device from Microchip Technology.[4] It has an on-chip EPROM for storage of 512 instruction words and 24 bytes of RAM for data storage. The program for this project uses 474 instructions and requires 23 bytes of RAM, so adding any features would require moving up the micro-controller product line to a device with more memory. The EPROM programmer used to load the code onto the microcontroller is from Parallax. This programmer comes with an assembler and debugging program. The manual for the programmer contains several sample projects that were very helpful in getting started. The debugger program also proved to be invaluable. I also put an EPROM eraser to good use.

The DDS Chip

The AD9850 is a complete DDS chip. It contains a 32-bit phase accumulator, a 14-bit look up table and a 10-bit digital-to-analog converter (DAC). It can be clocked at 125 MHz to produce a 41-MHz sine-wave output. The spurious-free dynamic range is greater than 50 dB at 40 MHz with a 125-MHz reference clock. A complete data sheet for the AD9850 can be downloaded from the Analog Devices web site at **http://www.analog.com/pdf/ad9850.pdf**.

The frequency control word for the AD9850 can be loaded byte-wide or serially. The serial mode is slower but is used in this project to minimize the number of output pins required on the microcontroller. Serial mode is selected as the default by wiring pin 2 of the AD9850 to ground while pins 3 and 4 are wired to the supply voltage. Pin 25 is the serial data input, and pin 7 is the data write clock. After shifting in 40 data bits, pin 8 is used to transfer the data from the chip's input register to the DDS core. The 40 bits are a 32-bit frequency control word, 3 control bits and 5 phase-modulation bits.

The AD9850 data sheet gives Eq 1 for calculating the required control word. In order to minimize program size, the actual calculation uses the algorithm shown in Eq 2. This algorithm has some round-off error, but the error is less than 1 Hz, which is small enough to ignore in this application.

$$F_{out} = (\Delta Phase \times ClkIn) / 2^{32} \qquad \text{Eq 1}$$

Where: $\Delta Phase$ = value of the 32-bit tuning word, $ClkIn$ = reference clock frequency in MHz, and F_{out} = frequency of the output signal in MHz.

$$\Delta Phase = \sum_{n} LCD\_Digit \times Digit\_Weight \qquad \text{Eq 2}$$

Where: $\Delta Phase$ is the control word sent to the AD9850, n is the range of 1 to 7, *LCD_Digit* is a one of the seven digits being display on the LCD, and *Digit_Weight* is a precalculated value given in Table 1.

The AD9850 DAC output is a differential current on pins 20 and 21. A resistor placed from pin 12 to ground determines the full-scale output current for the DAC as given in Eq 3. The current equation is valid provided the voltage across the DAC output pins is less than 1.5 V. Setting the resistor to 3.92 kΩ yields a DAC current of about 10.2 mA. With the parallel load of the filter terminator and an external 50-Ω load, this current results in a voltage swing of about 250-mV peak-to-peak.

$$I_{out} = 32(1.248\text{V} / R) \qquad \text{Eq 3}$$

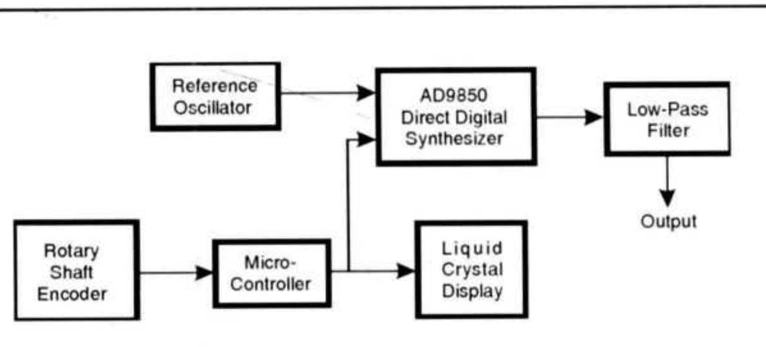
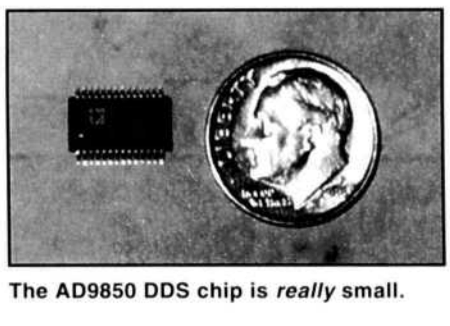


Fig 1—DDS VFO Block Diagram.



The AD9850 DDS chip is *really* small.

Rotary Shaft Encoder

The shaft encoder is optically coupled and has 32 detents. It has two digital output pins that cycle through a 00, 01, 11, 10 sequence as the shaft turns clockwise. Turning the shaft counterclockwise reverses the sequence. The control program monitors the shaft encoder outputs and decodes them to determine which way the shaft turned. At power-on reset the initial value of the VFO is 10 MHz. Each change of the encoder output causes the program to increment or decrement the LCD frequency display, followed by the calculation of a new frequency control word that is sent to the DDS chip.

The control program also counts the time elapsed between encoder output changes and will make bigger or smaller frequency steps depending on how fast the shaft is being turned. The particular rotary encoder shown in the schematic also has a built-in pushbutton switch. Turning the shaft with the switch closed causes the control program to change the frequency in 100-kHz steps. The combination of speed-controlled steps and 100-kHz steps allows rapid tuning across the VFO range. If there were only 10-Hz steps it would take 62,000 revolutions of the tuning knob to cover the VFO tuning range!

Liquid-Crystal Display

A wide variety of low-cost liquid-crystal displays are available, with different digit sizes or with back-

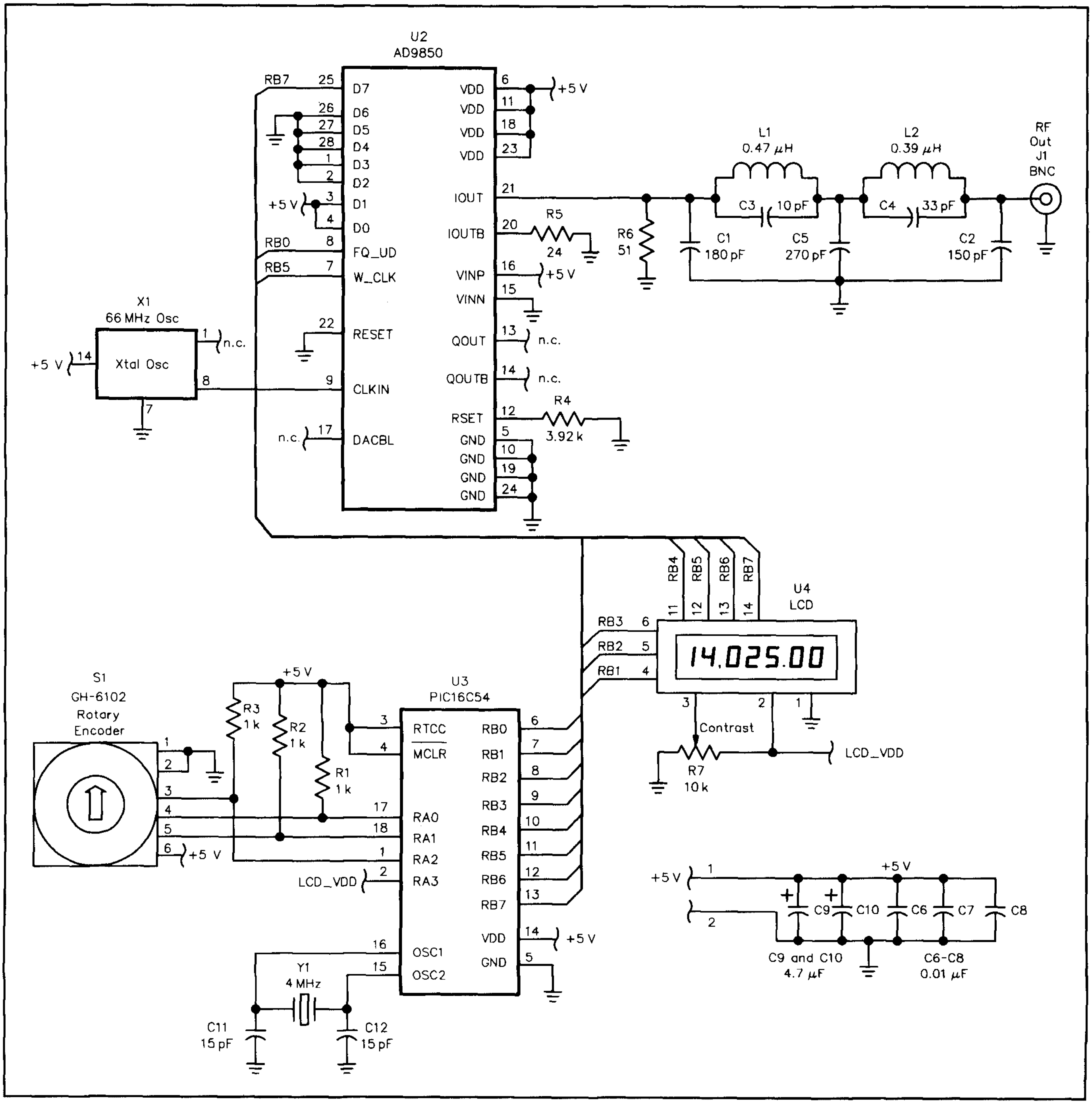


Fig 2—DDS VFO Schematic.

lighting. Fortunately, most of them use a common method of accepting data and control information. The control program assumes an LCD having a 16 by 1 display format. At power-on, commands are sent to the LCD that configure it for four data input bits instead of the normal 8-bit data mode. This minimizes the number of microcontroller pins required.

On the schematic, note that power for the LCD is being supplied from an output pin on the microcontroller. The reason for this is that the LCD was fussy about the turn-on time of the supply voltage and would not always reset properly. Using the microcontroller output pin for LCD power allows consistent power-on resets. Potentiometer R7 adjusts the LCD contrast.

Reference Oscillator

The reference oscillator is a standard clock oscillator module. The accuracy of the reference oscillator directly determines the VFO output accuracy. If the reference oscillator has a 100-ppm tolerance, so does the output. Clock oscillator modules up to 66.666 MHz are readily available. For many vendors, higher frequencies are special order items. Choice of a reference frequency will depend on the application. Changing the reference frequency requires updating the control program values in Table 1. One factor in the choice of reference frequency might be the locations of spurs. All DDS systems will have low-level spurious outputs.[5] The frequency of these spurs is very predictable. They are related to the reference clock frequency and harmonics of the output frequency.

Low-Pass Filter

The output of the DDS chip is a digitized or sampled sine wave. Such a wave shape has strong frequency components at the reference clock frequency plus or minus the output frequency. Filtering out these components produces a clean sine wave. For this project, with a reference clock near 66 MHz and a maximum output frequency of 20 MHz, the low-pass filter must cut off frequencies above 46 MHz while passing frequencies below 20 MHz. The fifth-order elliptic low-pass filter[6] shown in the schematic has 55 dB or greater attenuation at frequencies above 46 MHz. The filter requires a 50-Ω termination.

The calculated passband ripple of the filter is about 1 dB. However, a more significant amplitude variation can occur as the output frequency increases. As the output frequency goes higher, the digitized sine wave is constructed from fewer samples per cycle. At the DAC output pins, the wave shape begins to look less and less like a sine wave. As this happens the spurious frequency components constitute a larger portion of the total output power; meanwhile the desired output component is less.

Construction

A drawback to the AD9850, from a home-builder's point of view, is the package. The device is in a 28-lead SSOP, (shrink small outline package).

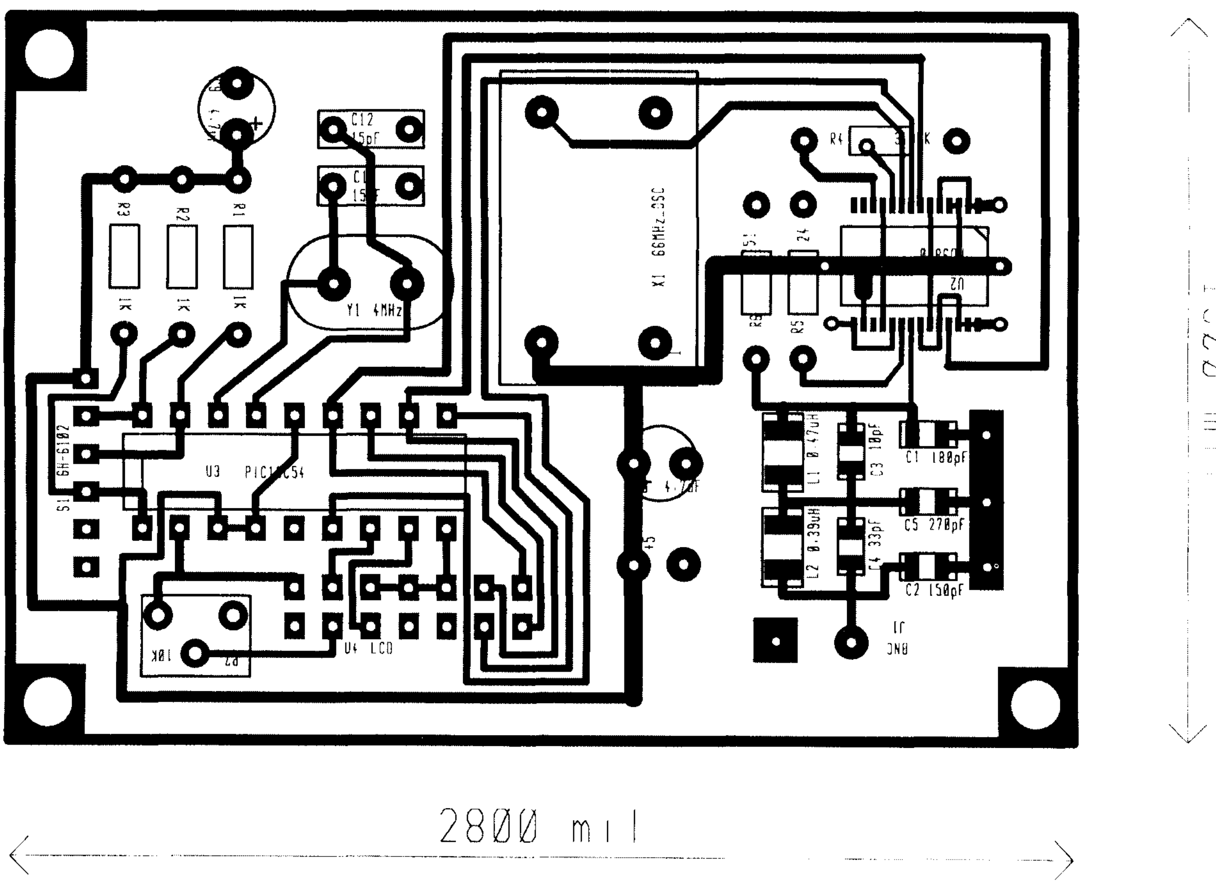


Fig 3—Component Layout.

The word for this package is tiny. The small size and close lead spacing made construction a challenge. A printed-circuit board is necessary. The use of surface-mount packages seems to be standard practice for suppliers of DDS chips. Some use leadless chip carriers, others use quad flat packs or small outline formats. Although the packages vary, experimenting with DDS means dealing with small packages.

The circuit board for the project was made with the use of iron-on transfer paper.[7] The layout was done using a program called *WinBoard*.[8] Fig 3 shows the layout that I used. One word of caution: this layout isn't necessarily an example of good layout practices. Simplifying board fabrication was the goal. The wiring around the DDS chip is very tight. I used a 600-dpi laser printer to print the pad patterns for this chip.

The board is double-sided, but only the component side is etched; the back side is a solid ground plane. Clearance on the ground side of vias was etched manually with a small drill bit. Eyelets were placed in the vias under the crystals. Decoupling caps for the DDS chip and clock module were placed on the ground side.

After making the board, the next challenge was getting the DDS chip soldered in place. With a 25-mil lead pitch, unintended solder bridges form between pins as if by magic. In a commercial environment the method for mounting such chips is to put a solder paste on the circuit board pads through a precisely made silk screen. The chips are positioned with a pick-and-place machine, then the whole board is sent through a heating chamber to reflow the solder paste. Lacking such equipment, I used a small soldering iron, some 25-gauge solder and a magnifying glass. There may be a better way, but at least one way to put the chip down is to tack it at one corner. Check and adjust the pin-to-pad alignment until it is right then tack the other corners. Once that's done just solder the pins, letting bridges form where they may, then use copper desoldering braid to wick the pins clean. The pads for the DDS chip demand careful use of the soldering iron. Too much heat or pressure will tear these off the board.

Table 1: Digit Weights for a 66.666-MHz Reference Clock

LCD_Digit	*Digit_Weight (in hex)*
10's	284
100's	192A
1k	FBA9
10k	9D49B
100k	624E13
1M	3D70CC1
10M	26667F90

Table 2: Parts List

Reference	*Value/Description*	*Part Number*
C1	180-pF chip capacitor	PCC181CGCT-ND
C2	150-pF chip capacitor	PCC151CGCT-ND
C3	10-pF chip capacitor	PCC100CNCT-ND
C4	33-pF chip capacitor	PCC330CGCT-ND
C5	270-pF chip capacitor	PCC271CGCT-ND
C6,C7,C8	0.01-µF chip capacitor	PCC103BNCT-ND
C9,C10	4.7-µF tantalum	P2011-ND
C11,C12	15-pF ceramic disk	P4014A-ND
L1	0.47-µH chip inductor	DN10471CT-ND
L2	0.39-µH chip inductor	DN10391CT-ND
R1,R2,R3	1-kΩ, 5%, 1/4 W	1KQBK-ND
R4	3.92-kΩ, 1%, 1/4 W	3.92KXBK-ND
R5	24-Ω, 5%, 1/4 W	24QBK-ND
R6	51-Ω, 5%, 1/4 W	51QBK-ND
R7	10-kΩ potentiometer	31G14-ND
U2	DDS chip	AD9850BRS
U3	Microcontroller (not-programmed)	PIC16C54/JW-ND
LCD1	16×1 liquid crystal display	73-1016-ND
S1	Rotary shaft encoder	GH6102-ND
Y1	4-MHz crystal	CTX006-ND
X1	66.666-MHz crystal oscillator	XC228-ND

Except for the AD9850 DDS chip, all these parts are known to be available from Digi-Key. Any equivalent parts can be used. A list of distributors for the AD9850 is available from Analog Devices. However, at the time of writing only FAI Electronics Corporation, (800-707-6040) had AD9850s in stock. Retail price was $21.00. For convenience, the Digi-Key stock numbers are shown. The EPROM programmer by Parallax is available from Digi-Key as is an EPROM eraser from the Walling Co.

Conclusion

While this project was probably overkill as the VFO for a basic radio, it was a great excuse to try out new things. DDS has significant advantages in many applications and its use will become increasingly common. There are a number of modulation techniques that can be used with DDS. Some CMOS chips with modulation capabilities are available. DDS can also be combined with phase-locked loop circuits to produce some interesting systems. Lots of possibilities!

Notes

[1] *IEEE Transactions on Audio and Electroacoustics*, Vol. AU-19, No. 1, March 1971, pp 48-56.

[2] Bergeron, "Direct Digital Synthesis; An Introduction," *Communications Quarterly*, Summer 1993, p 13.

[3] Cahn, "Direct Digital Synthesis—An Intuitive Introduction," *QST*, August 1994, p 30.

[4] Data sheets are available at **http://www.microchip.com/**.

[5] Bar-Giora Goldberg, *Digital Techniques in Frequency Synthesis*, pp 107-118.

[6] Stefan Niewiadomski, *Filter Handbook; A Practical Design Guide*, Chapter 2.

[7] For more information on this process see the web site at **http://www.dynaart.com/**.

[8] A demo version of this program is available at **http://www.ivex.com/**.

Meet the Vacker: The Simple, Stable VFO You've Been Looking For

This simple, classic VFO offers performance any builder will appreciate.

By Mark L. Meyer, WUØL

We've all had the experience—worked 'em, built 'em, used 'em, bought 'em or heard 'em. The experience I'm talking about is fussing with rigs with an unstable VFO.

With the proliferation of kits and homebrew QRP rigs over the past three or four years, including major commercial products with some well-respected brand names, there have been some pretty bad signals floating through the ether. But there isn't any good reason for the problem of an unstable VFO. Vacker developed a circuit decades ago that is simple and provides uncommon short- and long-term stability without the use of fancy phase-locked loops or other control circuitry. It is ready for use as soon as you turn on the power.

An examination of the circuit (Fig 1) gives insight as to why this is so. The tank inductor, L1, is not in the usual gate-to-ground configuration. The inductor is connected between the drain and gate. Then both the drain and gate are shunted to ground by very large capacitors, C1 and C2. L1, C1, C2 and C3, along with capacitors C4-C7 and tuning diodes D1 and D2, form a series resonant circuit that determines the VFO frequency. The key to stability is that C1 and C2 are many times larger than the internal device capacitances of Q1. Therefore, when the internal capacitances change as the device warms because of current flow, they have very little effect on frequency. Of course we must make sure that the inductor and capacitors are temperature stable to begin with.

The circuit shown in Fig 1 operates between 4.570 MHz and 4.646 MHz. This is for use with a 30-m rig that uses an IF of 5.528 MHz. You can easily modify the values of the inductor and the capacitors for any frequency you wish. Just multiply the values listed by the ratio of your desired frequency to this design frequency of 4.570 MHz. For example, if you wish to have a frequency of 3.500 MHz, you multiply C1-C7 and L1 by 1.3 (4.57/3.5 = 1.3). Frequencies higher than 4.75 MHz will result in a multiplier value less than one. Then all your components will have values smaller than those shown in Fig 1.

The basic Vackar oscillator circuit, using a good inductor and NPO or COG disc-ceramic capacitors or polystyrene capacitors, is very stable. When pow-

[1]Notes appear on page 11.

14153 West First Drive
Golden, CO 80401

ered up, without the tuning diodes connected, the frequency may drift downward 30 Hz or so in the first few minutes. Introducing variable-capacitance diodes for tuning adds a source of drift. I have not found these diodes to be stable without temperature compensation by use of capacitors having a negative coefficient. This is described below. If you wish to use a high-quality variable capacitor for the main tuning control instead of a diode, you can eliminate this problem.

I wanted to use diode tuning because of the ease of panel layout. Diodes also provide an easy method of implementing RIT (receiver incremental tuning). For frequency read-out, I use the KC-1 kit module put out by N6KR, Wilderness Radio, that provides the frequency read-out with audio. The KC-1 sends you the frequency in CW when you push a button by keying its sidetone oscillator, which is routed to your audio amplifier. It is accurate to within 1 kHz. It is easily programmed, using your keyer paddle, to accommodate any offsets required by your heterodyne scheme. It also provides a built-in electronic keyer. It is a great unit.

Nothing is critical in the values for the resonant frequency. All you have to do is get close with C1, C2, C3, C4, C5 and L1 and then start trimming. When testing the frequency, be sure you have the tuning diodes hooked up with—voltage on them—so their effect on frequency is known. Then you can use various values for C6 and C7 (or even more capacitors) to trim the frequency to exactly where you want it. Or a small air-insulated variable trimmer capacitor can be substituted for C7.

For the larger capacitors, C1 and C2, I used polystyrene capacitors because they were available. The 660-pF value was obtained by paralleling a 560-pF unit with a 100-pF unit. Polystyrene capacitors are usually very stable. COG or NPO disc-ceramic capacitors would also be a good choice for C1 and C2. C3-C7 and C11 should be COG or NPO disc-ceramic capacitors. Use good quality capacitors. Some of the miniature surplus units, even though they have a black painted top, are marginal. Digi-Key has a good selection of suitable capacitors. Use the 100-V or greater variety as they are physically larger and won't heat up as much with RF current flowing through them. In fact, using two parallel capacitors of smaller value to make up a larger value helps distribute the current and keeps the heat down, enhancing stability.

The trick to compensation and frequency trimming is patience. Tack-solder in a capacitor and let the circuit cool for an hour. Then fire it up and see how the frequency is affected. If you don't have a frequency counter, a synthesized general-coverage receiver will do just fine. Figure on spending a few minutes at a time for several evenings. Keep written records of changes made so that things don't get confusing.

D1 is the main tuning diode. I used an MV1662. This is a high-capacitance diode of about 250 pF at 4V. Other diodes can be substituted if you cannot find an MV1662, including paralleling devices with lower values. An NTE 618 or an MVAM109 AM tuning diode is a good substitute. These diodes provide even more capacitance, so series capacitor C8 can be reduced in value or the voltage range applied to the diode by the pot can be restricted. The 50-pF N150 capacitor in series with D1 limits the tuning range that can be obtained. A larger value will give you greater range; a lower value, less range. The negative coefficient (N150) compensates for drift introduced by the diode. The 10-kΩ tuning pot is a high-quality 10-turn unit. This provides very nice bandspread and smooth tuning. If you opt to use a high-quality variable

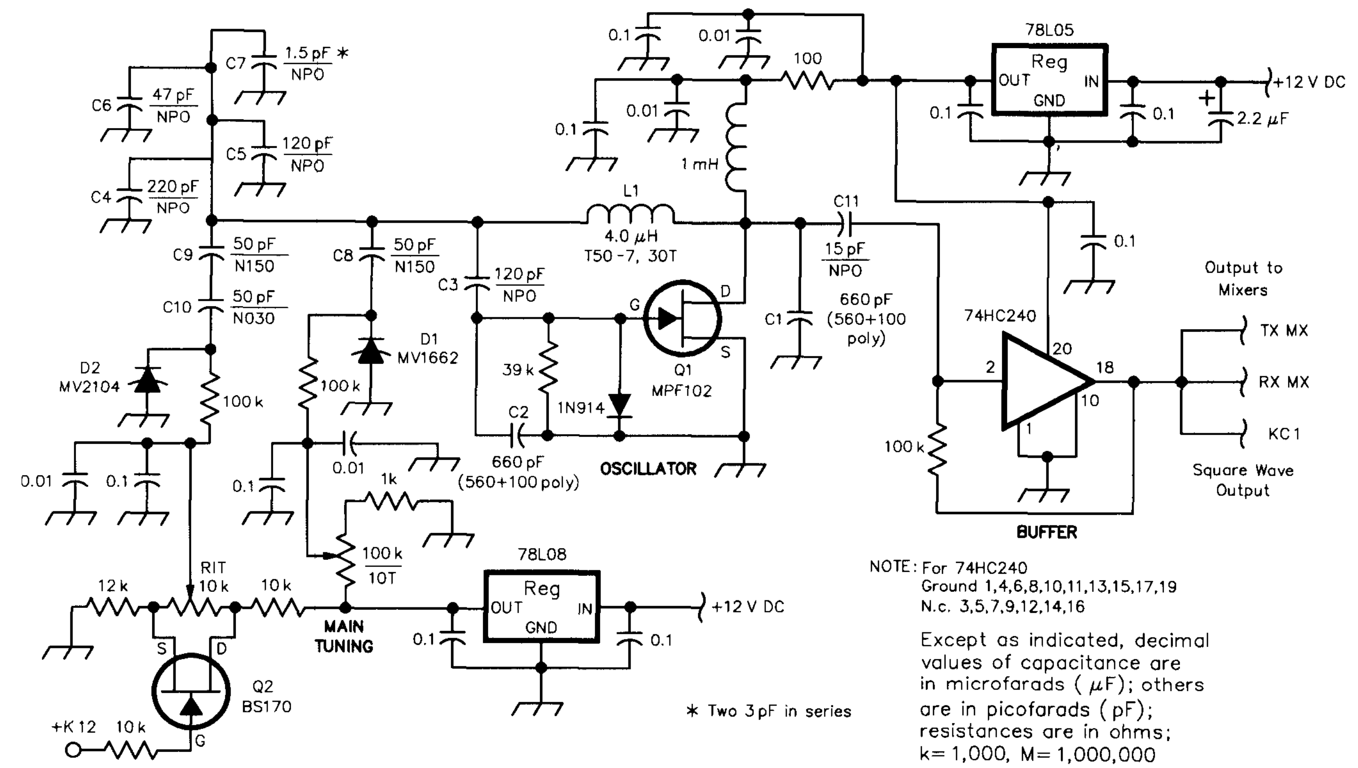


Fig 1—Vacker VFO and buffer.

capacitor instead of using D1 for tuning, a neutral coefficient using an NP0 capacitor will probably be suitable.

D2 provides the RIT. The MV2104 provides 13 pF at 4 V. C9 and C10 in series limit the range on my unit to about 5.5 kHz. Temperature compensation is performed as described for the main tuning diode. When +12 V is present during transmit, the BS170 is turned on and the 10-kΩ pot is shorted. This centers the frequency as determined by the 10-kΩ and 12-kΩ resistors on either side of the pot. My set-up allows me to tune up about 1.5 kHz and down about 3.0 kHz from the frequency centered when +12 V is present. This unequal split was done to facilitate listening down for DX when the DX operator instructs you to transmit up. Experiment with the resistors to give you the set-up you want. The 10-kΩ pot should be a high-quality unit and is single-turn.

Most oscillator circuits drift down as they heat up. Negative coefficient capacitors are the way to compensate for this drift. The capacitance of a negative coefficient capacitor goes down as it heats up. The lower capacitance causes the oscillator frequency to go up. Therefore it compensates for the natural drift down and makes the whole circuit frequency stable with regard to heat. If your circuit is under-compensated, it will drift down when powered up. If your circuit is over-compensated, it will drift up. Johnson Shop Products can provide a good selection of negative coefficient disc ceramic capacitors. The larger the N value on the capacitor, the more it will compensate for a given temperature change.

L1 is critical to stability. The ideal inductor would be wound on a ceramic coil form. However, solenoid type inductors should be mounted at least a diameter away in all directions from the wall of your enclosure for good stability. I needed to use a small box for my transceiver, so I chose to use a toroid core. The best core material with regard to temperature stability is #7 mix. I used a T50-7 core wound with 30 turns of #24 enameled wire to obtain 4.0 mH. After winding the toroid, I boiled it in water for two minutes. This is to relieve the stresses in the wire caused by winding it and presumably leads to better stability. I can't vouch for this, but it sure is easy so I do it. Then I coated the winding with two layers of Duco cement. Q-Dope or similar material also can be used. The #7 mix (painted white) is less common then #2 or #6 but is worth the effort to obtain. If you cannot find #7, use #6; it is more stable than #2. Buckeye Electronics can supply the #7-mix toroids.

The circuit in Fig 1 typically will not drift at all (my frequency counter is only accurate to 10 Hz) when first powered up. After five minutes or so, it will drift down 10 or 20 Hz. Then it will drift up past the original frequency perhaps 40 Hz in an hour. After that it will stay within 20-30 Hz for hours on end. My ear cannot distinguish a 20 to 30 Hz change in a beat note. This is with the unit sitting bare on the bench without being in a box at all. My work bench is in the furnace room where the temperature varies several degrees every time the furnace kicks on. As a test, I stuck the whole unit in the refrigerator for a couple hours at 42 degrees. Then I pulled it out and fired it up at a room temperature—72 degrees. The frequency changed 900 Hz, 90% of that in the first 25 minutes.

The circuit is built using Neanderthal surface-mount techniques. It is built on a piece of good-quality single-sided printed-circuit board. Be sure to use single-sided board as double-sided board forms small capacitors between the sides that are unstable. The connection pads for the components are formed by gouging out the unwanted copper with a utility knife gouge blade. Use whatever technique you like, but make sure things are mechanically stable. Make sure the ground leads from the capacitors that determine the frequency and their connection to the inductor all happen in one general area so the RF current doesn't have to go

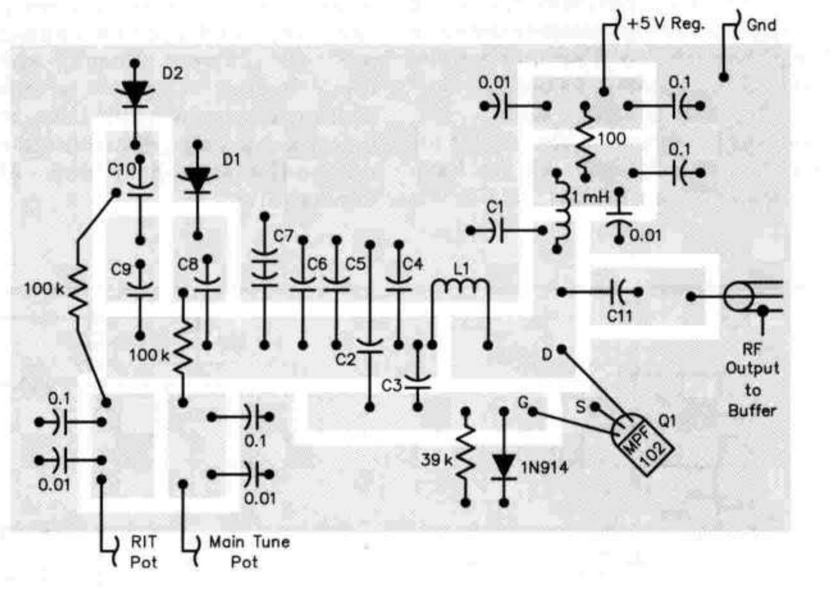


Fig 2—Parts placement.

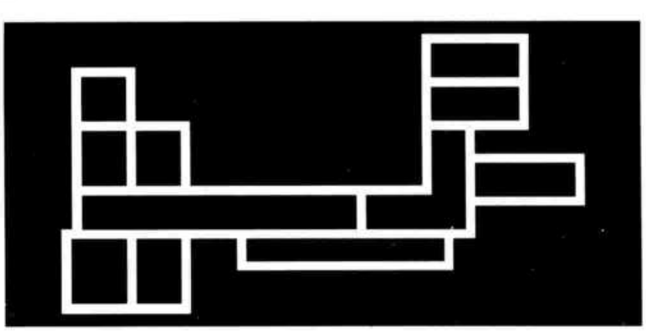


Fig 3—PC-board layout, full size.

running all around the board. I applied a generous coating of three-minute clear epoxy to the capacitors and the inductor to make the whole works mechanically stable when I was finished. This does not seem to affect the frequency once it cures for a few hours.

Only 5 V (regulated) is applied to the oscillator circuit. This keeps current and heating low. My circuit oscillated fine down to 4 V in a test. 8-V regulated is used with the tuning diodes to obtain greater range in tuning. Note the liberal use of bypass capacitors. It is best to keep unwanted RF out of your VFO circuits. I probably overdid it, but what's cheaper than a handful of bypass capacitors?

Only the oscillator and tuning diodes are mounted inside the VFO enclosure. The voltage regulators, buffer and tuning pots are all outboard. The dc voltages are brought into the enclosure using 1000-pF feedthrough capacitors. A small length of mini-coax brings the output to the buffer. Be sure to mount your VFO in its enclosure when making the final frequency adjustments. The enclosure will affect frequency somewhat.

For a buffer I have shown a digital CMOS circuit using a 74HC240. This buffer appeared in an April 1995 *QST* article by Lew Smith, N7KSB ("A Remote-Oscillator High-Frequency VFO"). This circuit provides a high input impedance so it doesn't load down the oscillator very much. The output is a square wave that mixers of the NE602 type prefer. If you use the 74HC240, be careful not to zap it with a static charge because it is CMOS. A simple FET buffer will work very well also. Look in the published VFO circuits for examples. The gate of an FET buffer also provides very high impedance and thus isolates the oscillator from the other circuits. Any buffer will load the oscillator slightly and thus change the frequency. When trimming your frequency, connect it to the buffer so you account for this effect.

Often hams don't want to tackle building VFOs. This shouldn't be the case. With this circuit, and a few general guidelines, building a VFO is a snap. It is fun to build, fun to test and calibrate and offers great performance; it's simple, it's stable.

Parts Suppliers

Tuning diodes:
Dan's Small Parts, PO Box 3634, Missoula, MT 59806-3634.
DC Electronics, PO Box 3203, Scottsdale, AZ 85271-3203.
Any NTE parts supplier.

NP0 and C0G capacitors:
Digi-Key, PO Box 677, Thief River Falls, MN 56701-0677.

Negative coefficient capacitors:
Johnson Shop Products, PO Box 2843, Cupertino, CA 95015.

#7-mix toroid cores:
Buckeye Electronics, 10213 Columbus Grove Rd, Bluffton, OH 45817.

KC-1 keyer and frequency counter:
Wilderness Radio, PO Box 734, Los Altos, CA 94023-0734.

About the Author

Mark L. Meyer, WUØL, was first licensed in 1965 as WNØNSY when he was a high-school senior. He graduated with a Bachelor's Degree in Electrical Engineering from South Dakota School of Mines and Technology in 1970 and has worked in the area of hydroelectric power generation and transmission since then. Mark is currently involved in the operation of the high-voltage electric grid in the western United States. His first transmitter was a homebrew, and he has been homebrewing ever since.

Exploring the 1:1 Current (Choke) Balun

Sound use of electromagnetic principles reveals what the current balun does—and does not—really do.

By William E. Sabin WØIYH

The goal of this article is to explore certain basic aspects of the 1:1 current, or choke, balun transformer (often called the Guanella balun[1]) that is widely used in radio equipment, antenna feed lines and antenna tuners. Several topics are included that have not been covered in detail in previous literature. The balun made with coax and ferrite cores will get special attention, and we will look mostly at the basic principles of the 1:1 balun in order to limit this article to a reasonable size.

Balun Description

Fig 1 illustrates an example, with a floating load and a connection at the input to some large reference plane that we will call ground. Sometimes it is very difficult to say exactly what ground is, and quite often it has no definite relationship to Mother Earth. Ground can mean one thing at one end of a long coax cable and something else at the other end. An impedance of some kind between these two locations can often be measured.

The differential input signal V_{IN} is transported, essentially intact, from an unbalanced input, through the transmission line (a pair of conductors with dynamic characteristic impedance Z_0) to the balanced load, R_L, which in this case is equal to Z_0. At the same time, a high impedance is presented to some external voltage source V_S (which may have a series impedance Z_S) that is common to both output terminals. In this sense the balun device is a *transmission line transformer*. The two windings, tightly coupled, in addition to being a transmission line transformer, also constitute a 1:1 *conventional Faraday* transformer. This important, sometimes overlooked, fact is easily verified both experimentally and from basic principles that will be discussed later. Also, we assume that the winding reactances of *both* conductors are large, but not infinite.

The various measured voltages with a floating load (no V_S or Z_{GND}) are as shown. The impedance from each output terminal direct to ground (excluding the transformer windings) is an open circuit. Voltages V_1 and V_2 are measured from the output terminals to ground using a high-impedance probe. The input V_{IN} (differential) is transmitted to the output with almost no voltage translation. That is, V_2 is almost at ground potential and V_1 is al-

1400 Harold Dr SE
Cedar Rapids, IA 52403
email: **sabinw@crpl.cedar-rapids.lib.ia.us**

most the same as V_{IN} (measurements confirm this, ignoring for now minor stray effects at higher frequencies).[2] Also, forward current I_F equals return current I_R, so the net current $(I_F - I_R)$ is zero, and therefore the flux in the cores, and therefore also V_{CM} (*common-mode*, or *longitudinal* voltage), are essentially zero (more about this later). That is, assuming pure inductive reactance in the windings,

$$V_{CM} = (I_F - I_R)X_B = I_{CM}X_B \qquad \text{Eq 1}$$

where X_B is the winding reactance and I_{CM} is the common mode current. In this article the subject of core flux will be related to V_{CM}, the voltage drop across each winding of the balun. The reasons for this will become clear presently.

If, as an option, some point on the load resistor R_L in Fig 1 is connected *directly* to ground ($Z_{GND} \approx 0$) as implied, the situation of the previous paragraph changes. Assume initially that the impedance of each of the two windings is much greater than R_L. The output voltage V_{IN} divides between the two resistor values. These voltages will adjust themselves so that very nearly the same value of current flows in each of the two segments of R. That is, because A and B are floating, the path from A-to-ground-to-B must have very nearly the same current. The two windings also have almost the same currents in opposite directions. For example, if a point near the top end is grounded, V_2 goes almost to V_{IN}. If R_L is grounded at the exact center, V_1 and V_2 then become very nearly equal to $V_{IN}/2$. This situation is approximated, as just one example, in antennas with "grounded" driven elements.[1]

If the grounded point of R_L is its center, V_{CM} now becomes, by inspection of Fig 1, $V_{IN}/2$, and we must now start to think about the presence of significant amounts of magnetic flux in the core(s) of the balun, as discussed in the next section. The presence of some kind of impedance path to ground at the output of the balun is critical to the design considerations.

The statements of the preceding two paragraphs need some clarification. With a direct ground connection (Fig 1, $Z_{GND} \approx 0$), the two opposing currents in the windings are not *exactly* equal.[2] There is a slight difference (which escapes to ground) and this difference, when multiplied by the large but finite reactance of the balun coils, produces V_{CM} (see Eq 1) and also the flux in the core(s).

[1]Notes appear on page 20.

$$\Lambda = L(I_F - I_R) = L\, I_{CM} \qquad \text{Eq 2}$$

where Λ is the balun flux linkages and L is its inductance. It is this flux that forces the two currents to be almost equal. It does this by creating $-V_{CM}$ across the bottom winding. This $-V_{CM}$ forces nearly all of the current to return through the bottom resistor and the bottom balun wire. Without the balun, the current to ground will not return through the transmission line back to the generator but will return directly to the generator via the ground path ($Z_{RET} \approx 0$). Without this return current in the transmission line, the line will radiate like an antenna because the magnetic flux generated by one of the line's conductors is not cancelled by the flux from the other line. This is true for a two-wire transmission line and it is true for a coax transmission line. With this perspective we have related balun current I_{CM}, voltage V_{CM}, flux and line radiation into a unified picture and we see why the balun is important and also how it does its job.

An intermediate and more complicated situation is that the ground connection to R_L has an impedance Z_{GND} associated with it, for example the wavelength of a ground connection or some other impedance or electromagnetic coupling in the system (near-field or far-field). An exact circuit analysis of this situation can sometimes be difficult, so a safe approach would be to assume a direct short to ground from the center of R_L as a worst case. This avoids a certain amount of guesswork and speculation. On the other hand, if this impedance is greater than zero (quite often it is), there is a division of common-mode voltage V_{CM} between it and the balun windings (Fig 1), and this reduces the stresses on the balun cores. If this situation can be reliably identified, the robustness of the balun design can possibly be relaxed considerably. One path to ground to consider is the balun itself and its feed line.

As R_L becomes greater than Z_0, and becomes comparable to the impedance of the windings, the accurate voltage division between two equal segments of R_L (with a ground connection in Fig 1), and therefore the equal current through each segment, deteriorates (verified by measurements). The same effect occurs at low frequencies, where the impedance of the windings becomes small.[1] In both cases, the approximation to a perfect (or ideal) transformer model is no longer valid.

Also, if power is constant, then as R_L (with ground path connected) increases, V_{IN} (Fig 1) increases and V_{CM} obviously increases, too. This is something to watch for. For a truly floating load, this effect is not nearly as severe. Moreover, for a constant value of V_{IN}, as R_L becomes large (or very small) at higher frequencies where the electrical length of the transmission line is a significant fraction of a quarter wavelength, a voltage transformation (step up or step down) occurs within the transmission line and this also increases V_{CM} and the flux. Usually, this effect is small.

For a certain value of power delivered to the resistive part of the load, any increase in load reactance in series with this load resistance will also increase the load voltage. If this load also has a path to ground so that V_{CM} becomes large, the increase in core flux can become large and increase the stress on the cores. Beware of this, too. In applications such as antennas, this path to ground very often exists in the form of near-field coupling to the local environment or other antenna elements. In some antenna types, a connection to a boom or a tower is involved. Another very fa-

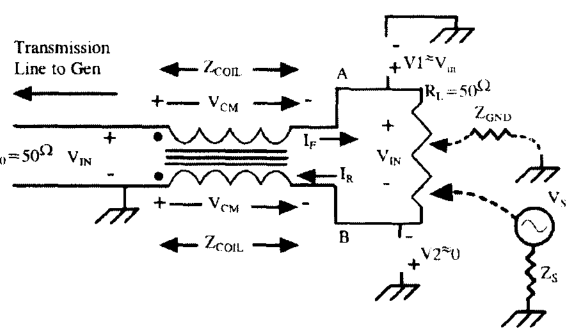


Fig 1—A 1:1 current balun with grounded input.

miliar problem is the current that is induced or conducted by an antenna onto a transmission line, for example the outside of a coax cable. We want to "choke off" this current to prevent feed-line radiation and also prevent contamination of the radio equipment. This can also increase V_{CM} (more about this later).

If the load at the balun output is a transmission line that has a high input impedance (due to a high VSWR) and a path to ground, changing the line's input impedance so that its magnitude is closer to Z_0 can improve the current balance and also reduce the V_{CM} voltage. This improvement can often be made by changing the transmission line length (this doesn't change the VSWR except very slightly) or by inserting a lumped L-C network or transmission-line stub (this *can* reduce the VSWR). Quite often a change of just a few feet in line length will be enough to make a big improvement. It is often possible (and often done) to find a low-impedance point in the line where a balun can be inserted.

Core Flux Density

In Fig 1, the rms common-mode V_{CM} voltage establishes a flux density in the core of a toroid according to Faraday's law:

$$B_{MAX} = \frac{V_{PEAK}}{0.0444\, A_E\, N\, F} \text{ (gauss)} \qquad \text{Eq 3}$$

where B_{MAX} is in gauss, V_{PEAK} is the peak of sine-wave voltage (1.41 times V_{CM}), A_E is the toroid core cross section in cm^2, N is the number of turns and F is in MHz. For specified values of F and A_E, the critical factor for flux density is the volts-per-turn (V_{PEAK}/N) as observed on any one of the windings. The diameter of the toroid is not mentioned explicitly in this formula but is determined by the number of turns N and the wire size or coax diameter that must be used to limit power loss in the copper. This formula is also used for inductors and power transformers, where a stacking factor for laminations is included.

For a string of beads (each bead is a toroid) with coax through the holes, the cross section A_E is the height of each bead times its thickness $T = \frac{1}{2}$ (O.D.–I.D).[3] We then require N beads, each with a certain voltage value, to support V_{PEAK} within the required B_{MAX} limit. For a long string of thick beads, B_{MAX} can be made quite small. This string also has less of the capacitive coupling effects that we find between adjacent turns in toroids. In either case this coupling reduces the parallel self-resonant frequency beyond which the common-mode impedance becomes capacitive and starts to decrease in magnitude. Nevertheless, beyond this parallel resonant frequency the impedance value can still be very large up to much higher frequencies. This is a common occurrence in RF chokes of all kinds. Series resonances must be avoided. See the "Experiments" section for data about the bead balun.

Amidon Associates,[4] in their reference manual, has a table of recommended values of B_{MAX} for different frequencies. These values are based (somewhat loosely) on measurements of temperature rise and apply to both powdered iron and ferrite. Using Eq 3 and the values of B_{MAX}, the maximum suggested voltage across the windings in Fig 1 can be calculated. These values of B_{MAX} are far below the saturation levels of 2000 gauss for ferrite and 5000 gauss for powdered iron and therefore help to ensure adequate intermodulation properties (especially the iron cores). They ensure a long life and reliability with a good safety factor. The increases in hysteresis, eddy currents and dielectric losses cause the reduction of the B_{MAX} rating at high frequencies. Also shown in the table is the product $B_{MAX} \cdot F \cdot 0.044$ which, from Eq 3, when multiplied by A_E, gives Amidon's suggested maximum volts-per-turn at a particular frequency. These numbers actually get larger at high frequencies.

Any RF choke relies on its magnetic flux and Lenz's law for its choking action; the choke balun is no exception. That is, if the common-mode voltage V_{CM} increases for any reason, the increasing common-mode current creates an increasing flux whose induced counter-emf (V_{CM}) opposes the action that is trying to increase the common-mode current (Lenz's law).

The voltage V_{CM} creates a magnetic field in the cores. If the balun cores are lossless the energy that is stored in the magnetic field in one part of the cycle is returned to the circuit in another part. If the cores have losses (ferromagnetic materials always do) then some power is "robbed" from R_L in each half-cycle. The power loss in the balun is V_{CM}^2/R_{BALUN}, where R_{BALUN} is the *parallel* resistive component of the balun's impedance at a particular frequency. Equipment such as a vector impedance meter or an RF impedance bridge is needed for this measurement.

A good approach to the design is to measure the V_{CM} across the balun windings (if they are accessible), using an RF voltmeter, then use Eq 3 to determine the flux density and compare that value with the Amidon table. Touching the balun (if it is accessible) to see if it gets more than slightly warm after a long key-down is an *ad hoc* approach that is often used, but the voltage measurement method is more credible and is preferred, at least at the start. For meaningful results, these tests should be made with the balun in its actual operating location. The touch test is always a good idea as a final checkout. If these tests are not feasible, calculated or, possibly, simulated estimates (or deliberate overdesign) will have to do. See the

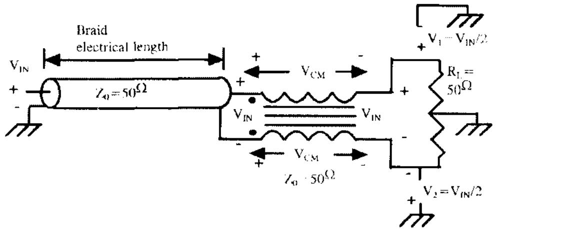


Fig 2—A balun with a long coax at the input.

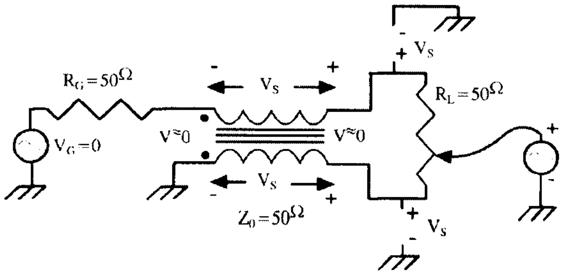


Fig 3—A balun with a common-mode stray voltage source.

"Experiments" section for more about this topic.

The flux ratings in Table 1 depend on temperature rise. In a later section, "Experiments," we will look at balun power dissipation calculations based on laboratory impedance measurements as a supplement to this discussion of Eq 3 and Table 1. When it is not feasible to make lab measurements, Table 1 and Eq 3 are a good starting point. We will also see that if a large core surface area can be made available for heat dissipation, the numbers in Table 1 are conservative. The bead (W2DU) balun is an excellent way to get this large area, when it's practical. In some systems, for example Yagi arrays, the toroid or rod balun is probably more practical from a mechanical standpoint.[1]

Table 1. Flux Density vs Frequency

Recommended Flux Densities

Frequency (MHz)	B_{MAX} *(gauss)*	$B_{MAX} \bullet F \bullet 0.044$
1	500	22.0
3.5	150	23.1
7	57	17.6
14	42	25.9
21	36	33.3
28	30	37.0

Conventional Versus Transmission-Line Transformer

If a certain coaxial (or wire) 1:1 balun, using some specific core, is connected as a conventional transformer with R_L across the secondary, say the center lead is the primary and the braid is the secondary, or vice versa, the voltage across each winding is V_{IN}. As a current balun, as shown in Fig 1, V_{CM} is quite small unless there is a ground connection at the center of R_L, in which case the common-mode voltage across each winding is $V_{IN}/2$. So the flux density in the core, for a given number of turns, is only reduced to 50% in this particular case. The power loss in the core would be reduced by 6 dB (which is substantial). If the ground is at A in Fig 1, there would be no reduction at all. Also, the conventional transformer has some difficult leakage inductance and capacitive problems that are less (but not negligible) in a broadband balun.

For the toroid or bead balun, with their better control of stray effects, we can use more turns in order to further reduce flux density, but we must keep the winding length much less than a quarter wavelength, say no more than 1/8 wavelength, at the high end of the frequency range (this is true also for the conventional transformer). Also, we can only go so far with this idea because, as I mentioned before, the high-frequency choke impedance will start to decrease beyond its self-resonant frequency because of distributed capacitance. This effect is more pronounced with high-μ cores (as toroids or bead strings), as measured by myself. In either case, I find it better to use large, medium-μ toroids with more turns, or more and thicker medium-μ beads, as the way to reduce flux density.

We might think that we could use a higher permeability μ to reduce flux density in the balun, but from Eq 3 we see that although the common-mode impedance may increase (this is an important goal on the lower frequencies), the flux density does not depend much on permeability. For example, for a constant value of V_{CM}, if we increase μ the common-mode magnetic field H_{MAX} is reduced, but since $B_{MAX}=\mu H_{MAX}$, B_{MAX} remains just about the same. (In practice, V_{CM} may increase a little.) Also, the higher permeability materials often tend to have increased losses at the higher frequencies.[1] If we use large, medium-μ cores with more turns or many large, medium-μ beads, we also get more surface area for heat dissipation, and this is a good idea.

The Ungrounded Balun

A somewhat different situation for the choke balun is shown in Fig 2, where a longitudinal impedance, for example a coax braid, separates the balun ground terminal from the reference ground plane. If the electrical length of the braid (velocity factor about 0.95) is an odd multiple of a quarter wavelength, the ground terminal of the balun input is at a high impedance with respect to actual ground and the common-mode voltage V_{CM} (and therefore the core flux) becomes very small. In fact, the balun is now floating and could possibly be eliminated. But if the length is an even multiple of a quarter wave, a virtual ground is present at the balun input and the arrangement of Fig 1 is a better model. In practice, the actual braid impedance can lie anywhere between these two extremes. If the impedance to ground at the balun input in a particular system is unknown or can vary for some reason, the worst-case situation of Fig 1 would ordinarily be assumed, and the balun would be designed for the maximum expected value of flux density. A careful system analysis is always a good idea.

Stray Generators

In Fig 3 a stray undesired *voltage*, V_S, is applied at the output to some arbitrary point on the load. Using the superposition principle for linear circuits, we temporarily set the normal input generator to zero. A voltage of some magnitude and phase with respect to ground then appears at each output terminal. Because of the combination of the large reactance of each winding and the tight magnetic coupling between the two windings, these two voltages will be forced to be nearly equal in both amplitude and phase. And if they are nearly equal, the small currents through the windings will also be nearly equal since they both have large and very nearly equal reactances. This is the mechanism by which the balun tries to maintain equal currents in both windings when there is a V_S. Note also that since the bottom lead at the input (Fig 3) is at a voltage ground, the top input lead is almost at a virtual voltage ground. However, a small leak-through voltage will appear across R_G. The total V_{CM} is the composite of the stray generator and the other causes we discussed before (V_S is also a path to ground).

On the other hand, if the stray generator is a perfect *current* source, the balun doesn't present a sufficiently large impedance. One-half of this constant current flows through R_G, and it can sometimes be large. The other half returns directly to system ground. Usually, the stray generator is something between a voltage source and a current source, and the current into R_G and ground can vary over a wide range, depending on the impedance of the stray generator. This is a basic limitation on the current balun that has to be considered in any system design where a stray generator is present. It is desirable, if possible (it's not always possible), to avoid a high impedance (current source) in this generator.

An alternate approach worth considering is to put an isolation transformer between the signal source V_G and the input of the balun. This breaks up the current path back to R_G. Fig 4 shows how this is implemented. The grounded center tap (optional) provides an escape route to ground for the stray current, if one is deemed necessary (it may not be). The opposing magnetic fields due to the $I_S/2$ currents cancel in the primary winding, and the Faraday shield within the isolation transformer helps to shunt currents to ground and away from R_G. The conventional isolation transformer could present a design problem, but if R_L (the load) is 50 Ω or so, it is fairly easy to build a good wideband 1:1 transformer using powdered-iron cores.[5] The Faraday shield can be the open-circuited braid of the coax used for the primary winding. One problem is knowing what to do with the center tap, since this current does not get into the generator output R_G but does get into the ground system of the radio equipment where it might cause trouble (and RF exposure). A short, low-inductance auxiliary path to an outdoor RF ground is one approach (where possible) that I have used to divert unwanted ground currents. Be cognizant of electrical code requirements when choosing a ground connection and lightning protection. A legal water pipe connection or some other ground dump (but not the house wiring system) or counterpoise may be a compromise approach. Or, it may be possible to not have *any* ground connection (let it float), but in this case an RF choke would be desirable as a static leakage path to electrical system ground if the system does not already have one.

The Balun as a Conventional Transformer

We have emphasized that the balun being considered has some of the properties of a conventional Faraday transformer. That is, there is a very efficient electromagnetic coupling between the two windings. If the windings are two identical wires, side by side, it is easy to see how this might be true because the visual similarity to a conventional transformer is more apparent. But for the case of a coaxial cable balun we should take a closer look. For example, it is often thought that the inner wire is somehow impervious to what is happening on the braid, especially on the *outside* of the braid. This is definitely not true at all, as I will show.

This does not suggest that the balun acts like a conventional transformer as it transfers the input power to the output load. It does declare that the electromagnetic coupling between the pair of wires, or between braid and center wire of a coax, is an important part of the balun function.

Fig 5 shows an electrically short segment of coax connected as a conventional transformer, with the RF generator V_G connected to the center wire. The current in this wire creates a magnetic H-field that encloses the wire as shown, in accordance with Ampere's law (for simplicity, only the H-field that is created by the *generator* is shown; that created by the load current is not shown). This H-field is tangential to the inner surface of the braid and is perpendicular to the direction of the coax.

This oscillating magnetic H-field creates an oscillating electric E-field which is an emf-producing field because it is a source of voltage (emf) in accordance with Faraday's law.[6] This E-field is in the same direction as the coax and is uniformly distributed along its length. It exists also within the dielectric material[7] and *within* the braid (it is continuous across the boundary[8,9]). V_{EMF}, the total emf induced, is the integral of this E-field over the length of the braid. This V_{EMF}, minus the voltage losses due to resistance and leakage reactance, is the terminal voltage V_L that appears across R_L. The voltage from one end of the braid to the other varies linearly from zero to V_L. A battery-operated RF voltmeter (Fig 8), isolated from ground, eliminates sneak paths that might cause false readings of the various voltages.

Note the relative voltage polarities and current directions in Fig 5. V_G and V_L are nearly the same, as a detailed analysis would show. The coax center-wire and braid impedances are not identical, and the coupling is very good but not perfect. Also, I_G is a little greater than I_L because of the magnetizing current that a transformer requires.

In Fig 6, the braid is the primary and the center lead is the secondary. In this case the braid current (the primary current) produces a magnetic H-field that lies entirely *outside* the coax (for simplicity, only the H-field that is created by the generator is shown; that created by the load current is not shown). For this reason, it would seem that a voltage could not be induced into the center wire since this H-field is not in contact with the center wire. It turns out that it is not necessary for the H-field itself to be in direct contact with the center wire. It is instead sufficient that the *E-field that is created by the H-field* be in contact with the center wire, and as a matter of fact, this is a common situation in nearly all types of conventional transformers in which the secondary winding is physically separated from the core (where almost all the flux is) by air and insulation. In these transformers the secondary winding encircles the core, and this is *precisely* what Faraday's transformer induction principle requires.[5,7,10] The same situation applies in Fig 6. Because of the common boundaries between the braid, dielectric and center wire, this transfer is quite efficient (see the "Experiments" section).[8] Again, I_G is a

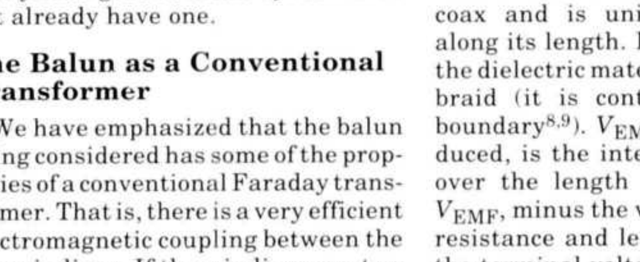


Fig 4—An isolation transformer diverts common-mode currents to an RF ground.

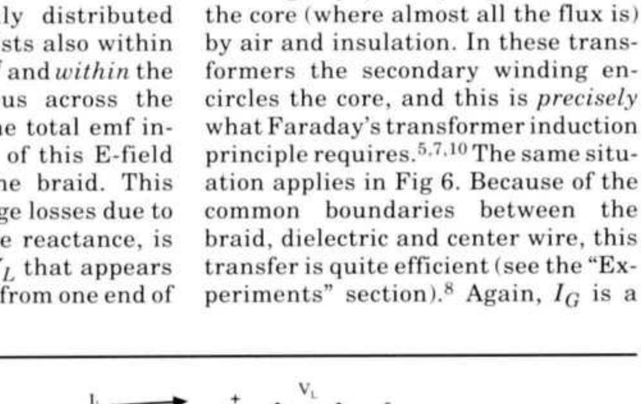


Fig 5—Coax connected as conventional transformer, center wire is primary.

little greater than I_L.

It might be thought that the coupling is due to dielectric capacitance rather than electromagnetic, but this is not true—although my actual measurements suggested that this capacitance may have a very small influence on the transfer. Looking at Fig 6, each location on the center wire is at about the same potential as the corresponding location on the braid. Therefore, there is at most only a very small displacement current through the dielectric capacitance at that location. There are also fringing effects at the ends of the coax to consider. In summary, the shield of the coax is a pretty good transformer primary, just about as good as the center lead in Fig 5.

In view of this electromagnetic coupling, there is another fact to consider. A current on the outside of the braid, created by some external activity, produces a skin effect on the outside of the braid. This current also produces a longitudinal voltage drop along the outside of the braid, due to its resistance, reactance and wavelength. This same longitudinal voltage drop also appears on the *inside* of the braid. That is, there is *no* skin effect between the inside of the braid and the outside of the braid at any particular location that would sustain a significant voltage difference between these two surfaces. Skin effect doesn't work like that.[12] The braid *thickness* has almost no voltage drop anywhere. It is only the *current* that is confined to the outside.

This paradox is explained as follows: skin effect increases the inductive reactance of the braid below the outer surface.[11] Because of this higher reactance inside the braid, the internal current is very much less, even though the longitudinal voltage gradient along the inner surface of the braid is the same as on the outer surface. The copper on the inner surface behaves just like an insulator (with the typical insulator's internal voltage gradient) for this outer braid current. But if we try to apply a voltage *between* the inner and outer surfaces at some particular spot, the picture changes drastically. There is a very-low impedance (resistance and reactance, but no skin effect) for this voltage. In this respect, the copper is much different than a true insulator material such as polyethylene.[13]

This longitudinal voltage drop is then induced onto the center wire by transformer action exactly as we discussed previously. A balun would present a high impedance for both of these voltages. Because these two voltages are quite similar, the voltage difference between the two coax conductors is very small (but maybe not zero), therefore the two voltages do not produce much *differential* signal in the coax. It is by this mechanism that a coax in mid-air shields against longitudinal external influences, even though the braid (in mid-air) is not grounded. The "Experiments" section gives experimental proof of this behavior and the theory involved is solid basic principles.

A possible source of error with this explanation, at the higher frequencies and most notably for toroids, involves stray capacitance between adjacent turns, where the signal on the braid tends to bypass the transformer coupling that we want, or to cause phase shifts that degrade the cancellation effect that we want.

If a coax braid is connected to a true ground plane all along its length, or is grounded frequently, this transfer of longitudinal voltage does not occur. But for an electrically long coax in mid-air or in a balun, the braid cannot always be considered to be close to ground potential, as we saw in the discussions of Figs 1 through 4. This behavior is easily observed experimentally, and it is almost exactly the same as in a two-wire balun. All things considered, the coax balun and the two-wire balun are equivalent in their behavior except for small differences in resistance and reactance of the braid and the center-wire of the coax. A solid shield has lower resistance and reactance and also less leakage than a braid but it is still a very good transformer primary, as verified by experiment.

Balun Impedance

To get the desired amount of common-mode current choking action from the balun, its common-mode impedance should be at least several (≥4) times as large as the sum of the source impedance R_G, the load impedance Z_L and the impedance Z to ground at the load, as shown in Fig 7. As mentioned before, it is questionable whether the braid longitudinal impedance can be included, but sometimes it can. R_G can be the output impedance of a transmatch or a PA and can often be difficult to know. Z_L is usually known fairly well using simulation software, by theory or by measurement.

In general, the best approach is to get as much common-mode impedance as possible over the desired frequency range, without letting the self-resonant frequency get too low. I prefer the bead string (W2DU) type for this reason when its use is mechanically feasible. The best way to verify good common-mode current reduction is by experiment. Use a clamp-on RF ammeter. I use the Palomar PCM-1 (low cost and excellent), but homemade meters are doable and not too difficult to calibrate with some effort.[14] Measure common-mode current with the balun removed. Slide the meter along to find a location of maximum current. Then install the balun

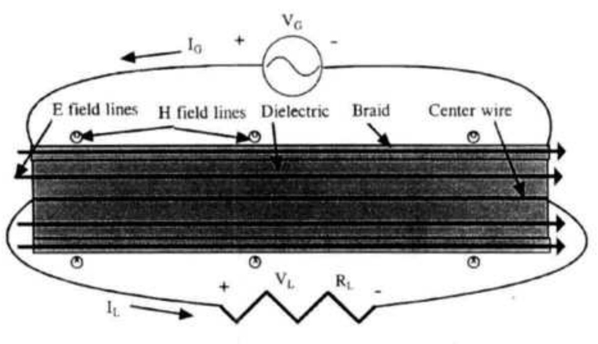


Fig 6—Coax connected as conventional transformer, braid is primary.

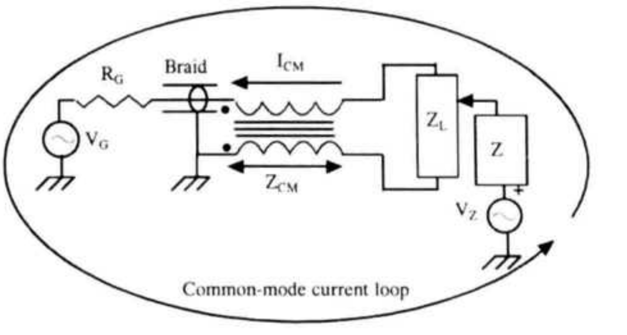


Fig 7—Common-mode current loop for a 1:1 current balun.

and slide along and measure again to see how much the maximum current is reduced.

Common-mode voltage measurements on the balun itself, as previously described, are excellent indicators of balun effectiveness and adequate design, when it is feasible to make them. It is worthwhile to go to some trouble to do this—if it can be done without personal physical risk. Two ways to make this measurement are: 1) connect temporarily to the braid at both ends of the balun, or 2) run a very thin enameled wire through the bead string and use that as a one-turn secondary. This voltage measurement can be made at a low (measured) power level and the voltage V_{CM} extrapolated to a high power level using the formula

$$V_{HI} = V_{LO} \bullet \sqrt{\frac{P_{HI}}{P_{LO}}} \qquad \text{Eq 4}$$

Common practice is to put a balun right at the antenna feedpoint, to improve the symmetry of the radiation pattern and to reduce conducted antenna currents onto the line. For a feedpoint at a high impedance this is not a good idea, especially with high power and high V_{CM}. A balun at a high-impedance point causes impedance bumps (irregularities) in the system that can complicate the efficient flow of power. They can cause dBs of power loss, which is important even at low power levels.[15] The guidelines suggested in this article should help to reduce the guess work, or at least to add additional understanding. The balun design depends on the lowest and highest frequencies, the common-mode voltage V_{CM} (or equivalently, common-mode current I_{CM}), the self-resonant frequency and the temperature rise.

Another important use for a balun is at the equipment end of the line, where it can often help to reduce the transfer of household noise up to the receiving antenna. It also provides a significant reduction of parallel currents, induced by the antenna, into the station. The one right at the antenna does not always do these jobs well enough.

Keep in mind that the coax balun increases the longitudinal impedance of the center wire as well as the braid. This important fact is often overlooked.

Experiments

This section describes three experiments to augment and verify some of the conclusions in the previous sections.

The measurements of Figs 5 and 6 were made using the simple RF voltmeter in Fig 8. A signal generator supplied the RF, and a battery operated DMM was isolated from ground to eliminate sneak paths that might affect the readings. As the movable probe slides along the braid (3 feet long), the voltage rises linearly from zero to a maximum value. The readings of the braid and center lead were within 1% of each other, showing that the length of coax is a very good conventional transformer across the HF spectrum that was tested. This circuit is suitable only for 1- to 10-V signals.

The test setup of Fig 9 demonstrates that the toroid induces a longitudinal E-field (in the neighborhood of the toroid) in a 6-foot RG58 coax, that it produces a voltage from one end of the braid to the other end, and that this longitudinal E-field penetrates the braid and dielectric and appears within 1% from one end to the other of the center conductor. The differential measurements at either end were at least 40 dB less than the end-to-end voltages. These results are the same with one resistor, both resistors or open circuits, which eliminates capacitive coupling as a possible explanation. The assertion about true-transformer coupling within the coax is certainly confirmed by this simple test. Also, the optional circuit shown in Fig 9 verifies that the emf and current on the braid that are due, for example, to antenna current are duplicated in the center conductor. A dual-beam 100-MHz Tek scope shows equal amplitude, in-phase V_1 and V_2 if $Ra = Rc$ and $Rb = Rd$. If $Ra \neq Rc$ (most probable in practice) the transformer-coupled option eliminates any problem due to this unbalance at the left end. Without this balance, an induced field can produce a differential signal at the output that might, for example, deceive even a well-shielded SWR bridge. A two-wire line behaves in exactly the same way as the coax. There is a common belief that the coax braid is a kind of cocoon that always isolates (more or less) the center wire from the outside world. Definitely not true; electromagnetic principles *never* take a holiday.

Fig 10 is a plot of the impedance magnitude and parallel resistance (measured with a vector impedance meter) of some bead-type (W2DU) baluns. One (Fig 10a) uses 50 type FB-77-6301 beads, μ=2000, $L_{LOW\ FREQ}$ = 100 µH, and another (Fig 10b) uses 100 FB-43-6301 beads, μ=850, $L_{LOW\ FREQ}$ = 82 µH. These beads both have an O.D.=0.953 cm, I.D.=0.493 cm, H=1.04 cm, and therefore A_E=0.24 cm^2. They fit on RG58/59 with the vinyl jacket removed. If we assume a worst-case ground at point A in Fig 1 and a ground right at the input, the values in Table 1 (at 7 MHz) and Eq 3 say that for Z_0 and R_L=50 Ω (please note the 50 Ω) the 50-bead balun (type 77) is okay for 440 W and the 100 (type 43) balun is okay for 1770 W, continuous duty at 7 MHz. Because of the better heat dissipation of the large sur-

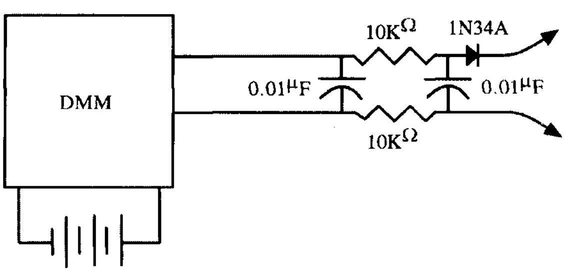


Fig 8—Diagram of an RF voltmeter using a battery operated DMM (digital multimeter).

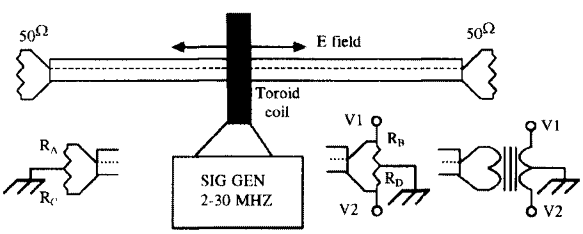


Fig 9—Test setup to verify penetration of the shield onto the center wire by a longitudinal E-field on the outside of the coax. Also shown are ways to ensure cancellation of the differential output.

face area of the beads, as compared to a toroid, the values in Table 1 are a little pessimistic. Also, the ground connection at point A in Fig 1, rather than near the center, is usually pessimistic. The low-duty cycle of SSB/CW (where applicable) also helps.

The 100-bead balun is definitely a better balun. That is, more medium-μ beads are better than fewer high-μ beads. Beyond the resonance peaks at about 6 MHz the impedance is capacitive, but in Fig 10b the impedance and parallel resistance of the type-43 balun are better over the entire frequency range, except possibly at 1.8-2 MHz.

Now refer to Fig 10c. For the same length of the bead string (≈3.3 ft), 100 of the thicker type FB-43-5622 beads, A_E= 0.40 cm^2, will fit over the RG58/59 vinyl jacket, have 28% more inductance and handle 2.8 times as much power (in a 50-Ω load) as the smaller beads, under the same worst-case conditions as the previous two examples, and according to Table 1 and Eq 3. This balun could also use 36 of the longer type FB-43-5621 cores.

For this balun, from the plot of parallel resistance in Fig 10c, at 28 MHz, 1500 W into 50 Ω would produce $(237)^2/1600/100$ = 0.47-W per bead with a worst-case ground at point A in Fig 1 (verified <5°F at 850 W). For a ground at the center of R_L (more realistic) the dissipation is 0.12 W (6 dB less) per bead. In a bead string this would be okay. We might expect further improvements in an actual system, as we discusssed before, but they would have to be verified. This is an excellent balun for the entire 1.8- to 30-MHz region, but we must keep in mind that 50 Ω is the preferred load and it should be within an SWR of, say 2:1, or must be derated accordingly and verified, either experimentally (preferably) or by computer simulation (if possible).

A detailed study of the resistance values in Fig 10 suggests that at, higher power levels, a temperature check, especially in the 10- to 30-MHz range, is an advisable precaution, especially if long key-down operations at maximum power are planned for some reason.

The self-resonant frequency of any balun involves stray capacitance, possibly affected by the packaging. The longer string of medium-μ beads should be less sensitive in this respect. Note also that the self-resonant impedance may be large compared to other system impedances in series

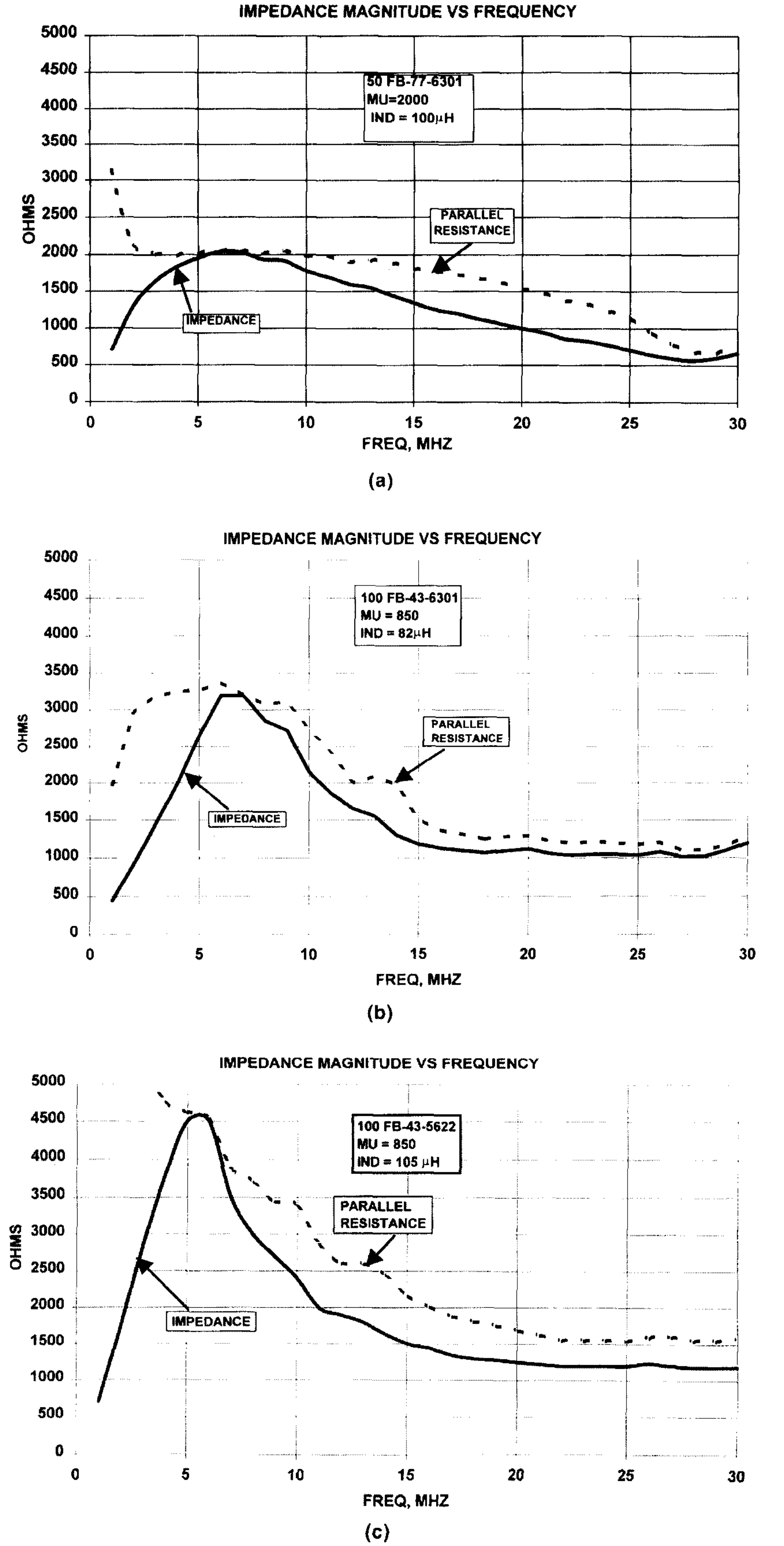


Fig 10—Frequency vs impedance magnitude and parallel resistance for three bead-type (W2DU) baluns. a) 50 type FB-6301-77 beads, b) 100 type FB-6301-43 beads and c) 100 type FB-5622-43 beads.

with the balun, and nearly all of the V_{CM} voltage may appear across the balun windings rather than being distributed as we considered in the "Balun Description" and "Balun Impedance" sections. But the high value of parallel resistance (Fig 10) will minimize power dissipation at these frequencies.

Last, but certainly not least, SSB-intermodulation distortion (IMD) caused by cores is much less than PA distortion if we use many large, medium-µ beads (or large toroids) and if we design conservatively (this is highly recommended). As the final step, IMD integrity should be verified by spectrum measurements of the *radiated* signal under SSB QSO conditions with a cooperating local operator, possibly using a narrowband CW receiver as a spectrum analyzer. Temperature rise of the balun may not always be a completely reliable measure of IMD performance, and verification on all of the bands in use is well advised and justified (and good citizenship).

Conclusions

For some additional information and test data about baluns, Note 16 is recommended. Note 1 provides a wealth of valuable practical information about balun construction. I also appreciate the very helpful comments and suggestions by Frank Witt, AI1H. This article has tried to relate the basic balun-design principles to the overall system design and requirements. Hard and fast rules can sometimes be difficult to come by, and judgment calls based on an understanding of basic principles can be very helpful.

The conclusions about controlling V_{CM} apply also to the baluns used in antenna tuners; this is an excellent way to keep an eye on them. Measure the voltage on any one of the windings and use Eq 3.

Notes

[1] Sevick, Jerry, W2FMI, *Transmission Line Transformers*, Second Edition, ARRL, 1990. Also, "Baluns Revisited," *Communications Quarterly*, Summer, 1992.

[2] The absence of voltage translation from input to output in a lossless, matched line is explained by the mutual cancellation by the two conductors, averaged over one RF cycle, of longitudinal electric fields. For this cancellation to occur, the forward current I_F and reverse current I_R must be equal in amplitude and opposite in phase. This cancellation automatically occurs at each location by virtue of E- and H-field interactions between the two conductors at that location, provided that the current path is complete and that none of the return current I_R is diverted from the line.

[3] Maxwell, Walter, W2DU, "Some Aspects of the Balun Problem," *QST*, March 1983, p 38. Also, Lewallen, Roy, W7EL, "Baluns: What They Do and How They Do It," *ARRL Antenna Compendium Volume 1*, ARRL, 1985.

[4] Amidon Associates, 3122 Alpine Ave, PO Box 25867, Santa Ana, CA 92799, (714) 850-4660. Their user's guide is a valuable resource for cores and uses.

[5] A link-coupled transmatch is also a good approach. Combined with a balun between the transmatch input link and the transmitter, this provides excellent isolation for common-mode currents. This approach is especially useful for balanced high-impedance transmission lines.

[6] Kraus, J. D., W8JK, *Electromagnetics*, Fourth Edition, McGraw-Hill, 1992. See Fig 10.2, p 422 for a discussion of Faraday's law. Also a footnote on p 422. See also p 194, section 5-10 for a discussion of E_e "emf-producing fields," which are *not* the same as the E_c fields that are due to static charges.

[7] Skilling, H. H., *Fundamentals of Electric Waves*, Wiley Books, 1948, p 80.

[8] Harnwell, G. P., *Principles of Electricity and Electromagnetism*, pp 544-545, McGraw-Hill, 1938. "If common boundaries exist, circulating currents induced in the shield produce emfs in the common bounding surface which react on the second surface that is to be shielded." See also p 313 regarding boundary relations.

[9] If the braid were looped into a short circuit, the E-field within the braid would be very small because of the loop's ability to carry a large RF short-circuit current. This current generates a large RF H-field whose RF E-field greatly cancels the E-field at the boundary and also within the braid. In the absence of this short-circuit current, a substantial RF E-field exists *within* the braid. The resistance (including a small radiation resistance), and imperfect coupling would also limit the loop current.

[10] Shadowitz, Albert, *The Electromagnetic Field*, Dover Publications, New York, 1975, p 389. The vector potential concept ($\nabla \times B = A$) is used for this explanation.

[11] Terman, F. E., *Electronic and Radio Engineering*, Fourth Edition, McGraw-Hill, 1955, pp 21-23.

[12] Johnson, W. C., *Transmission Lines and Networks*, Chapter 3, McGraw-Hill, 1950.

[13] The absence of a significant voltage drop between the surfaces of the braid of a coax casts considerable doubt on the concept of separate, independent currents, differing in phase, flowing on the inner and outer surfaces, especially in opposite directions. This would require substantial instantaneous voltage differences between the two surfaces.

[14] Lau, Zack, KH6CP, "A Relative RF Ammeter for Open-Wire Lines," *QST*, October 1988.

[15] Michaels, Charles, W7XC, "Physically Small Inductors May Lead to Unknown Losses," *QST*, August 1996, p 71.

[16] Witt, Frank, AI1H, "Baluns in the Real (and Complex) World," *ARRL Antenna Compendium Volume 5*.

RF

By Zack Lau, W1VT

A High RF-Performance 10-GHz Band-Pass Filter

Why do I call this a high-performance filter? It offers a very high ratio of Q to size and wastes very little volume in coax transitions or empty space. This is quite important when building a compact portable transverter.

The design is based on one of the very first 10-GHz filters I built—the 2-pole band-pass filter in the TNT, or Tuesday Night Transverter.[1] Unfortunately, the authors omitted two important details—the input/output coupling of the filters and the measured insertion loss. While their design had the potential for great performance, I doubt that many people realized it—even those who built one!

This filter is made from a 1.5-inch section of WR-90 rectangular waveguide. An enclosed cavity is formed by soldering brass sheet stock to the open ends of the waveguide. A pair of posts down the center line creates a pair of

[1]Notes appear on page 23.

225 Main Street
Newington, CT 06111
email: zlau@arrl.org

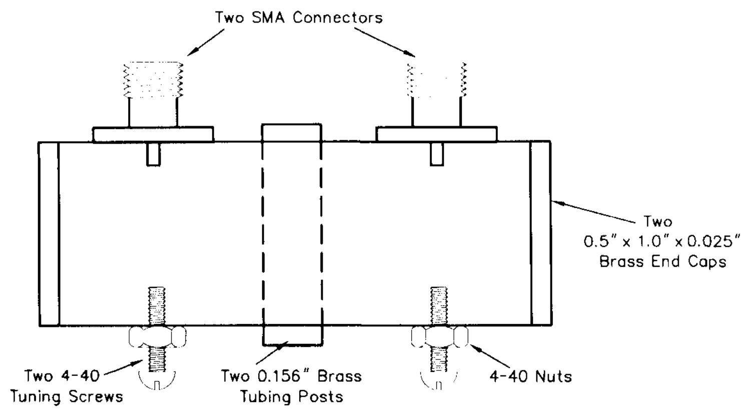


Fig 1—Diagram of the 2-cavity 10-GHz band-pass filter.

coupled cavities. The diameter of the posts determines the coupling—narrower posts increase the coupling between the cavities.

The filter is capable of surprisingly low insertion loss, if you can get the coupling adjusted properly. I chose to use the center conductors of captivated contact SMA connectors as probes. The hard part is adjusting them—just ask anyone who has spent hours filing away! My solution was pretty obvious—use shim stock. After all, isn't this what they sell shim stock for? I made a bunch of spacers of different heights—this allowed me to quickly adjust the height of the coupling probes. With this design, it appears necessary to adjust the probes and resonator tuning for lowest insertion loss. The tuning interacts with the probes, so it didn't appear terribly useful to look at the passband shape when setting the probe coupling. This differs from most other filters, where looking at the passband is an excellent guide.

Using a mixer and isolator to measure the insertion loss, the design shown in Fig 1 has 0.6 dB of insertion loss with a 3-dB bandwidth of 106 MHz. The 144-MHz IF image rejection is 33 dB. This should offer excellent performance for receive applications.

For transmit, it might be useful to trade a little insertion loss for better image rejection. With the original design's center coupling using a pair of 3/16-inch posts, the insertion loss increased to 1.3 dB with an image rejection of 47 dB. The 3-dB bandwidth is 36.7 MHz.

Construction

The first step is to get the waveguide, brass metal stock and SMA connectors. Either 2- or 4-hole flange SMA connectors will work, but it is important to use connectors with captivated contacts. Otherwise, the probes will move around, changing the filter alignment. I suggest using brass tubing since it's easier to solder. On the other hand, it might be practical to tap solid brass rod after it's soldered in place—so you can easily mount the filter with screws. Don't forget the clearance required for the 4-40 tuning screws.

As a starting point, I'd try 100-mil probes for the wider filter and 80-mil probes for the narrower filter. These lengths are shortened a bit by the waveguide walls—only 30 to 50 mils actually sticks into the cavity. Adding shim stock reduces the length of the probes, so you might cut the probes long and add shim stock to get the 100-mil starting lengths. The narrower bandwidth version with 3/16-inch posts required shorter probes—about 80 mils long.

Once you have the SMA connectors prepared, you can cut and drill the WR-90 waveguide. I filed the cut waveguide edges smooth. If you have 2-hole flange connectors, you need just half of the 2-56 mounting holes shown in the drilling guide. When drilling the holes for the posts, use several drill sizes to get up to the desired hole size. This will help to ensure a tight-fitting round hole that matches the tubing.

After deburring the holes and polishing up the waveguide, I soldered the posts into place. If you do a good job of drilling tight-fitting holes, it shouldn't be difficult to solder the end

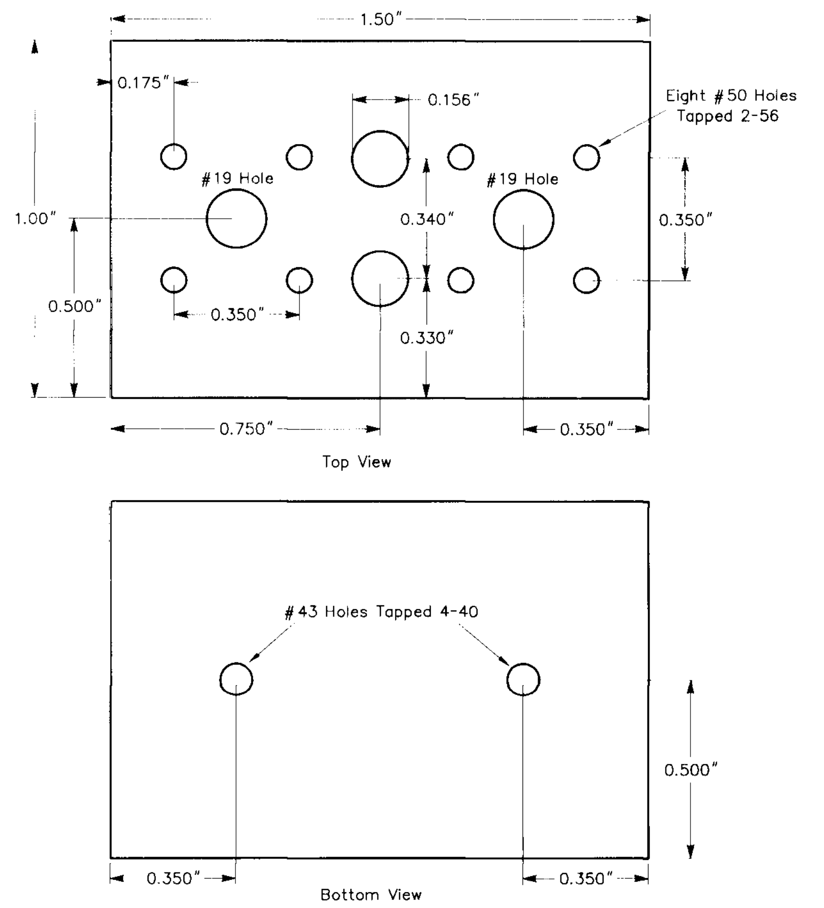


Fig 2—WR-90 waveguide drilling template.

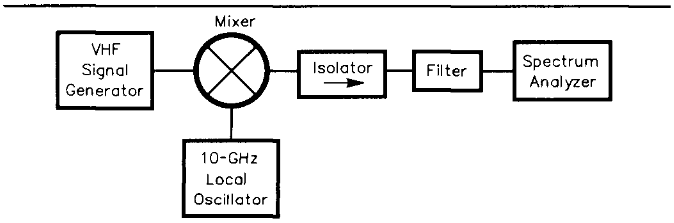


Fig 3—Diagram of a filter alignment test fixture.

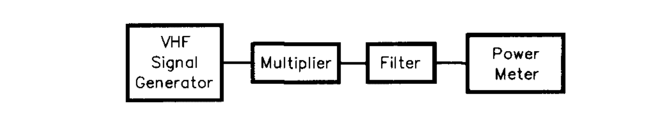


Fig 4—A simpler test fixture that doesn't work as well.

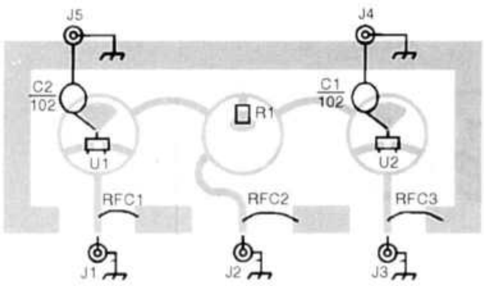


Fig 5—Dual 10-GHz mixer on 5880 Duroid.
C1, C2—0.001 μF, 50V ceramic disc capacitors.
J1-5, SMA panel-mount female jacks.
R1—51 Ω 1/10-W chip resistor.
RFC1-3—0.21-inch #28 wires.
U1, U2—HSMS 8202 diode pair.

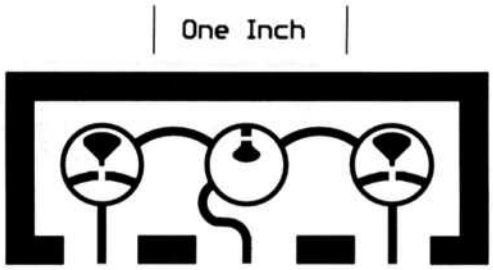


Fig 6—Full-size etching template for the microstripline mixer.

plates into place without having the posts unsolder themselves, even if you use a propane torch. I use a C-clamp to hold the end plates in place during soldering. I used ordinary 60/40 rosin-core solder.

To tune up the filter, I used a 10-GHz mixer/local oscillator to up-convert a signal generator to provide a suitable signal source, as shown in Fig 3. I prefer this technique because I have a spectrum analyzer available to sort out the various mixing products. It is a considerable improvement over the simpler setup shown in Fig 4 using a frequency multiplier and a power meter. The frequency multipliers that I have built are rather frequency sensitive, requiring a calibration plot. Because most filters reflect rather than absorb unwanted signals, the reflections often disturb the operation of mixers and multipliers. The isolator works well with a mixer since all the big signals are near the same frequency. In contrast, a multiplier may have strong signals distributed over a relatively wide frequency range. A resistive attenuator may work better with a multiplier than an isolator designed to work over just an octave.

I estimate that the tuning screws were inserted 0.10 inches into the cavities. Obviously, the exact tuning will vary due to construction tolerances. Adjust the length of the probes and the tuning for minimum insertion loss.

Fig 5 shows a dual-mixer circuit board designed for use at 10 GHz.

It uses 15-mil 5880 Duroid (ε_r=2.2). It's a slight improvement over the one in the June 1993 *QST*, since it uses less board area. This article explains how to tune these mixers for best performance.

A brass frame is soldered around the mixer board using 0.5 × 0.025-inch brass strips. The strips are drilled and tapped to hold SMA connectors. I use 2-hole flange connectors to offset the center pins of the IF connectors so they clear the ground conductors. I also tap holes on the two opposing strips without connectors to hold the aluminum cover. Fortunately, I have access to a metal brake to bend the cover. The best way to locate the holes is to install the cover and drill #50 holes through the cover and the frame. Then the cover holes are enlarged with a #44 drill to pass the 2-56 screws.

Notes

[1]Bailey, Kirk, N7CCB, Larkin, Robert, W7PUA, Oliver, Gary, WA7SHI, "TNT for 10 GHz," *Proceedings of the Microwave Update 1988.*

Feedback

More on "A Simple T/R Sequencer," October 1996 *QEX*

◊ Unfortunately, the T/R sequencer doesn't actually protect its outputs from reversed polarity, though the sequencer itself survives reverse polarity. Incandescent test lamps instead of LEDs are recommended for anyone experimenting with such circuits. The problem isn't easy to solve—power FETs contain a parasitic diode across the junction. Normally, protective circuits get around this problem by reversing the source and drain connections, but this doesn't work here. That fix would disable the use of the FET as an on/off switch—it would be effectively on all the time.—*Zack Lau, W1VT*

◊ Mario Miletic, S56A, suggests an improved PTT input circuit for "A Simple T/R Sequencer," that appears on page 16 of the October 1996 QEX. He also suggests that the decoupling capacitors C1 and C2 be increased to 10 nF, through the author thinks that 270 pF is more appropriate for a VHF/UHF/SHF station.

Upcoming Technical Conferences

Central States VHF Society Conference

The 31st Annual Central States VHF Society Conference will be held July 24-26, 1997, at the Clarion Resort on the Lake, Hot Springs, Arkansas.

A dinner and sight-seeing cruise aboard the *Belle of Hot Springs* riverboat is planned for Thursday evening; papers will be presented on Friday and Saturday; Friday evening you can enjoy the Mountain Music Jamboree, Hot Springs' original #1 music comedy stage show. Dinner will be available at Homestead's Restaurant prior to the show. Buses depart from the hotel; Saturday evening is the annual conference banquet; Sunday will wrap up the conference.

The special conference rate at the Clarion Resort on the Lake is $89 per night. Call and make your reservations at 501-525-1391, or toll-free at 800-432-5145. Don't forget to ask for the CSVHF rate.

There will be a family program offering tours of several area attractions.

For more conference information contact: Joel Harrison, W5ZN, 528 Miller Road, Judsonia, AR 72081.

Eastern States VHF/UHF Conference

The 23rd Eastern States VHF/UHF Conference will be held August 22-24, 1997, at the Harley Hotel, Enfield, Connecticut.

For information on this year's conference please contact: Fred Stefanik, N1DPM, tel: 413-569-0116, ext 211; Stan Hilinski, KA1ZE, tel: 860-649-3258; or Ron Klimas, WZ1V, tel: 860-768-4758.

Microwave Update '97

Microwave Update '97 will be held October 23-26, 1997, at the Holiday Inn Conference Center in Sandusky, Ohio.

A "Surplus Tour" is scheduled for Thursday; conference papers will be presented on Friday and Saturday; noise-figure measurements and a microwave flea market are planned for Friday night; Saturday night dinner will be a Bar-B-Q; and Sunday, conference wrap up and possible tour of the W9JK "Big Ear" at Ohio State University.

Registration before October 2, 1997 is $40; after October 2, 1997 it's $45. Conference fee includes one copy of the Proceedings. Additional copies are $10. Saturday night BBQ Dinner is $15.

Hotel rates at the Holiday Conference Center are: Single, $69.95 per night; Double, $95.90. Price includes buffet breakfast and lunch on Friday and Saturday and breakfast on Thursday and Friday.

A ladies program will be offered. Planned activities include sightseeing on the north coast area (Lake Erie) and shopping. For information on area events, contact the Lake Erie Visitor's Bureau at 1-800-255-ERIE, or check their Web page at **http://www.buckeyenorth.com/**.

For more conference information contact: Tom Whitted, 4641 Port Clinton East Road, Port Clinton, OH 43452; tel: 419-732-2944.

ARRL and TAPR Digital Communications Conference

The 16th Annual ARRL and TAPR Digital Communications Conference will be held October 10-12, 1997, at the Holiday Inn BWI Airport, Baltimore, Maryland.

This year's local host is the Amateur Radio Research and Development Corporation (AMRAD).

Call for Papers: Papers for inclusion in the proceedings are due by August 20, 1997. They should be sent to Maty Weinberg at ARRL HQ.

This is a conference for all, beginners to digital experts. Topics include: APRS, satellite communications, TCP/IP, digital radio, spread spectrum and more.

Friday will include an all-day symposium covering APRS; late Friday afternoon there will be a half-day seminar entitled "RF Basics for Computer Weenies: Helping the RF-challenged get the most out of the new high-speed wireless toys"; papers will be presented on Saturday; a seminar on "Spread Spectrum System Design and Theory" will be conducted on Sunday morning.

A block of rooms have been reserved at the Holiday Inn BWI Airport at the special rate of $89 per night. The rates are good on reservations made before September 9, 1997. (Rooms cannot be guaranteed after that date.) For reservation call the Holiday Inn BWI Airport at: tel: 410-859-8400 or fax: 410-684-6778. Ask for the Digital Conference rate.

Preregistration before September 10, 1997, is $42, after September 10, or at the door it's $47. Registration includes one copy of the Conference Proceedings, sessions, meetings and lunch on Saturday. Saturday dinner: $20. Seminars/Symposiums—Friday, APRS 1-8pm: $25; Friday, RF Basics for Computer Weenies 3-7pm: $20; Sunday, Spread Spectrum System Design and Theory 8:30 am-1:30 pm: $20.

To register for the conference, or for more information, contact TAPR at: 8987-309 E Tanque Verde Road #337, Tucson, AZ 85749, tel: 817-383-0000, fax: 817-566-2544 or Internet: **http://www.tapr.org/dcc/**.