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Oscillators and Multipliers

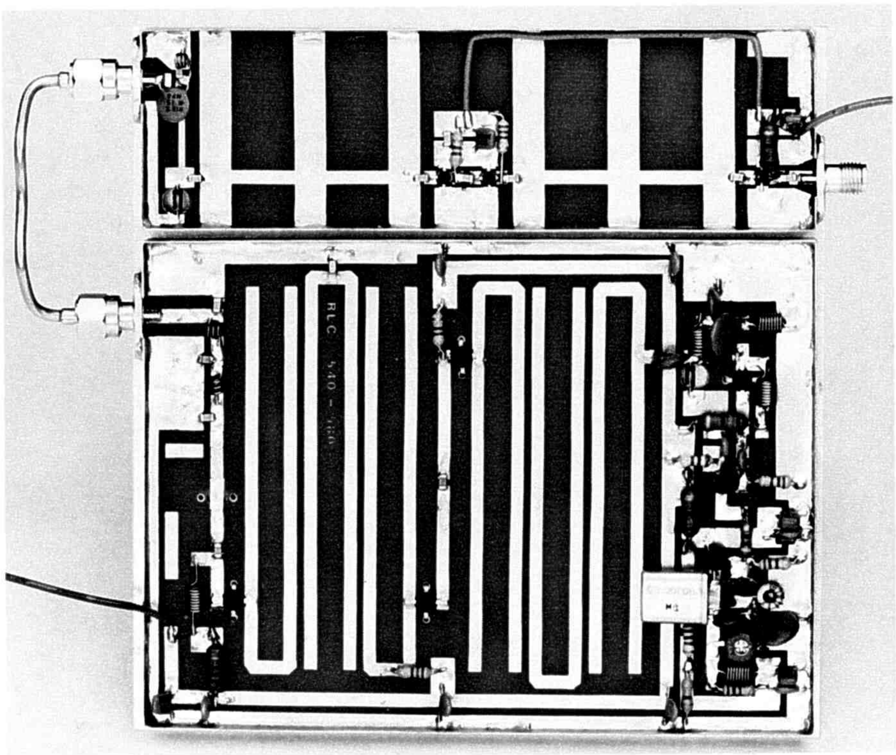
Contents

A Clean, Low-Cost Microwave Local Oscillator

By Rick Campbell, KK7B

(From *QST*, July 1989)

Obtaining a suitable local oscillator (LO) has traditionally been a major obstacle in amateur microwave work. This article describes a straightforward, inexpensive, easily constructed microwave LO. The oscillator may be combined with a simple low-noise preamp[1], an image filter[2] and an off-the-shelf doubly balanced mixer to build a complete high-performance receiving converter for OSCAR Mode S. This LO can also be used as a building block in a transverter for the 2304- or 3456-MHz bands.



The complete microwave LO is built on two PC boards. The larger (bottom) board provides a signal anywhere from 540 to 580 MHz, depending on the crystal frequency. The smaller board is a ×4 multiplier that provides an output from 2160 to 2320 MHz, depending on input frequency. Used separately or together, these boards have a wide variety of UHF and microwave applications.

All of the critical microwave circuitry in this LO is readily taken care of by a pair of fiberglass-epoxy (G-10) PC boards. The remaining parts include noncritical chip capacitors for interstage coupling and bypassing, standard ¼- and ½-W bias resistors, inexpensive, plastic-cased monolithic-microwave integrated-circuit (MMIC) amplifiers, a pair of 99-cent diodes, a few hand-wound inductors, disc-ceramic capacitors, and a 90-MHz, 5th-overtone crystal oscillator. PC board manufacturing tolerances, component variations, and construction tolerances have all been allowed for in the design. There are no RF tuning adjustments except for the 90-MHz oscillator tank circuit.

Design Goals

This project began with a list of design goals:

1) No tuning adjustments should be required.
2) All frequency-sensitive elements are printed on G-10 board.
3) Use inexpensive, readily available components.
4) Offer sufficient output to drive a standard-level mixer.
5) Use a single 12-V power supply.
6) All spurious outputs are more than 40 dB down.
7) Have electrical, mechanical and thermal stability consistent with portable CW operation on mountaintops in bad weather.

These goals have been met, with one minor exception: The 90-MHz crystal oscillator tank circuit must be tuned to make the oscillator start reliably. This adjustment can be made by listening for the crystal-oscillator output on an FM-broadcast radio. The electrical, mechanical and thermal stability are impressive. One of these LOs was still operating after an airline baggage-handling event left the aluminum transverter case so badly bent that the top had to be removed with a hammer!

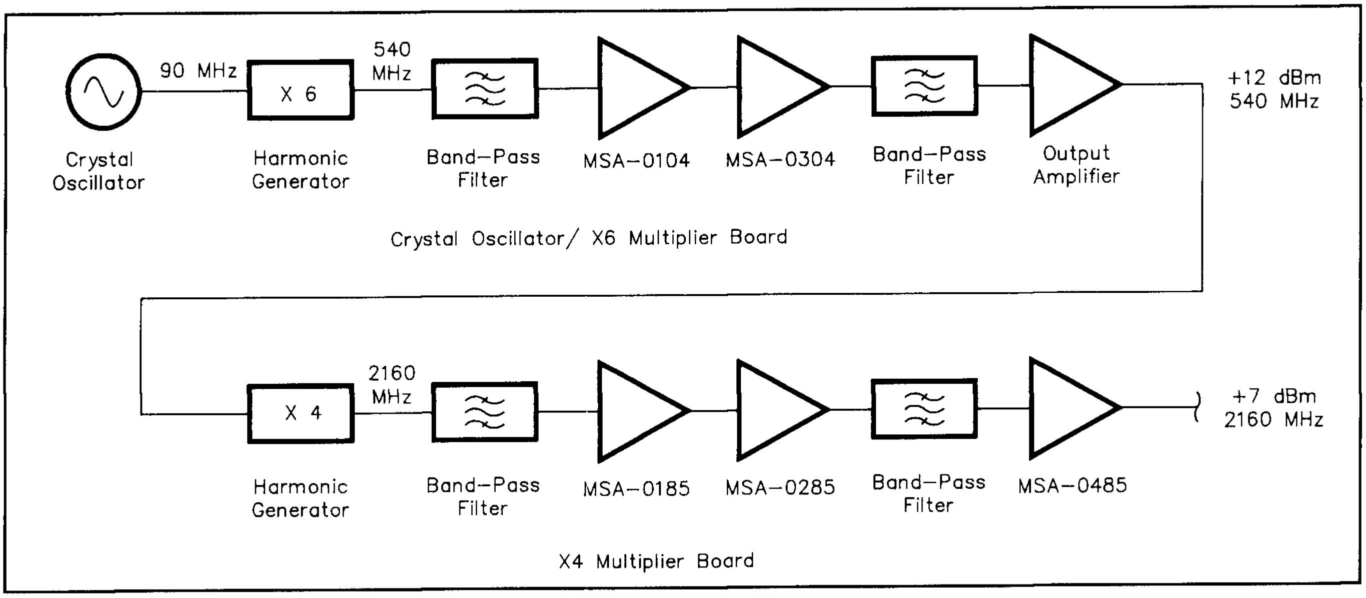


Fig 1—Block diagram of the 2160-MHz LO.

System Description

The complete microwave LO, shown in block-diagram form in Fig 1, consists of two PC boards: a crystal oscillator and times 6 (×6) multiplier board; and a ×4 multiplier board. The crystal oscillator/×6 multiplier board can generate any frequency between 540 and 580 MHz; simply choose the appropriate crystal. The output level depends on the device chosen for the output amplifier. An Avantek MSA-0404 is used for the output amplifier in the version described here. (See this article's Amplifiers section for more details.)

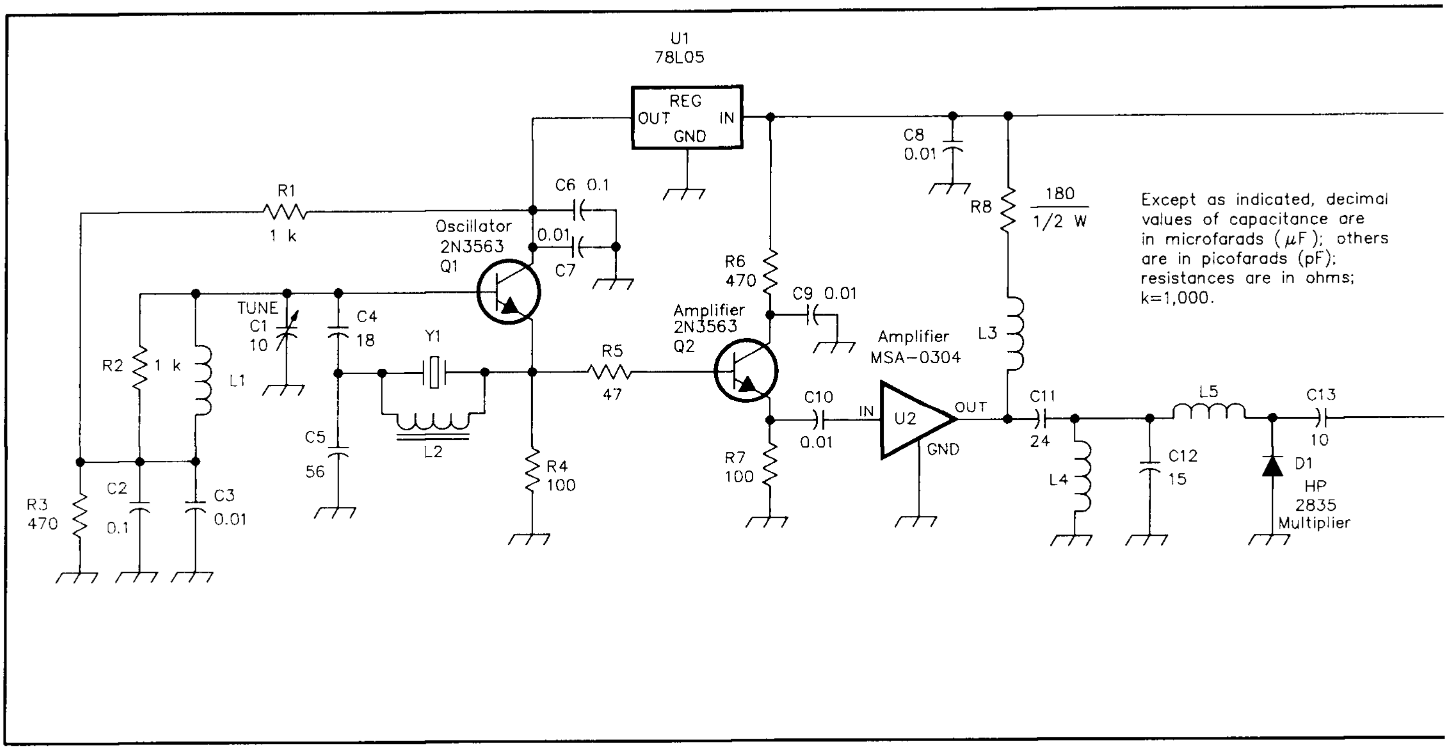


Fig 2—Schematic diagram of the crystal oscillator/×6 multiplier board. Resistors are ¼-W carbon-film types unless otherwise indicated. Capacitors are 50- or 100-V disc-ceramic types unless otherwise noted.

C1—8- to 10-pF trimmer capacitor. Ceramic-piston trimmer preferred; standard ceramic trimmer acceptable.

D1—Schottky diode; Hewlett-Packard 2835, 2800, 2811 or equiv. See text.

J1—SMA female chassis-mount connector preferred. See text.

FL1, FL2—Band-pass filters printed on PC board.

L1, L3, L4, L5, L6—8 turns no. 28 enam wire, 0.1 inch ID, closely wound.

L2—10 turns no. 32 enam wire on T-25-6 toroid core, or 0.33 µH miniature RF choke.

L7, L8—3 turns no. 28 enam wire, 0.0625-inch ID, spaced 1 wire diam.

The ×4 multiplier board can be used for any output frequency between 2140 and 2360 MHz. The harmonic-generator components are sufficiently broadly tuned that the board works equally well as a ×3 or ×5 multiplier. Any input level between +7 and +13 dBm is fine, and inputs as low as 0 dBm may be used, at reduced output levels.

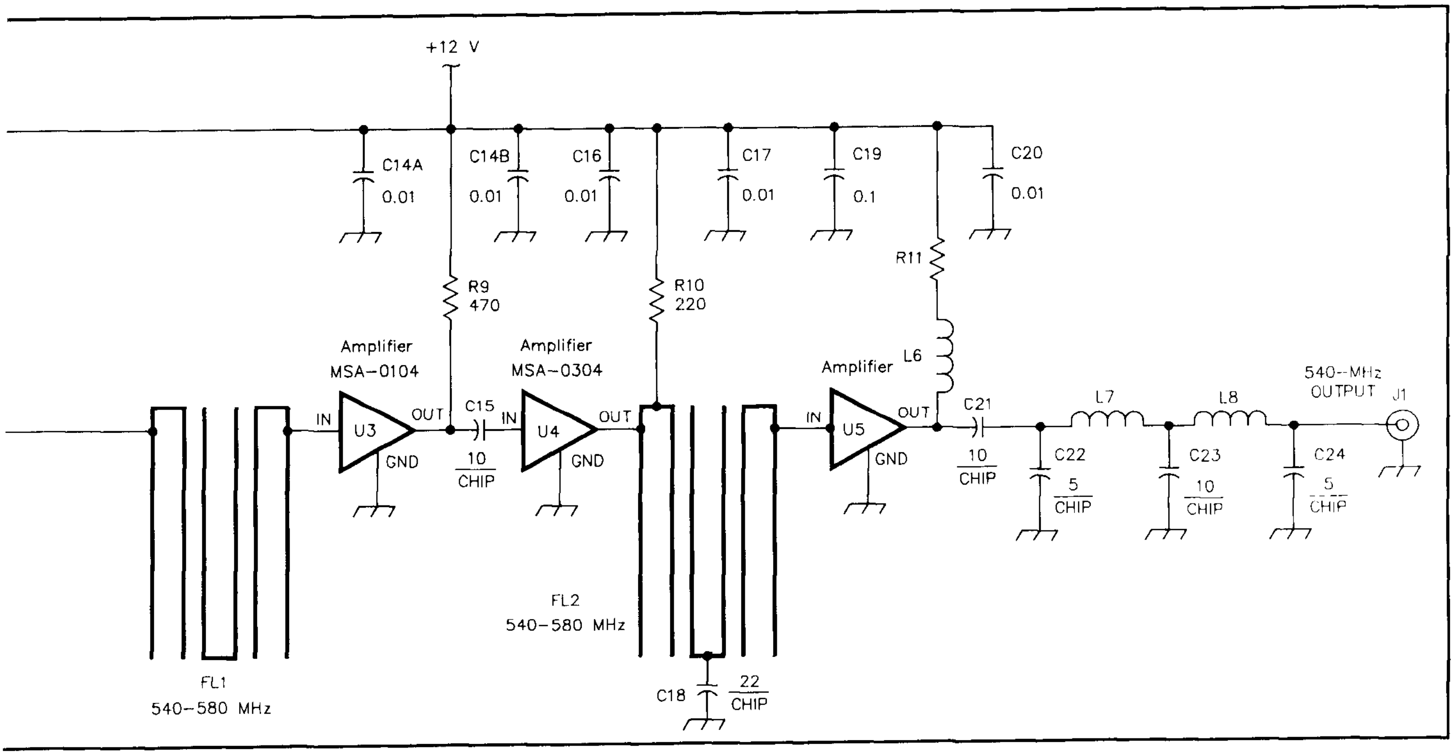
These two boards can be used independently—in fact, they were developed for two separate projects. The 540- to 580-MHz board was developed at the suggestion of Jim Davey, WA8NLC, who wanted a simple 552-MHz driver for his single-board 3456-MHz transverter.[3] The 2140- to 2360-MHz multiplier board was developed as part of a no-tune 2304-MHz transverter that was described in the *Proceedings of Microwave Update '88*.[4]

The local-oscillator system shown in Fig 1 has four functional blocks: the 5th-overtone crystal oscillator; the Schottky-diode harmonic generators; the printed band-pass filters; and the MMIC amplifiers. Each of these blocks is described in the following sections. A schematic of the crystal oscillator/×6 multiplier board is shown in Fig 2, and Fig 3 shows a schematic of the ×4 multiplier board.

Crystal Oscillator

The crystal oscillator generates the signal that is subsequently multiplied into the microwave region. In this design, the 90-MHz crystal-oscillator signal is multiplied by 24 to produce the final output signal (2160 MHz). Any long-term drift or "warblies" on the 90-MHz oscillator will be 24 times worse at the output frequency. Common crystal-oscillator circuits that work well in a 144-MHz, or even 432-MHz, receiving converter may be unacceptable when the output frequency is multiplied into the microwave region.

The Butler emitter-follower circuit shown here was originally suggested to me by Al Ward, WB5LUA, and modified to the present circuit by Jim Davey, WA8NLC. (I don't waste much time arguing with those two—when they express an opinion, they generally turn out to be right.) This oscillator will free-run on the tank-circuit frequency if the crystal and its shunt inductor (L2) are replaced with a 47-Ω resistor. This characteristic is especially useful if you want the tank circuit to operate at another frequency. After initially testing the prototype oscillator with inexpensive 2N5770 transistors, I tried replacing them with some 20-year-old pullout 2N3563s, some MPS3563s, a pair of 2N5179s that I found on the floor under the bench, and some new AT-42085 microwave transistors from Avantek. All of these devices worked in this circuit. I also discovered that the value of R1, which sets the operating points of Q1 and Q2, can be varied to change the power output. A 1-kΩ resistor was fine for all the transistors except the AT-42085s. The output power from the AT-42085s was about



Q1, Q2—2N3563, MPS3563, 2N5179 or equiv. See text.
R11—If U5 is an MSA-0404, use 120-Ω, ½-W resistor. If U5 is an MSA-1104, use a 100-Ω, ½-W resistor. See text.
U1—5-V, 100-mA, 3-terminal regulator.
U2—MSA-0304 MMIC preferred. MSA-0404, MSA-0385, MSA-0485, MAR-3 or MAR-4 also usable. See text.
U3—MSA-0104 MMIC preferred. MSA-0185, MSA-0685, MAR-1 or MAR-6 also usable. See text.
U4—MSA-0304 preferred. MSA-0285, MSA-0385 or MAR-2 also usable. See text.
U5—For +12 dBm out, use MSA-0404. For +16 dBm out, use MSA-1104. See text.
Y1—90-MHz, 5th-overtone, series-resonant crystal.

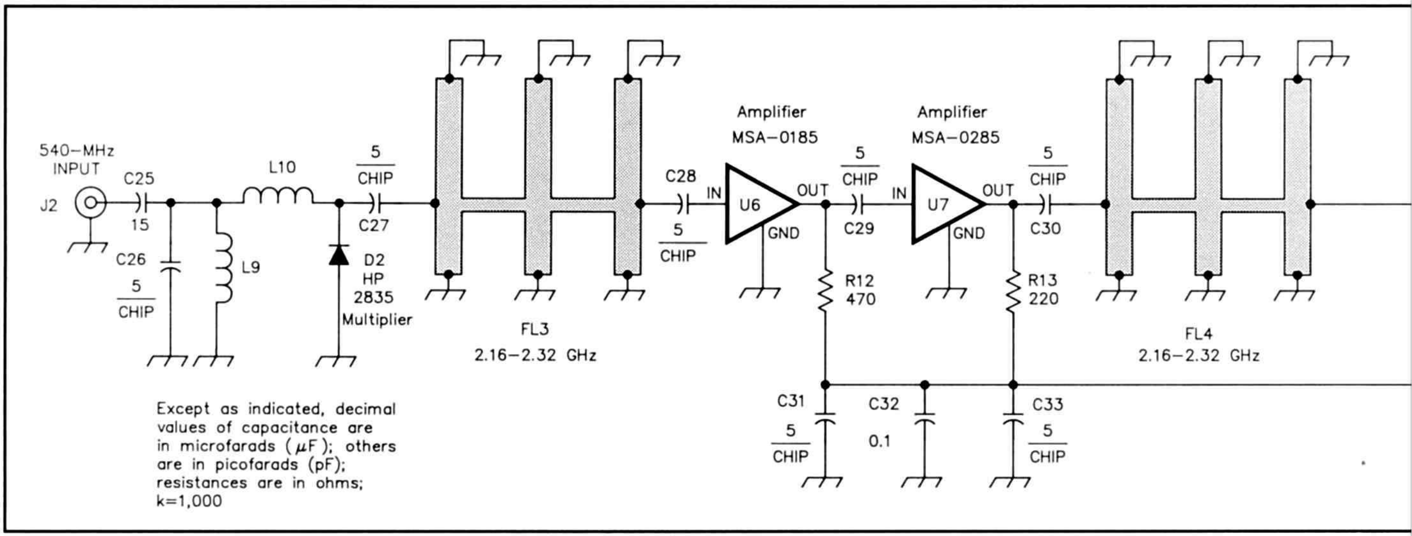


Fig 3—Schematic diagram of the ×4 multiplier board. Resistors are ¼-W carbon-film types unless otherwise indicated. Capacitors are 50- or 100-V disc-ceramic types unless otherwise noted.

D2—Schottky diode; Hewlett-Packard 2835, 2800, 2811 or equiv. See text.
J2, J3—SMA female chassis-mount connector preferred. See text.
FL3, FL4—Band-pass filters printed on PC board.
L9—3 turns no. 28 enam wire, 0.0625-inch ID, spaced 1 wire diam.
L10—Inductor printed on PC board.
U6—MSA-0185 or MAR-1 MMIC preferred. See text.
U7—MSA-0285 or MAR-2 MMIC preferred. See text.
U8—MSA-0485 or MAR-4 MMIC preferred. See text.

+6 dBm—a little too much drive for the MSA-0304 buffer (U2).

The LO shown in the photo on the first page of this article varies slightly from the schematic in Fig 2. The photo shows a Zener-diode regulator in Q1's collector circuit. When this board is used in a setup with a battery supply, the difference in voltage when switching from receive to transmit may be enough to cause an observable frequency shift. This problem is eliminated by using a 3-terminal, 5-V regulator (U1), as shown in Fig 2.

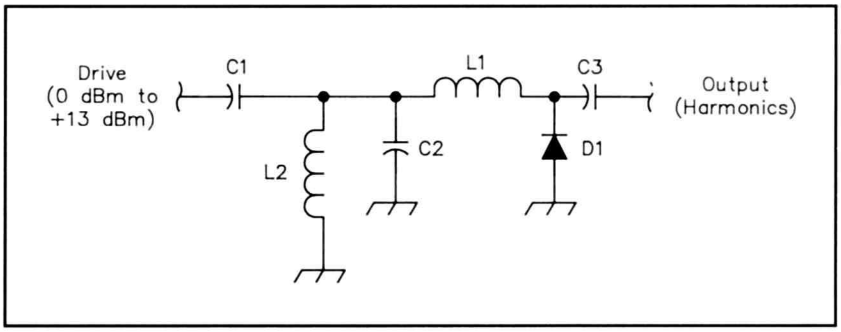


Fig 4—Schematic diagram of the basic harmonic generator used on both boards. See text for discussion.

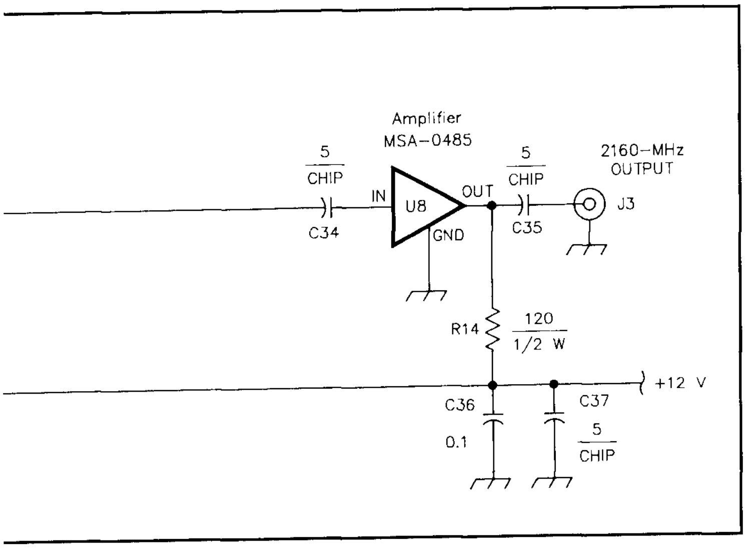
Harmonic Generator

Harmonic generation is easy—or, at least, not generating harmonics is very difficult. Solid-state power amplifiers must be low-pass filtered to get rid of the harmonics generated by the nonlinear characteristics of the transistors. In fact, if you do anything to a sine wave—clip it, drive a class-C amplifier with it, half-wave rectify it—anything that distorts its perfectly sinusoidal shape, the resultant signal will contain harmonics.

Harmonic generation is difficult only if you want to do it with high efficiency. High efficiency in microwave-LO harmonic-generator stages was important in the 1960s and 70s because amplifying a low-level microwave signal to the +7 dBm level required for many diode-ring mixers was expensive. Back then, tuning up efficient multipliers that used expensive step-recovery or varactor diodes required hours in front of a spectrum analyzer tweaking a handful of $5 piston trimmers to within a half turn of oblivion. And you had to do it all again if the drive level, load or temperature changed.

Now that unconditionally stable, broadband MMIC amplifiers are available for less than a dollar, multiplier efficiency is a minor consideration. By taking advantage of readily available modern components, we can build a broadband multiplier—with no RF-tuning adjustments—that is unconditionally stable with variations in temperature, load and drive level.

The harmonic generator shown in Fig 4 is just a half-wave rectifier with a simple low-pass filter (L1) feeding in the fundamental, and a simple high-pass filter (C3) picking off the harmonics. A half-wave rectifier based on an ideal diode generates only odd harmonics. A Schottky diode (D1 of Fig 4) has an offset voltage of a few hundred millivolts, so the conduction angle is less than 180 degrees. Odd and even harmonic levels are approximately equal for drive levels up to about +10 dBm. This basic harmonic generator is used on both boards. For higher drive levels, a bias circuit with



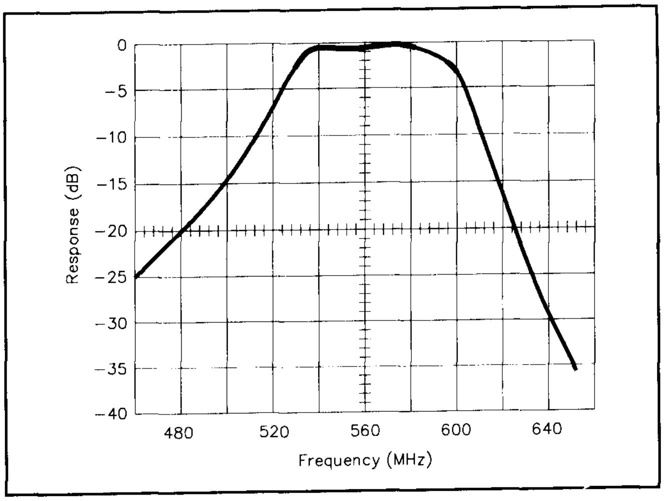


Fig 5—Frequency response of the 540- to 580-MHz hairpin filters printed on the crystal oscillator/×6 multiplier board.

a trimmer potentiometer is recommended.[5]

Other diodes will work in this circuit. Schottky diodes like the HP-2800 and HP-2811 are suitable. A silicon switching diode like the 1N4148 works fine, and even provides more output than the specified Schottky diode, but it has one major disadvantage: It may oscillate! If you don't believe this, compare the circuit of Fig 4 with that of a parametric amplifier circuit in an early VHF manual. Better yet, connect the input of a 1N4148 multiplier to a signal generator and the output to a spectrum analyzer and try it. At some combination of input frequency and drive level, the output noise floor will rise considerably, and the output spectrum will contain many discrete output signals—typically, subharmonics of the drive signal modulating the desired output. This circuit needs to operate reliably from a motorcycle battery on a mountaintop in a rainstorm, so the use of switching diodes is discouraged.

Filters

The filter selects the desired harmonic output from the harmonic generator. In the past, amateur-built frequency multipliers usually were limited to multiplication factors of 2 or 3, because of the difficulty of tuning to the correct harmonic. With fixed-tuned filters having steep skirts and flat tops, it is easy to build multipliers of much higher order. The theory behind hairpin filters (FL1 and FL2 in the 540- to 580-MHz board) and off-center-tapped half-wave filters (FL3 and FL4 in the ×4 multiplier board) is covered in the amateur and professional literature.[6-9] Only the practical aspects are mentioned here.

The primary design goals for these filters were low cost and reproducibility without tuning adjustments. To achieve the first goal, I've specified G-10 board. As a result, the filters are more lossy than equivalent designs on a more expensive substrate, such as Teflon fiberglass. The loss for each of the three-element sections used here is about 3 dB. Because a 10-dB-gain MMIC capable of compensating for this loss costs only about $1, and because better substrates may cost $100 a square foot, G-10 is an attractive trade-off.

To achieve reproducibility without tuning adjustments, the filters are made broadband (with lots of low-Q resonators), rather than narrowband (with only a few high-Q resonators). The resulting passband characteristic, shown in Fig 5, looks more like that of an SSB filter in an HF rig than something out of an LO chain! This filter characteristic is fundamentally different from a single-tuned circuit that must be tuned exactly on frequency—a signal anywhere in the 40-MHz-wide passband (540 to 580 MHz) passes through, but the undesired 5th and 7th harmonics 90 MHz above and below the passband are greatly attenuated by the steep filter skirts.

There are two major advantages to using flat-passband filters in a crystal-controlled LO. One is that the frequency may be moved anywhere in the passband by simply changing the crystal. The other is that allowance for manufacturing tolerances and variations in circuit-board material can be designed in before the circuit is built, rather than having to be tuned out afterward. For example, if the PC-board manufacturer is a little sloppy and the production boards are 1% smaller than the prototype, the passband will be 5 MHz higher. The desired signal will still get through, and the undesired signals will still be attenuated.

Even a change in circuit-board material from G-10 to its fire-retardant variant, FR-4 (which has a slightly different dielectric constant), results only in a well-behaved upward shift in the passband. The FR-4 boards provide a few decibels more output at 576 MHz, and the G-10 boards are a few decibels better at 540 MHz. The passband shape is the same for both G-10 and FR-4 materials.

In an attempt to discover how tolerant the 540- to 580-MHz hairpin filters are, I reduced the length of several hairpin resonators by one millimeter. The output dropped about 2 dB, worst case, and the spurious outputs remained acceptably low. I also sprayed a complete 2160-MHz LO with a thick coat of clear Krylon, and could not detect any change in the module's output. These filters work better than double-tuned circuits; require no tuning; are tolerant of manufacturing, device and construction variations; allow a range of frequencies to be generated from a single board layout; and cost no more than the PC-board material required to make them.

Amplifiers

The MMIC amplifiers serve two important functions: They raise the level of the desired harmonic, and they provide broadband, 50-Ω interstage terminations. Many suitable amplifiers are available, and substitutions require only a little care concerning output level. The MMICs most popular with hams are available from Avantek and Mini-Circuits Labs. The Avantek line has part numbers starting with MSA-; the MSA-0404 is an example. Mini-Circuits parts numbers start with MAR-; the MAR-4 is an example. The devices used here were chosen because they were closest to the top of the pile of parts on my workbench. Drive to D1 should be about +10 dBm—an MSA-0304 or '0404 works well for U2. The desired harmonic coming out of FL1 is at about –17 dBm, so a device with a 50-Ω input and more than 10 dB of gain is useful for U3. The MSA-0104, MSA-0185, MSA-0685 and MAR-6 are good choices.

The second interstage amplifier, U4, must deliver about +10 dBm and provide a 50-Ω load for FL2. MSA-0304, MSA-0285 and MAR-3 devices have all been used in this stage. The final stage, U5, may be omitted if +6 dBm is enough output. If U5 is not used, also omit R11, L6, C20 and C21. A good plan, however, is to use a high-power device that will not saturate, and then add a resistive attenuator to reduce the output to the desired level. An MSA-0404 will provide +12 dBm output, and an MSA-1104 works well for +16 dBm out.

It's not really necessary to operate any of these amplifiers linearly, but the output stage should be kept at or below its 1-dB compression point, or the spurious outputs will rise to an unacceptable level. MMIC selection is left up to the builder. Just about anything will work—and who knows what will be available next year for 49 cents? If you'd rather not think about which parts to use, just use those specified in the schematics. If you use devices different from those specified, be sure to use an appropriate bias resistor, as specified in the manufacturers' data sheets.

Construction

Both boards are etched on 0.062-inch-thick, double-clad, G-10 PC board material. All components are surface mounted on the etched side; the other side is unetched and acts as a ground plane. The components necessary to complete this project are available from several suppliers, and etched PC boards and complete parts kits are available as well.[10]

The first construction step for the crystal oscillator/×6 multiplier board shown in Fig 6 is to drill the component holes and file the edges flat. The filter elements need not be grounded, so the exact size of the board is not critical. The edges of the upper and lower ground foils must be connected all the way around the board, though. A plated-through hole or shorting wire every half inch or so works fine. Copper tape wrapped the whole length of each edge and soldered top and bottom also works well. The method I recommend is to solder brass or tin walls all the way around the circuit board, making sure to solder to both the top and bottom of the circuit board. This results in a nice, rigid box with solderable shield walls suitable for mounting feedthrough capacitors and the output connector.

The next step is to add the bridges across the unused breaks on the crystal oscillator/×6 multiplier circuit board (between C21 and C22), and a piece of copper tape to short the cold end of multiplier diode D1 to the ground plane. The MMIC ground leads are bent at right angles to the device body, passed through holes drilled in the board, and soldered directly to the

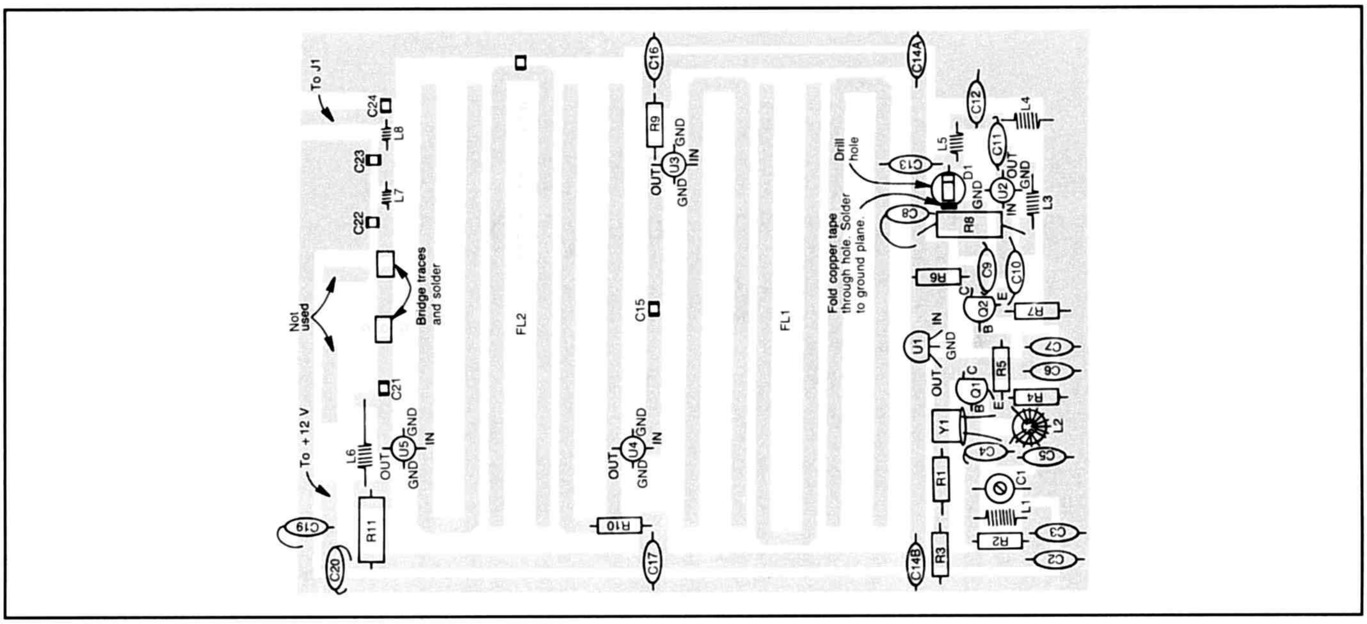


Fig 6—Part-placement diagram for the crystal oscillator/×6 multiplier board (not shown actual size). All components mount on the etched side of the board.

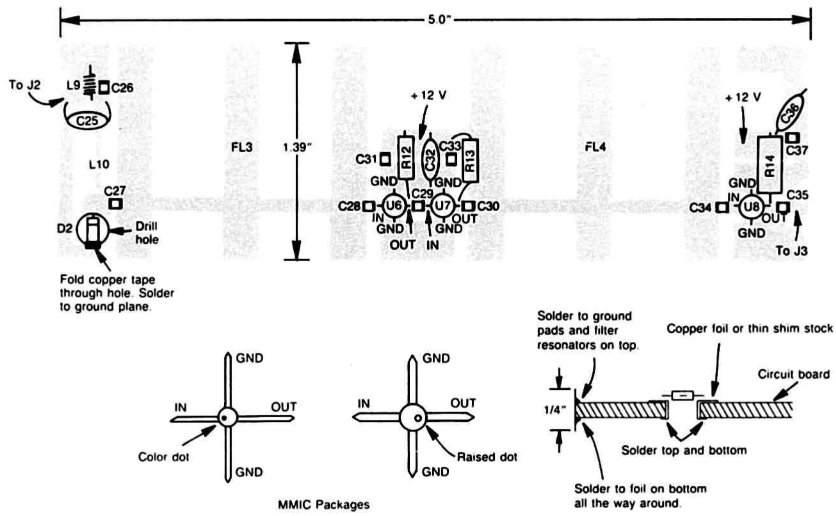


Fig 7—Part-placement diagram for the ×4 multiplier board (not shown actual size). All components mount on the etched side of the board.

ground plane. The 5-V regulator's ground lead also goes through the board and is soldered to the ground plane. I usually put the chip capacitors on first, followed by the inductors, resistors, disc-ceramic capacitors, diode, transistors, 5-V regulator IC and MMICs, in that order. The crystal can be installed as shown on the parts-placement diagram by sticking it to the board with a small piece of double-sided foam tape.

I prefer to use SMA end-launch connectors, even at the output of the crystal oscillator/×6 multiplier board,

because they are small and easy to use. (I also use an SMA connector at the output of the ×4 multiplier board because it's an excellent microwave connector. By standardizing on connectors, I don't have to keep switching adapters on my power meter.)

After you've carefully checked all your mounted components against the parts-placement diagram, you can apply 12 V to the board. Tune C1 until you hear the 90-MHz signal in a nearby FM broadcast radio. Then turn the power supply on and off a few times to make sure the oscillator starts reliably. The crystal oscillator/×6 multiplier board is now complete.

The filter topology on the ×4 multiplier board (Fig 7) differs from that of the ungrounded hairpins on the crystal oscillator/×6 multiplier board. The width of the ×4 multiplier board determines the resonant frequency of the shorted half-wave filter elements. The correct length for all the half-wave resonators is obtained by cutting the circuit board precisely to the width shown in the drawing, and then soldering the board's brass wall to the ground plane on the bottom, and to the end of each resonator on the top of the circuit board. Plated-through holes, ground wires and copper tape wrapped around the board edges will not work with this layout.

I obtain the correct board dimensions by scribing a line on the top of the circuit board at exactly the correct place. Then I cut the board slightly oversize. Next, I lay a large, flat file on my workbench and work the circuit board back and forth until the board edge is filed to the scribed line. This results in a nice square edge as shown in Fig 7. Only the width of the ×4 multiplier board is critical. Because FL3 and FL4 form a bandpass filter with a flat passband response, construction errors of up to about 0.032 inch do not significantly affect the output level at 2160 MHz.

After soldering the side and end walls to the ×4 multiplier board, add copper tape to ground the MMICs and multiplier diode as shown in Fig 7. Then add the chip capacitors, disc-ceramic capacitors, inductor, diode, MMICs, bias resistors and SMA connectors. No adjustments need be made to the ×4 multiplier board

Performance

Fig 8 shows the output spectrum of the crystal oscillator/×6 multiplier on G-10 board with a 90-MHz crystal and an MSA- 0404 output device. The plot is from dc to 1 GHz. The largest spur, at 450 MHz, is 70 dB below the +12-dBm, 540-MHz output. The 360-MHz and 630-MHz spurs are just barely visible at about 75 dB below the 540-MHz output. The harmonics at 1.08 GHz and 1.62 GHz (not shown) are more than 55 dB below the 540-MHz output, and are not measurable because of the limited dynamic range of this spectrum analyzer. This is a clean LO!

Different crystal frequencies result in different spurious-output levels. Worst-case spurious outputs are about –45 dB for any output between 540 and 580 MHz.

Once the board is built and tested, frequency stability can be enhanced considerably by thermally insulating the crystal oscillator. I usually tape a small piece of sponge packing foam over the oscillator and then package the entire system in a box to keep rain and cold mountain breezes out.

Fig 9 shows the output spectrum of the ×4 multiplier board

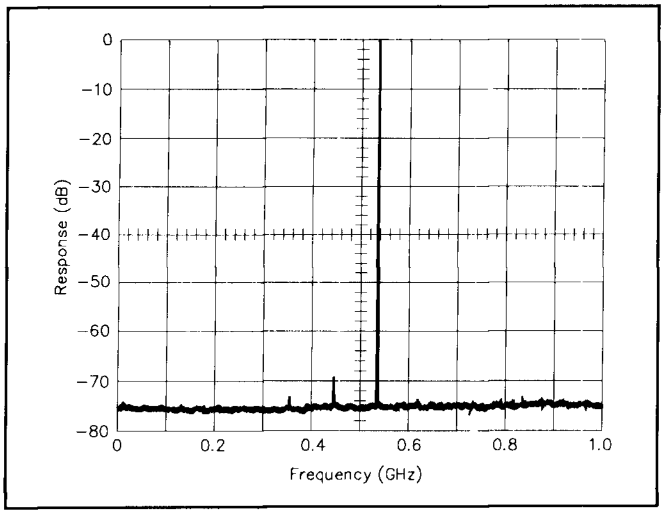


Fig 8—Output spectrum of the crystal oscillator/×6 multiplier board. The desired signal is +12 dBm at 540 MHz.

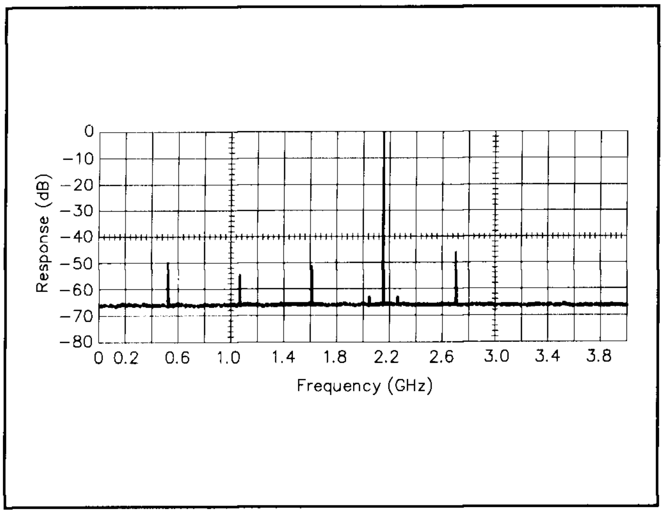


Fig 9—Output spectrum of the ×4 multiplier board. The desired signal is +8 dBm at 2160 MHz.

driven with the signal shown in Fig 8. The plot covers dc to 3.7 GHz. The largest spur, at 2.70 GHz, is 45 dB below the +8-dBm, 2.16-GHz output. The lower frequency spurs, at 0.54, 1.08 and 1.62 GHz, are more than 50 dB down. The second harmonic, at 4.320 GHz (not shown), is only about 25 dB down. However, harmonic spurs are not too important on LO outputs, since the mixer generates harmonics of the LO signal anyway. If the output of the ×4 multiplier board is used to drive an antenna or another multiplier stage, then a filter (as described in reference 5) may be added.

A word about LO drive level is in order here. Many engineers, both amateur and mercenary, agonize because they have a +7-dBm mixer and only +6 dBm of LO drive. It's true that 1-mW (0 dBm) is probably not enough drive for a 5-mW (+7-dBm) mixer, and that 100 mW (+20 dBm) is too much—but there is some latitude. A few decibels either way won't make any measurable difference in most systems. If you really don't want to get on the air, having only +5-dBm drive for your +7-dBm mixer is as good an excuse as any. But the guy down the street with 10 countries worked via moonbounce is probably running +3 dBm into the RF port of an unknown surplus mixer.

Applications

These two boards have been used for a surprising number of applications. I use the 2160-MHz LO described here in a pair of 2304-MHz transverters with 144-MHz IFs, and in a 3456-MHz transverter with a 1296-MHz IF. Simply change the crystal to 96 MHz, and you have a low-power CW transmitter for 2304 MHz. A 94-MHz crystal will provide a 2256-MHz LO for OSCAR Mode S. The ×4 multiplier board may be used with a suitably modified 70-cm FM exciter to generate 2304-MHz FM (460.8 × 5 = 2304 MHz).

The crystal oscillator/×6 multiplier board is even more versatile. With a 96-MHz crystal providing 576-MHz output, it serves as the LO for my single-board 1296-MHz transverter. It may be used with a ×4 multiplier for 2304-MHz CW output, a ×6 multiplier for 3456-MHz output, a ×10 multiplier for 5760-MHz output or a ×18 multiplier for 10.368-GHz output. If an untuned Schottky diode multiplier is used, it will provide useful signal levels for receiver and filter alignment on the calling frequencies of all the amateur bands from 2.3 through 10 GHz. With 92-MHz crystal and an MSA-1104 providing +16 dBm at 552 MHz, it can serve as the LO board for the WA8NLC single-board 3456-MHz transverter described in June 1989 *QST.*

Acknowledgments

This work would not have been possible without the forum for amateur microwave technology exchange provided by Don Hilliard, WØPW, in the form of the Microwave Update Conferences. The speakers and attendees at Microwave Updates '85 through '88 provided much of the basic information and all of my motivation for pursuing this work. In particular, I thank Al Ward, WB5LUA, for the wealth of information on MMICs he has made available to the amateur community, and Jim Davey, WA8NLC, for bringing us hairpin filters and encouraging me to keep pushing the state of the art.

Notes

[1]A. Ward, "Simple Low-Noise Microwave Preamplifiers," *QST*, May 1989, pp 31-36.

[2]R. Campbell, "2.3 GHz Transverters," *Proceedings of Microwave Update '88* (Newington: ARRL, 1988), pp 9-32.

[3]J. Davey, "A No-Tune Transverter for 3456 MHz," *QST*, Jun 1989, pp 21-26. *Proceedings of the 21st Conference of the Central States VHF Society* (Newington: ARRL, 1987), pp 51-57. (Reprinted in this book.)

[4]See note 2.

[5]R. Campbell, "A Clean Microwave Local Oscillator," published in *Proceedings of the 1296 and 2304 Conference*, Estes Park, CO, Sep 1985. Reprinted in *Proceedings of the 21st Conference of the Central States VHF Society.* Newington: ARRL, 1987), pp 51-57.

[6]J. Wong, "Microstrip Tapped-Line Filter Design," *IEEE Transactions on Microwave Theory and Techniques*, Vol MTT-27, No. 1, Jan 1979, pp 45-51.
[7]G. Beebe, "Analysis of a Class of Microstrip Bandpass Filters," MSEE thesis, Michigan Technological University, Feb 1988.
[8]J. Davey, "Microstrip Bandpass Filters," *Proceedings of Microwave Update '87* (Newington: ARRL, 1987), pp 42-53.
[9]J. Davey, "Microwave Filter Update," *Proceedings of Microwave Update '88*, pp 1-8. See note 2.
[10]Most of the parts for this project are available from Microwave Components of Michigan. Etched PC boards and parts kits are available from Down East Microwave.

Phase Lock Control Circuit For Use With Brick Type Oscillators

By C. L. Houghton, WB6IGP and Kerry Bane, N6IZW

This project was started to overcome the inherent stability problems in the California Microwave and Frequency West "brick" type oscillators. We have found that the basic brick oscillator's crystal (operates in the 100-MHz range) can vary 1 kHz or more (short term stability). This system if left unchecked can vary in output frequency up to 40 kHz or more (at 10 GHz) when left on for long-term operation. This drift is due to the brick's internal 100-MHz oscillator temperature and aging characteristics.

Several modifications to the brick crystal oscillator have improved the basic stability of this high-frequency control oscillator. These include: (1) Better temperature control circuitry for the crystal housing, and (2) Changing some of the capacitors in the oscillator to temperature coefficient types, which improved overall stability. This improved system stability, but the improvements still missed the mark. Something several magnitudes better was sorely needed.

We designed the phase-lock control circuit to improve upon existing stability problems looking for a high state of accuracy. What we developed was a circuit that controls the brick's crystal oscillator, by phase locking this crystal to a high accuracy external reference source (Fig 1). This improved the basic brick oscillator to what we feel is the ultimate, with very simple circuitry. Part of the design goal was to minimize modifications to the brick oscillator. The phase-lock circuitry would be external to the brick, and would control the 100-MHz oscillator by comparing it to a very high-accuracy standard. In practice, we were able to stabilize the crystal oscillator to about 100 Hz at X band.

Kerry, N6IZW, did the initial work and came up with the design while testing two different prototypes. The first test circuit used two separate high stability low frequency oscillators to divide the high frequency brick oscillator control crystal (a fractional value) down to a common PLL frequency. This system worked well, but with two different oscillators the circuit was cumbersome and might be difficult for others to duplicate. After quite a bit of number crunching on a computer Kerry discovered that a single 10-MHz standard could be used in the system eliminating the second oscillator.

The original design was modified adding an additional 74HC163 programmable divider and providing pinouts for any other possible divide by ratio to be programmed in. Our system used a single 10-MHz high stability oscillator as the system reference, although other frequency combinations are possible. We have included a program that was used in our system allowing other oscillator combinations to be compared to your brick oscillator frequency. First a little about our system.

The 10-GHz frequency that we have standardized on for SSB work is 10.368 GHz. The local oscillator system uses a 10.223 GHz (Brick Frequency) for a lower side mix incorporating our Frequency West Brick oscillators. The fractional crystal frequency for a Frequency West brick operating at 10.223 GHz is 100.2254902 MHz. This fractional frequency is quite a challenge to phase lock but I believe you will like the scheme used to do the job.

The single oscillator approach mixed the tenth harmonic of the 10-MHz oscillator with the 100.225-MHz signal from the brick oscillator leaving a difference of 0.2254902 MHz. The computer program (attached) will give the relationship between this number and 10 MHz. In this case it was found that dividing 10 MHz by 1020 and 0.2254902 MHz by 23 produces a very close match to 0.01 Hz. Note other combinations are possible but this was the simplest choice. This new frequency (9.803913043 kHz) primary reference is fed to a 4046 phase detector along with the crystals divided down input. The output of the 4046 PLL controls a varactor located in the brick's crystal adjust circuitry trimming the crystal to exactly the proper frequency.

System error using this system of phase-lock control results in errors of less than 1/100th of a cycle at 100 MHz, (actually it's 0.008526 Hz.). Using the system, the 100.2254902-MHz brick oscillator will lock up in about 2 minutes after applying power up the circuit and brick. The circuit that was finalized is quite simple in keeping with the original design goals. The harmonic mixer is nothing more than a 2N918 with its collector circuit tuned to the low frequency product (0.2254902 MHz) in this example. The output of the mixer is coupled into a VN10KM VMOS FET, amplified and squared up in two sections of a 74HC04.

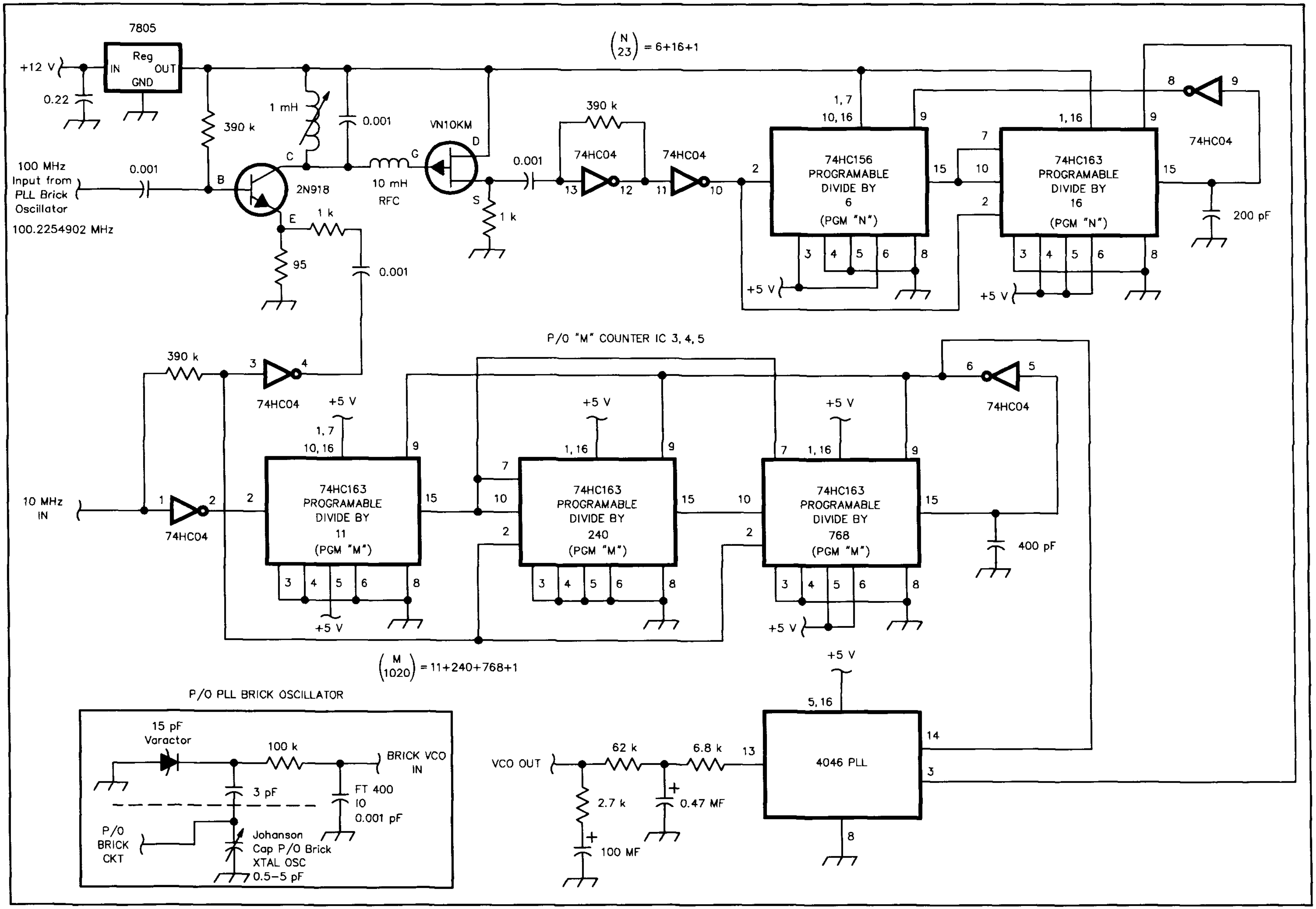


Fig 1—10-GHz phase-locked "brick" oscillator stability circuit designed by N6IZW and WB6IGP.

This signal is then coupled into a programmable divide by a string of 74HC163s. Note that all programming pins are brought out, allowing custom programming for different division ratios. The 10-MHz oscillator input is squared up in a 74HC04 inverter and feeds two signal paths. One is the harmonic mixer and the other is the programmable M divider for the master reference. The 4046 PLL compares the output of the M and N dividers and adjusts the control voltage for the varactor in the brick. This varactor is in series with a small 3-pF capacitor which is tied in parallel to the existing Johanson capacitor in the Frequency West brick.

The test results of this system, using identical phase-locked circuits, at either end of the test path, were excellent. From a cold start, as soon as the high-stability main-reference oscillators were stabilized, we attempted communications on 10-GHz SSB. We selected 14.01 MHz for both SSB IFs. The RIT control was the only adjustment needed to clarify SSB operations at both ends of the test path, as both signals were in each other's passband on 2 meters. The error was about 100 Hz.

An off-shoot modification to the basic high-stability circuit is just the opposite but can be very useful. Our 2-meter radios are SSB hand-helds. Frequency setting is done by small BCD switches. Searching for other stations proved cumbersome. The solution to the problem was to disable the high-stability circuitry. In its place, we installed a pot to vary the voltage on the brick's varactor. This gave us about 100 kHz of frequency tuning (at 10 GHz). Fine control was obtained by using a 10-turn pot. This gave us a method to search and scan quite a bit of frequency in short order, instead of flicking BCD switches. Once you verify what frequency the other station is using, you can return to high stability and let the other station adjust to your frequency.

Perspective

The use of high-stability oscillators brings greater opportunities for weak-signal work. If you know where a signal is without question, you can start some interesting applications. One such possibility is to copy a repetitive signal being transmitted over a very long path that normally is not received. With special techniques, it is possible to reach under the noise floor and pull these signals out. Very high-stability oscillators and computers are necessary to further these techniques.

Brick Oscillator Number Generation Program

The program listed in Table 1 generates the proper division ratio for other combinations of brick oscillator frequencies and reference frequency input. Additionally, the program will define the M and N counter division numbers and give the data bits 0 or 1 to be preset into each counter. Program step 190 is set at 1 to start division comparisons. This number can be raised if you find that higher order numbers are needed. In our case we set up to 500 or so for most work with a 10-MHz input reference in step 190.

We hope you will adapt this circuit to your phase-locked oscillators, as the improvement in stability is quite noticeable. After all, 100-Hz stability at 10 GHz with simple circuitry is something to brag about!

Table 1
BASIC Program Listing

```
10  DEFDBL A-Z
20  DEFINT I
30  DIM P(13)
40  REM THIS PROGRAM CALCULATES THE DIVIDE
     BY "M" AND "N" COUNTER
50  INPUT "ENTER BRICK FREQUENCY",BF
60  INPUT "ENTER REFERENCE OSCILLATOR
     FREQUENCY",RF
70  REM RH IS HARMONIC NUMBER BEING COM-
    PARED
80  REM HD IS DIFFERENCE BETWEEN RH AND BF
90  REM M IS THE NUMBER BY WHICH REFERENCE IS
     DIVIDED
100 REM N IS THE NUMBER BY WHICH DIFFERENCE
     IS DIVIDED
110 REM S & U SET PRECISION OF DIVISION MATCH
120 RH=CINT(BF/RF)
130 PRINT "BRICK OSC FREQ= ";BF:"MHZ"
140 PRINT "REFERENCE OSC FREQ= ";RF;"MHZ"
150 HD=ABS(BF-RF*RH)
160 PRINT "DIFF FREQ IS ";HD;"MHZ"
170 IF RF*RH>BF THEN PRINT "PHASE SEN REV"
180 PRINT " M ";" N"; " CALC DIFF FREQ IN MHZ"
190 M=1
200 N=M*ABS(HD)/RF
210 S=SBS(RF*INT(N)/M-HD)
220 IF S<.00001 THEN PRINT M; INT(N);RF*INT(N)/M
230 IF S<.00001 THEN 300
240 U=ABS(RF*INT(N+1)/M-HD
250 IF U<.00001 THEN PRINT M;INT(N+1);RF*INT(N+1)/
     M
260 IF U<.00001 THEN GOTO 295
270 M=M+1
280 GOTO 200
290 REM THIS CALC M CNTR PRESET FOR M=1
295 N=INT(N)+1
300 D=2048
301 PRINT
305 M=M-1
320 FOR I=12 TO 1 STEP -1
330 IF M<D GOTO 360
340 M=M-D
350 P(I)=0 GOTO 370
360 P(I)=1
370 D=D/2
380 NEXT I
385 PRINT "M CNTR PRESET LSB ";
390 FOR I= 1 TO 12
400 PRINT P(I);
410 NEXT I
480 PRINT
485 REM THIS CALC N CNTR PRESETS FOR N-1 CNT
490 D=128
495 N=INT(N)-1
500 FOR I=8 TO 1 STEP -1
510 IF N<D THEN GOTO 540
520 N=N-D
530 P(I)=0 : GOTO 570
540 P(I)=1
570 D=D/2
580 NEXT I
```

Frequency Multipliers Using Silicon MMICs

By Jim Davey, WA8NLC

Introduction

Two recent articles describe the use of MMICs as frequency doublers.[1,2] The advantages cited are high efficiency and unconditional stability. MMICs can also be used for higher order multiplication. In low-power applications, the MMIC multiplier offers several advantages over Schottky-diode multipliers including lower drive power and less current for the total oscillator system.

Background

My work with no-tune transverters for 2304 and 3456 MHz[3] required the use of ×4 and ×6 multiplication on the board. This was done by driving a Schottky diode with a high-level source to generate harmonics. Filters and MMIC amplifiers were added to select the desired harmonic and amplify the output to a usable level. The success of MMIC doublers cited in the references intrigued me. How would a MMIC multiplier perform in a ×4 or ×6 stage? Would the MMIC offer any advantage over the Schottky diode?

This paper describes a series of experiments I performed with silicon MMICs to see what could be done in a ×4 and higher mode. The experiments were performed on local oscillators for 2160 and 3312 MHz. The results show that MMICs work well at multiplication orders up through at least ×6, with less drive power than the Schottky diode. Efficiency was enhanced by selecting an optimum MMIC-to-filter spacing.

Test Methodology

Experimental data was obtained by following this general procedure:

1. Measured the relative strengths of the harmonics in the output of the MMIC.

2. Varied the drive level to see over what range of input power the MMIC delivered the highest harmonic levels.

3. Added an output filter to the MMIC to select one of the higher harmonics. Checked for instabilities resulting from the VSWR of the filter.

4. Using one harmonic (6th), varied the device bias current to see what correlation it had to the level of the 6th harmonic.

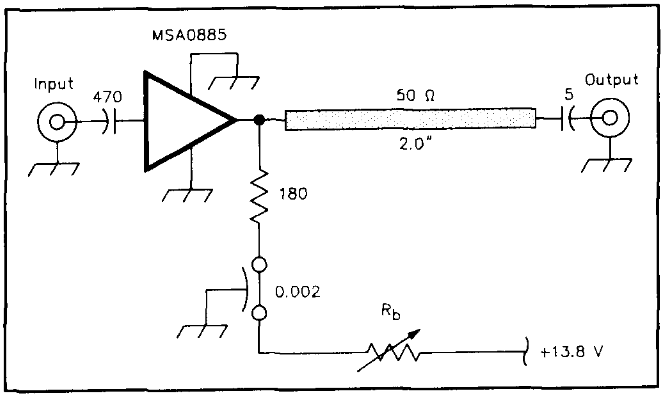


Fig 1—Schematic of the 540-MHz test fixture.

5. Varied the electrical distance between the MMIC and the filter to see if the "filter gain" phenomenon also worked for active devices. Recorded output level data for several values of bias at each spacing.

Test Details

I chose to limit the test to multiples of 540 MHz, since I had an extra 540 MHz no-tune oscillator available that could deliver +14 dBm. For the active device, I used an Avantek MSA 0885 (Fig 1). The gain of the 0885 is 26 dB at 540 MHz with 1-dB compression at an output level of +14 dBm. Saturation should begin at an input of −12 dBm.

The MMIC was mounted on a small 50-Ω test fixture made of 0.031-inch TFE-glass board, to allow bias and dc isolation capacitors to be added. SMA connectors were edge-soldered to the microstrip for input and output connections. The output was measured on an HP 8551B spectrum analyzer and HP 431B power meter.

In the first tests I looked at the effects of drive and bias on the 4th and higher harmonics. Initially, a wide range of input levels was tried to see what input level the MMIC preferred for best harmonic output and spectral purity. At the same time, the device bias was varied to test the effect of bias current at each level of input power. In general, for each combination of input power and harmonic number, I found a unique bias condition that gave the best output. The bias used in this test were lower than the specification value to prevent excessive current in the MSA 0885 when driven at +10 dBm.

The practical upper limit for input drive appears to be about +10 dBm. Above that point no additional output power was observed with any level of bias. Overall efficiency dropped as drive levels approached +10 dBm, too. Nearly as much output was available at lower drive levels if the bias was adjusted to optimize the output level. For this reason I decided to conduct the remaining experiments at a lower input level of 0 dBm. This level is about 12 dB higher than the level required to cause +l dB compression of the fundamental in the MMIC.

Next, I filtered the output and tried to optimize the 4th and 6th harmonic. Microstrip bandpass filters were attached to the test fixture and the drive and bias were varied while watching for greatest output. The optimum drive and bias conditions found in these tests did not correlate well with the unfiltered test data. The reasons for this were apparent after the next test was performed.

"Filter Gain" Test

A Hewlett-Packard application note[4] describes some tests HP made on a step-recovery diode multiplier/filter setup. It showed that each harmonic can be peaked by locating the diode at specific electrical distances from the reflective-type filter. At these locations the reflected energies in the other harmonics are returned to the diode in proper phase to convert more of the unwanted harmonic power into power at the desired harmonic. Of particular interest was the fact that the optimized filtered output level at any chosen harmonic was higher than the unfiltered level in the broadband output, the added power coming from the other harmonics. If this principle also applied to the active MMIC multiplier, higher efficiency may be achievable than the unfiltered output would suggest.

To test this principle, I chose the 6th harmonic at 3312 MHz and constructed a test fixture of a 3312 bandpass filter (10% BW, 1.5 dB loss) preceded by a long 50-Ω input line. The input line was made over 360 electrical degrees long, to be able to see the whole range of phase conditions. The line was marked off in steps of 0.1 inch, which corresponded to about 15 degrees on this dielectric. The 50 Ω line was driven by an Avantek MSA 0885.

Data was then taken at four levels of bias from 12 mA to 29 mA at each distance to the filter, down to 15 degrees. The data was then plotted, revealing that for each bias condition, there are spacings that definitely enhance the efficiency of the multiplier. When plotted on top of each other, it was evident that for most bias conditions, 210 to 230 degrees gave the highest output. I could not correlate these results to the HP application note because HP's tests were performed on 8th and higher harmonics. I believe, however, that the effect is adequately demonstrated for active multipliers.

The one variable I held constant in these tests was the input power level. It is possible that combinations of input power and bias could be found that would give even greater outputs. In addition, the type of filter used may have an effect on the overall results. The hairpin filter used in these tests looks like an open circuit to the highest level signal in the output, the fundamental at 540 MHz. If another type of filter is used that appears as a short circuit, such as an interdigital, the phase of the reflection will be altered and different results would be expected.

The "filter gain" effect is seen in a comparison of the data. The level of the 6th harmonic in the unfiltered test was –24 dBm for an input of 0 dBm. Under the same bias conditions, the level of the 6th harmonic was –13 dBm when passed through a bandpass filter located 220 degrees away. The apparent "gain" is 11 dB. This is in the range of the HP test results.

Comparison to Schottky-Diode Multiplier

The Schottky-diode multiplier layout in Fig 2 is the one used in my 3456-MHz transverter. In that design, the diode spacing to the filter is between 15 and 30 degrees. Diode bias and input matching are optimized for maximum output. Total conversion loss for the diode is 29 dB.

Applying the results of the above test, the multiplier chain shown in Fig 3 could be realized. Signal levels are kept below 0 dBm until the last amplifier. Shielding and isolation require-

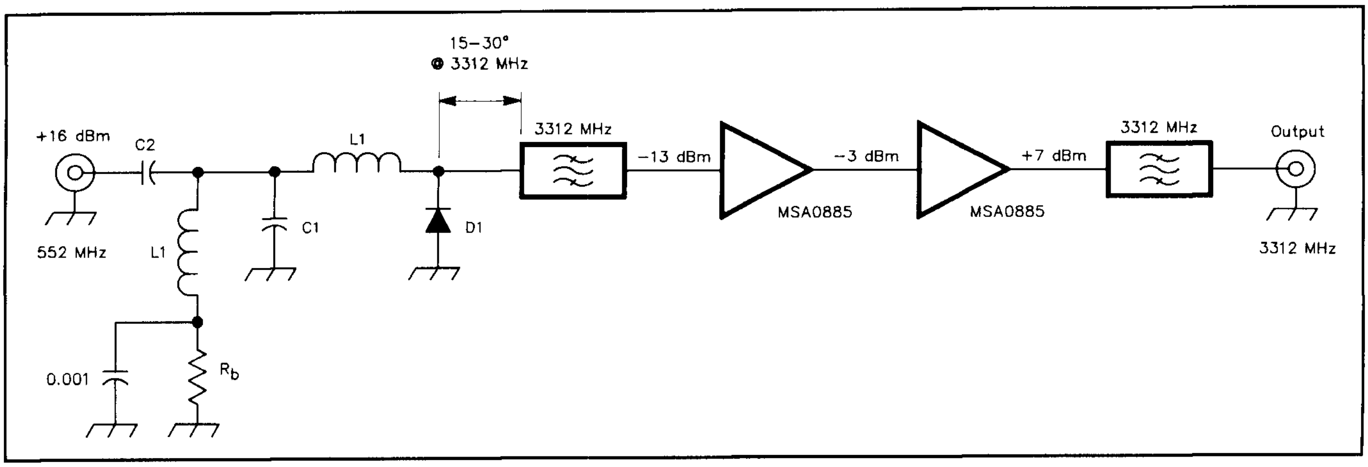


Fig 2—Block diagram of a Schottky-diode multiplier for use in a 3456-MHz transverter. L1-C1 and L2-C2 resonate at 552 MHz. D1 is the Schottky diode.

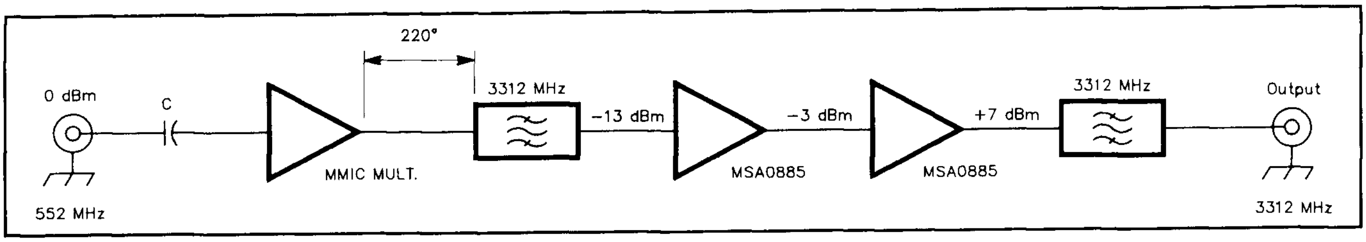


Fig 3—Block diagram of a multiplier for the same application, but using a MMIC instead of a Schottky diode.

ments would be greatly reduced with this design.

Although not usually an important design consideration in amateur equipment, the total current required in the MMIC version is lower. The parts count is about the same when the MMICs bias and dc blocking components are considered.

Summary

Simple frequency multipliers can be made with inexpensive silicon MMICs well into the microwave range that offer an alternate to Schottky diode multipliers. Drive requirements are lower with the MMIC design. The principle of "filter gain" can be effectively used to achieve maximum efficiency in MMIC multipliers. For each set of operating conditions, there is an optimum MMIC-to-filter spacing for best results. Experiment!

Notes

[1]J. Hinshaw, "MMIC Multiplier Chains for the 902-MHz Band," *Ham Radio*, Feb 1987, pp 72-79.
[2]J. Hinshaw, "MMIC Active Multipliers," *RF Design*, June, 1988, pp 64-68.
[3]J. Davey, "A No-Tune Transverter for 3456 MHz", *QST*, June 1989, pp 21-26.
[4]"Comb Generator Simplifies Multiplier Design," *Hewlett-Packard Application Note 983*, July 1981.

Weak-Signal Sources for the Microwave Bands

By Paul Wade, N1BWT

Introduction

There are a few areas of the country where microwave experimenters have the luxury of beacons or local hams to provide a signal source for checking out microwave equipment. The rest of us need a local weak-signal source for tuneup, and to be able to verify that things are working when no signals are heard. Relying on harmonics of VHF equipment can be very misleading—I learned this the hard way on top of Burke Mountain in northern Vermont during the 1990 UHF contest.

These weak-signal sources provide a local signal source that is strong enough to be heard at a reasonable distance, perhaps a few hundred feet, but not strong enough to ride through directly into the receiver, so the entire system, including preamps, coax switches, feed line, and antenna, may be verified. The crystal frequency can be chosen so that none of the harmonics fall in the IF band; most of those listed in the table on the schematic avoid all ham bands.

The design is based on the no-tune transverter[1] and oscillator[2] work of KK7B, and is very similar to the LO section of the 903 transverter. The only significant differences are in scaling of the filters for 903, 1152, and 1296 MHz; also, I flipped the artwork to avoid bending the base leads of common transistors under the can. Fig 1 is a schematic of the source.

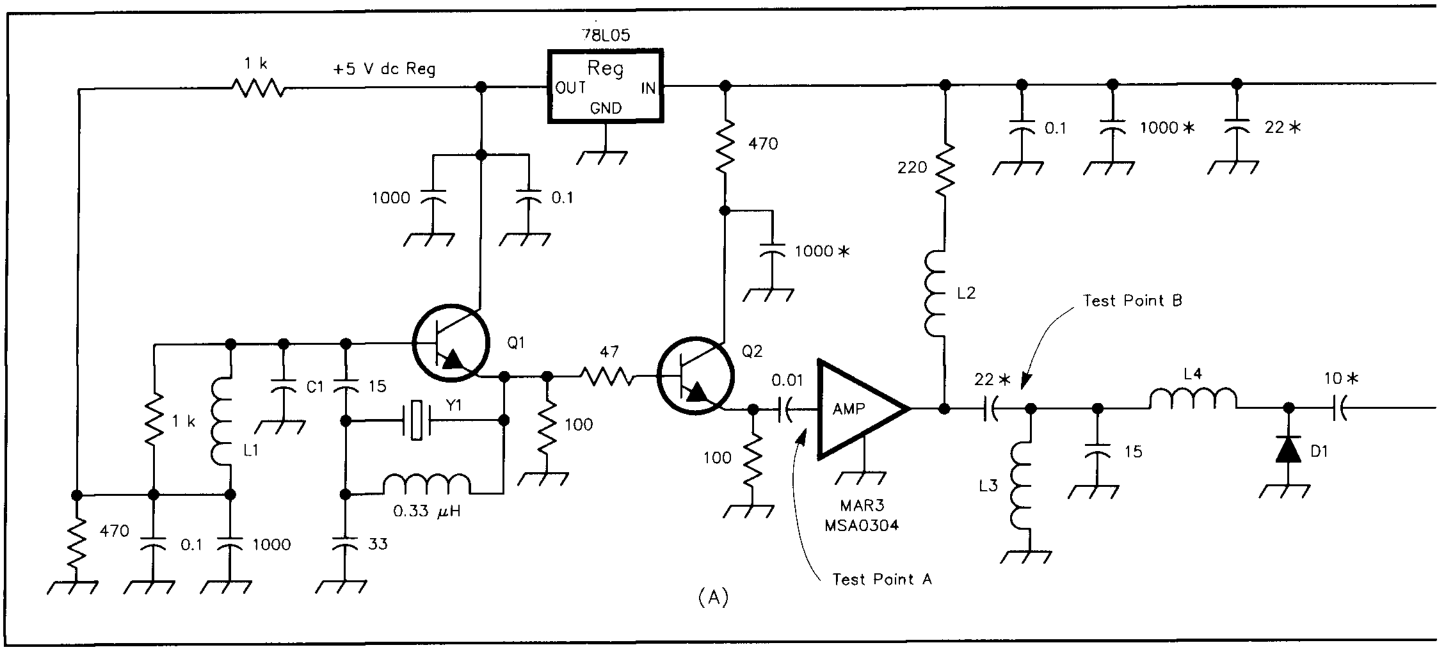


Fig 1—Schematic diagram of the N1BWT weak-signal source. The inset at B is an optional amplifier. See text. Capacitors marked * are chip capacitors. The source can be built in a Radio Shack 270-238 enclosure.

C1—10 pF trimmer.
D1—Schottky diode, Hewlett-Packard HP2835 or equiv.
L1—Below 90 MHz: 11 turns no. 24 enam wire, 0.1-inch dia. Above 90 MHz: Same, except 10 turns.
L2-L5—8 turns no. 28 enam wire, 0.0625-inch ID, spaced 1 wire diam.
Q1—2N5179, 2N4124 or equiv.
Y1—Crystal; see text and Table 1 for details.

Tuneup

Tuneup is straightforward, and easiest if the source is built in stages, following the schematic from left to right. Most of the component values aren't critical—the ones mentioned below require some attention. The first step is to install all parts up to Test Point A, and measure the output at the test point. Adjust trimmer C1 for max output, checking to see that the crystal is running on the right overtone. Most of the crystal frequencies can be checked by listening on an FM broadcast radio (if it isn't near the right frequency, replace the crystal with a 47-Ω resister and adjust L1 until C1 tunes through the desired frequency). I don't like to leave trimmers in the circuit, so after peaking C1, I estimate its capacitance and replace it with a fixed capacitor, then squeeze and stretch L1 to repeak the output. The second stage is to add components up to Test Point B and check the output there. I typically see one milowatt (0 dBm) at Test Point A and 10+ milliwatts (10 to 12 dBm) at Test Point B.

Now it is time to add the multiplier and next two amplifiers. There isn't enough output after the multiplier to measure unless you have a spectrum analyzer (that's cheating!), so the amplifiers give enough signal to measure, typically −10 to −20 dBm, depending on the number of times the crystal frequency is multiplied. This is plenty of output for a weak-signal source, but if you feel a need to peak it, squeezing and stretching L4 sometimes makes a difference.

The optional amplifier is not recommended for 903 and 1296, but may be useful at 1152 to get to the higher bands. To install, drill a hole in the output stripline for the MMIC and make a cut with a knife for the output blocking cap. With this amplifier there is enough gain in the box to make things touchy; an external amp might be better.

Table 1
Suitable Crystal Frequencies for Weak-Signal Sources

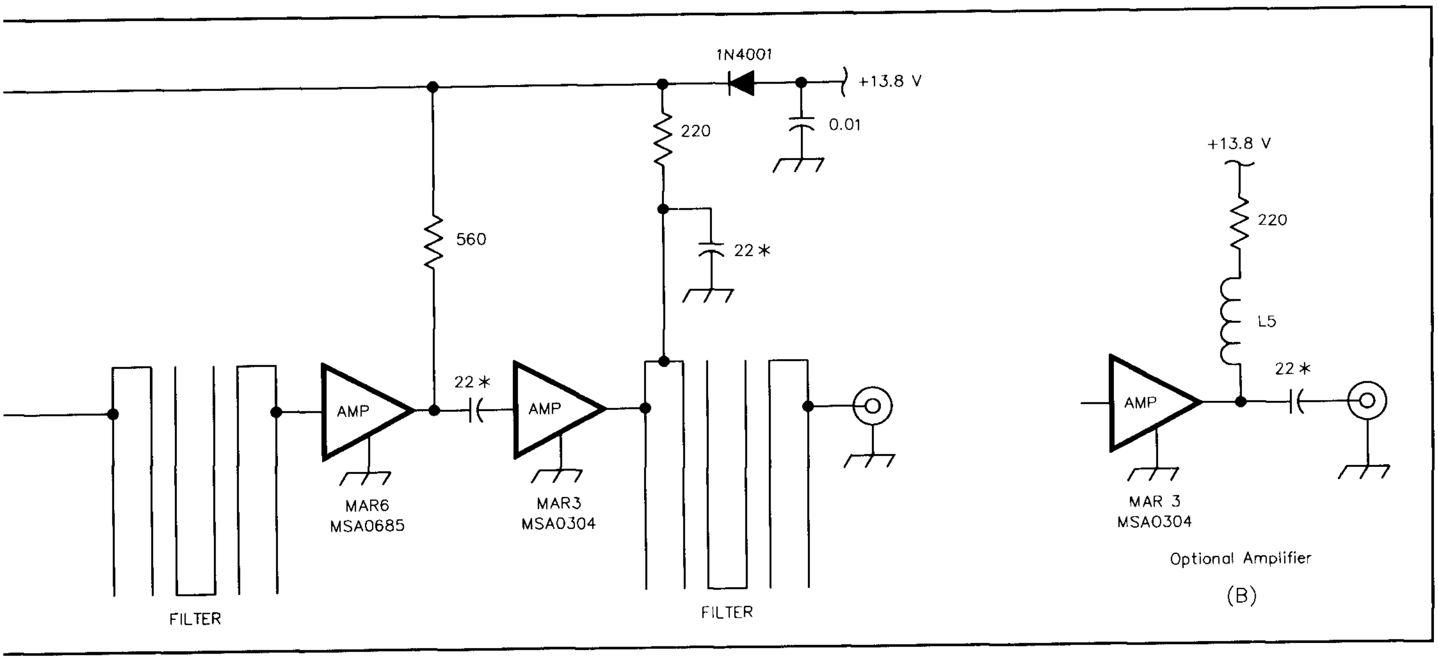
Output Frequencies (MHz)		
903.1	*1152*	*1296.1*
72.26	76.802	81.01
82.1	82.287	86.41
90.31	88.617	92.58
100.344	96.002	99.70
112.89	104.73	108.01

Board

Boards and kits are available from Down East Microwave.

Higher Bands

All the standard calling frequencies above 2 GHz are multiples of 1152 MHz (+100 kHz), so this is the starting point. I made this one with the optional amplifier, which provides measurable (about −30 dBm) harmonic output at 2304 and 3456 MHz to use as a weak signal, preferably after filtering. Output at 5760 and 10368 was not detectable, so I used a very simple harmonic generator and high-pass filter: a waveguide diode detector and a waveguide-to-coax adapter, both surplus, bolted together (Fig 2). The 1152-MHz output



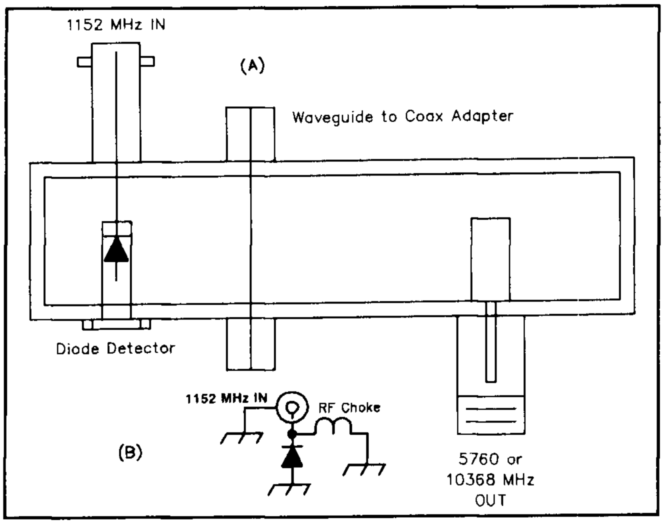


Fig 2—Waveguide harmonic generator for use with weak-signal source. At A, mechanical construction. B shows a schematic diagram of the harmonic generator, including an external RF choke required for use at low drive levels.

drives into the detector BNC output, and 10 GHz comes out the coax end of the waveguide to coax adapter. Drive of 8 to 10 dBm (from an external MMIC amplifier) produces an output of around –50 dBm at 10.368 GHz, but 0-dBm drive produced no output; apparently this is not enough to produce harmonics from a 1N23 diode. The waveguide section does rudimentary filtering, with all harmonics in the waveguide passband appearing; a real filter could be inserted here.

I haven't tried this with C-band waveguide components, but there's no reason it shouldn't work just as well for 5760 MHz.

Notes

[1] R. Campbell, KK7B, and D. Hillard, WØPW, "A Single Board 900-MHz Transverter with Printed Bandpass Filters," *Proceedings of Microwave Update 1989*, ARRL, 1989, p 1.

[2] R. Campbell, KK7B, "A Clean Low-Cost Microwave Local Oscillator," *QST*, July 1989, p 15.

Phase-Locked Microwave Sources

By Greg McIntire, AA5C

Introduction

The many phase-locked microwave sources circulating in surplus have been a real boon to microwave enthusiasts, particularly those building equipment in the 3456, 5760, and 10368-MHz bands. Their stable operation greatly simplifies the frequency variable when making microwave DX shots. Building around a good source simplifies much when working with microwaves.

This article briefly describes the principles of operation of these sources, and provides information and procedures that should allow you to get your LO up and running.

Principles of Operation

Most phase-locked microwave sources are based on the block diagrams shown in Figs 1 and 2, although there are many variations of the same theme. A microwave cavity oscillator serves as the fundamental RF oscillator for the source. The output of this oscillator is either the direct RF output or it drives a multiplier block, from which the output RF is obtained. A comb of harmonics extending into the frequency range of the fundamental oscillator is generated from a 5th- or 7th-overtone reference crystal oscillator. The fundamental oscillator is then phase locked to a harmonic of the reference oscillator.

The phase difference between the reference harmonics and the fundamental signal is obtained in a phase detector, basically a mixer. When the fundamental and one of the reference harmonics are equal in frequency, a pure dc level is obtained from the phase detector. An ac signal rides atop the dc level when the signals differ in frequency. The output of the

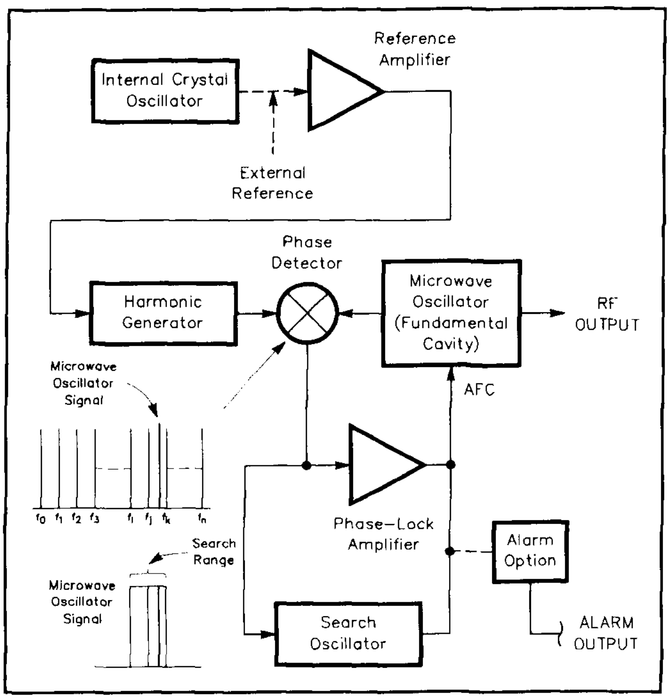


Fig 1—Block diagram of a phase-locked, mechanically tunable microwave signal source.

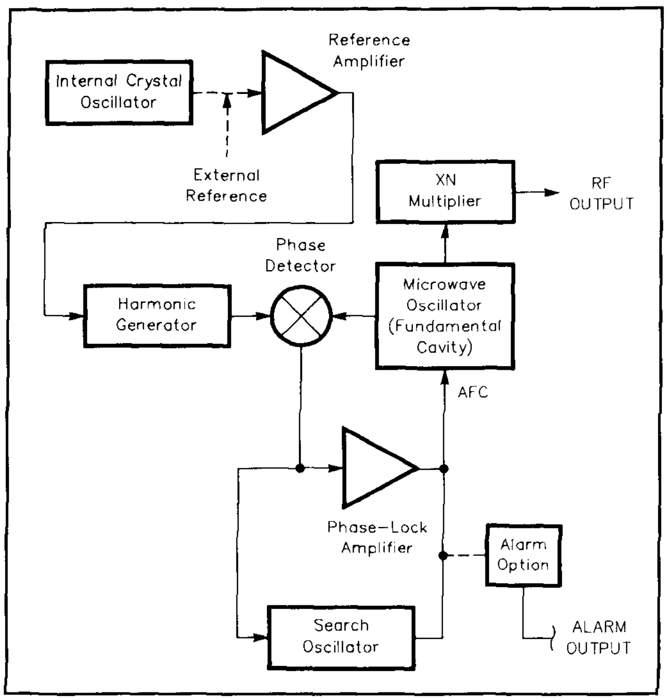


Fig 2—Block diagram of a phase-locked, mechanically tunable cavity-plus-multiplier microwave signal source.

phase detector is automatic-frequency control (AFC) of the fundamental cavity, typically via a varactor diode inside the cavity.

A search oscillator is used to initially acquire phase lock. The search oscillator slews the fundamental oscillator frequency, via AFC, so that it sweeps across the reference harmonics. Once lock is obtained, the search circuit is disabled The search oscillator automatically kicks in if lock is broken on most units Some require manual initiation of a new search.

The multiplier block, when used, consists of a step recovery diode (SRD) and an interdigital filter. A relatively high level signal from the fundamental cavity, typically +27 dBm, hits the SRD and generates the comb of harmonics. The interdigital filter selects the desired harmonic.

The main advantages of this type of source include:

1. High spectral purity output
2. High rejection of reference sidebands, typically 120 dB
3. Long term stability is that of the reference oscillator

Models

I'm aware of at least three companies that manufacture the more common "brick" sources for the 3-18 GHz range: Communications Techniques Inc. (CTI), Frequency West, and California Microwave. Sources from these companies are based on the same scheme and most are mechanically and electrically interchangeable. Non-"brick" sources I have worked with include 6- and 11-GHz units from Collins (Rockwell International). Almost all of the units operate from negative power supplies, common in the telecommunications industry.

The CTI, Frequency West, and California Microwave sources use a fundamental cavity operating in the 1-2 GHz range. A fifth overtone crystal in the 90-120 MHz range is the reference. All of the units I've seen have multiplier blocks. Depending on the fundamental cavity range, the sources can be used directly without multipliers for 1152 and 2160 MHz high level LOs (most have threads matching a female chassis mount BNC connector). A representative drawing of interconnections and adjustment points for a Frequency-West source, is shown in Fig 3. Options and variations I've come across include:

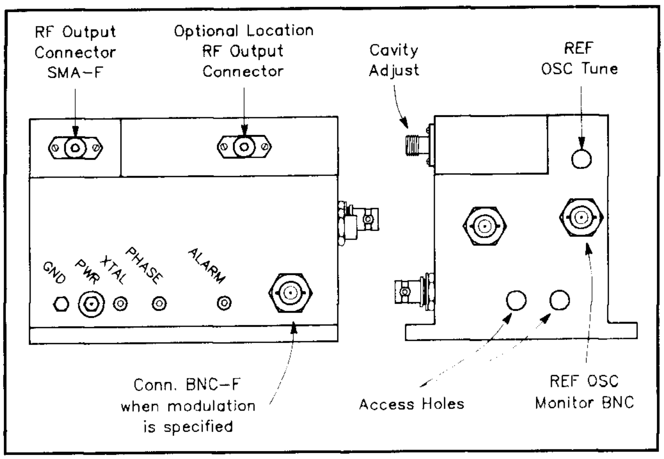


Fig 3—Locations of connectors and adjustment points on a typical Frequency West "brick."

- Output power level: 5, 10, or 20 mW
- Output connector: SMA, TNC or WR-75 output (WR-75 on some 10-15 GHz sources)
- Alarm (Y/N) and output type: relay contacts on older units, an open emitter transistor on newer ones
- Modulator input (Y/N): typically via a BNC-F connector
- "IF" output for external phase-lock control circuit
- External reference: reference oscillator not in basic unit
- Supply polarity: positive or negative (generally −19 or −20 V dc)
- Precision reference: the reference oscillator is phase locked to a precision reference, usually 20 MHz
- Multiplier: from 3 to 17 depending on manufacturer and frequency

Collins designed and built its own 6- and 10-GHz phase-locked sources. The Collins sources both use a fundamental cavity operating on the intended RF output frequency, ie, no multipliers are used.

The fundamental cavity in the Collins 6-GHz units tunes 5.9-6.4 GHz without modification. Units with P/N 622-5829-001 have two output SMA connectors, a tuning control and a bolt on the cavity. Units with P/N 622-5828-001 have a single SMA output connector, a tuning control, and no bolt on the cavity. Both 6-GHz units operate from +12 and −12 V dc, and have the following characteristics:

- Supplies: +12 V @ 110 mA, red lead
−12 V @ 450 mA cold, 180 mA warm
Common is black lead
- Output: +7 dBm via SMA connector
- Alarm: open emitter 2N2907, gray lead, on indicates unlocked
- Phase lock: blue lead, −6 Vdc indicates unlock—settles near 0.1 Vdc when locked
- Crystal: fifth overtone, 100-116 MHz
- Multiplier scheme: Fo=(F × Xtal × M)−25 MHz(M=25, 26, 27, 28 . . . so output falls within cavity range)
- Reference Output: SMA, 2× crystal frequency

A 3579.545-kHz color-burst crystal is multiplied by seven to provide the 25-MHz offset oscillator. I presume this was used for some sort of subcarrier, but have not investigated further.

The Collins 10-GHz sources were designed for the high end of the 10-GHz or the low end of 11-GHz range. I've put them on 10,224 and 10,368 MHz successfully, but the cavity output power is dropping off at the lower frequencies, which makes phase-locking more difficult. A tank-type reference oscillator is phase-locked to 4 times the reference crystal frequency. The reference oscillator is phase locked to the fundamental cavity. The Collins part number for these sources is 622-2983-002 and characteristics include:

- Power Supply: −20 Vdc @ 1.2 A cold and 0.8 A warm
- Output: +7 to +10 dBm via SMA connector
- Alarms: Cavity—yellow lead; reference—green lead; both open emitter 2N2907

• Phase lock: white lead, 0 to –18 Vdc swept during power up

• Sweep button: manual initiation of sweep when unlocked

• Reference indicator LED: on indicates unlocked reference

• Oscillator indicator LED: on indicates unlocked fundamental

• Crystal: fifth overtone 105-120 MHz

• Multiplier scheme: (crystal ×4)×22, 23, 24 . . . in range of cavity oscillator

• Reference output: SMA connector, 4× crystal

• RF monitor port: coupled output, SMA connector

• Test points: a) crystal oscillator level (peak for –4 to –5 V, b) reference AFC (tune for about –3 to –5 V with reference LED off), 3) oscillator RF power (should read –0.4 to –0.6 V) paralleled with brown lead, 4) oscillator AFC (same as phase-lock indicator)

Figuring Out What You Have

Visual Check

The first thing you should tweak when you get a source is NOTHING! A little bit of investigation before you do anything can save a lot of time later. Get any model number and frequency information from the source. For the brick sources, the model number should be on a label atop the multiplier block. It's pretty obvious if this label has been removed. If intact, you can generally look up the model number to get the multiplication factor, crystal range, and power level. The output frequency and crystal frequency are often also on the unit.

Generally, you'll find that the fundamental cavity of a Frequency West brick tunes about 300 MHz. You can multiply this up to determine potential output tuning range. The range of the multiplier block generally limits the tuning range.

Operational Check

Next, remove the side cover and see if there is a reference crystal in the unit. If so, get the crystal frequency. Now apply regulated power, usually –19 or –20 V dc. First verify that you have output RF. Check the output level and frequency (if you can measure the RF frequency directly). If you can't measure the RF frequency directly, you'll absolutely need to know the multiplication factor.

The pickup probe on the multiplier block fits in a hole in the fundamental oscillator cavity. You can remove the multiplier block and a female chassis-mount BNC connector will directly thread into the threaded hole. Measure the frequency here (1-2 GHz range) if you can't measure the RF frequency at the output directly. Remember, the signal level here is about ½ watt! Different loading than the multiplier block provides can force the source to lose lock. Padding sometimes helps attain phase lock. Check the 0 terminal for a dip when lock is achieved. (Reference oscillator operation can be verified by a voltage peak on the XTAL test point when tuning the reference adjust control.) This verifies operation of the phase lock circuitry, reference oscillator and fundamental cavity.

After you've checked for phase lock, check the range of the fundamental cavity. The range is usually great enough so that three multiples of the reference oscillator fall within the range of the fundamental cavity, ie, the 10th, 11th and 12th. You'll need this information to help pick the reference crystal frequency. Now you should know whether the unit is worth modifying, and the multiplication scheme.

Ordering the Reference Crystal

Run through several multiplication schemes for the complete source. Pick a crystal frequency that falls in the middle of the cavity range when multiplied. Reference crystal frequencies from 95-113 MHz are common for the brick style oscillators.

All of the units I've seen use a TO-5 style crystal case. These have gotten rather expensive from many vendors, about $50 each. A recent order to International Crystal resulted in a $13 price for a TO-5 packaged crystal, however. An HC-18 style crystal holder can be used with a little extra effort. The cost of this style crystal holder typically ranges from $12-20. To use holder, you have to unsolder the machined oven from the oscillator printed circuit board. You also have to remove the lip at the bottom of the hole the crystal fits in, by drilling it out (first remove the ceramic disk with the socket pins for the TO-5 type case). Then install the crystal directly on the board and reinstall the oven over the crystal.

The Collins sources use a plug-in oven. The crystal leads can be plugged directly into the socket pins on the board. Placing the oven over the HC-18 style case is awkward, but will provide some temperature control.

Crystal specifications (ideal) are:

Mode: Fifth overtone
Oven: 75° C
Shunt capacitance: 4.5 ± 0.75 pF
ESR: 60 Ω max
Frequency tolerance: ±0.0005% @ 75° C
Temperature stability: ± 0.002% from 70-89° C
Spurious response ratio: 2:1

Sources for crystals include:

Bliley (814) 838-3571 (TO-5)
2545 W. Grandview
P.O. Box 3428
Erie, PA 18508

EG&G Cinox
(513) 542-5555 (TO-5)
4914 Gray Road
Cincinnati, OH 45232

International Crystal (both)
10 North Lee
Oklahoma City, OK 73102

JAN Crystals (HC-18)
P.O. Box 06017
Fort Myers, FL 33906-6017

Crystek (HC-18)
2351/2371 Crystal Drive
P.O. Box 06135
Fort Myers, FL 33906-6135

Modification Procedure

CTI, Frequency West and California Microwave Sources

The technique I've generally followed to put a brick source on a new frequency is to install the new reference crystal and then "walk" the multiplier block and fundamental cavity to the new frequency. Going directly for the final output frequency may result in power inadequate for any indication, when tuning the multiplier block.

Crystal Installation

Install the crystal and apply power. A XTAL test point is available on all of the sources except those designed for external reference. Monitor the voltage at this point and adjust the crystal tune control for peak voltage, generally –1.5 to –2.0 V. A check of the reference frequency can be made on the SMA connector. Final adjustment for frequency should be done once the source is installed in your system.

"Walking" the Fundamental Cavity and Multiplier Block

There are several styles of interdigital filter adjustments on the multiplier blocks. Some permit adjustment from above, using small lock nuts. The best way to adjust these is with a tiny, hollow socket, which will pass a hex wrench to the socket on the setscrew. (Although more difficult, the job can be done with long-nose pliers or hemostats!) With a socket tool, you can hold the locking nut close to final position while adjusting the setscrew. This arrangement is convenient, in that you don't have to remove the block for filter adjustment. You still have to remove the block to adjust the output connector coupling.

Another construction style has locking setscrews accessible from the top, with adjusting setscrews on the side facing the brick. You must remove the multiplier block to adjust the screws. Fortunately, by design, the multiplier block will sit "backwards" on the brick for adjustment. Multiplier blocks with waveguide flanges won't seat completely for adjustment, however. Tweaking these is a trial and error process the best I can tell . . . remove the block to tweak and replace to see how well you did.

Walking the source to the new frequency involves adjusting the fundamental cavity tuning screw to slew the cavity a hundred megahertz or so, then tweaking the multiplier block to maintain signal. Tweak the output connector probe adjustment (usually a slotted "screw") and all filter setscrews for maximum signal. The adjusting setscrews nearest the cavity probe are the most critical. Then slew the fundamental frequency further toward the desired frequency and retweak. Monitor the 0 test point when you think you're close to final frequency You should be able to achieve lock before making the final multiplier block adjustments. Sometimes, however, the multiplier block needs to be close to final frequency to provide a good match for the fundamental cavity in order to achieve phase lock. When you achieve lock, make a final pass on the filter and probe tuning screws.

Generally, I've not lost too many dB in output power from the original power level, depending on how far the desired frequency is from the design frequency. One LO I put on 3456.350 MHz was originally set for about 4.0 GHz. When finished, the tuning screws were fully engaged to get maximum power, which was about 0 dBm. Don't despair if this is your case. An outboard MMIC amplifier easily brings up the signal level for driving a mixer. Remember, we are generally using the sources well outside their design range and losing a few dB can be expected.

If you're feeling adventuresome, one tweak which usually makes a significant difference is the depth the multiplier block probe inserts into the fundamental cavity. The probe is threaded. Some units use nuts to lock the threads. Unfortunately, the lead from a ⅛-W coupling resistor is wrapped around the probe underneath the lock nut, in these units. I can assure you this resistor is fragile! Other units tack solder the probe in place. Although the solder wicks into the threaded sleeve, I've been able to remove it, make the adjustment for maximum power, and resolder. This adjustment also affects the match of the probe to the cavity for both styles. You should simultaneously monitor for phase lock and peak power.

Collins 6-GHz Sources

Remove any existing reference crystal and slew the cavity oscillator to the new frequency. There is no multiplier block/filter to tune on these sources since the fundamental cavity is tuned to the RF frequency. Tuning the cavity below the 5.9-GHz range requires capacitive loading. The units with a bolt allow the most tuning range. A Teflon rod is used for the tuning adjustment on these sources. I've soldered Teflon rods from other units to the cavity side of the brass bolt and thus inserted a second Teflon rod into the cavity. Some experimenting is usually required to trim for the 5616-MHz range. The additional loading biases the cavity and reduces the tuning range to less than 200 MHz.

Install the reference oscillator crystal and tweak until you see a signal on the reference oscillator test point. These units will slew the fundamental oscillator over a broad range, so you should see the source lock up without further action on your part if you're at all close on the fundamental frequency. If not, tune the fundamental in increasing increments each way until you obtain lock.

Final trim of the reference crystal should be done after you've installed the source in your system with its own regulator. Remember, the signal at the reference SMA test connector is twice the crystal frequency.

Collins 10-GHz Sources

Like the 6-GHz Collins sources, no multiplier is used in these units. Remove the reference crystal and slew the fundamental cavity to the desired frequency. You may want to monitor the cavity power test point and get a feel for how much it drops off as you tune to the ends of the cavity range.

Next install the reference crystal. Monitor voltage at the test point for the crystal oscillator and adjust for maximum voltage. Then adjust the reference tuning capacitor (REF). A red LED on the unit is quite useful in indicating when the reference tank oscillator locks to the reference crystal oscillator. It will go out when lock is obtained. The reference oscillator operates in the 425-450 MHz range and locks to the fourth multiple of the crystal. Now adjust the reference tuning while monitoring the REF AFC test point. Tune for about –4 volts. You can tune for lower voltage but spectral purity of the reference signal, and thus lock stability, becomes a problem. The "cleanliness" of the reference signal is very sensitive to the

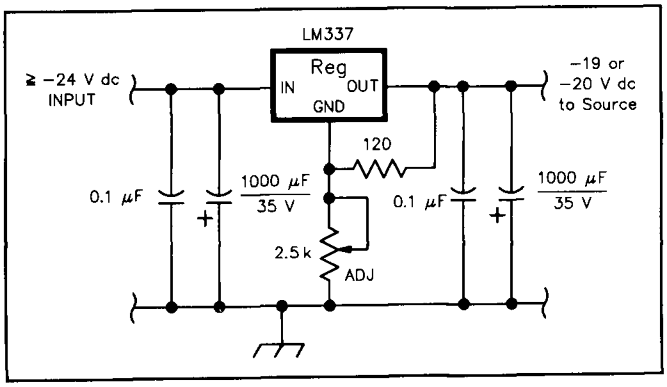


Fig 4—Schematic diagram of a regulator for "brick" signal sources that operate from negative supplies.

REF TUNE adjustment. I've sometimes had to make this adjustment using a spectrum analyzer to get a clean reference signal. Consider trying this if you have trouble obtaining lock.

The source may have already locked at this point. If not, press the red "SWP" button. When the fundamental cavity is close in frequency, and a sufficient sampled signal from the cavity is available, lock should occur. Monitor the "OSC AFC" test point. You should see the voltage swing close to the power supply rails (0 to −18 volts) and then settle in mid-range when lock occurs. The "OSC" red LED will go out when the cavity is phase-locked to the reference.

System Installation

A most critical aspect of installing a phase-locked source in your system is power supply regulation. Small variations in supply voltage result in "FMing." I strongly recommend using a dedicated regulator close to the source. My favorite is an adjustable negative regulator based on the LM337 IC (Fig 4). I substitute a 47-Ω, 2-W resistor for the source, to get the supply close to (less than) final voltage before connecting to the source. Tweak to the final level when you've connected the source.

Source output isolation is important for sources, particularly for those without multiplier blocks or if you're keying the output; for example, a PIN diode switch for beacon applications. Isolators directly on the source output are the best answer. In keying applications, 20 dB of isolation has been sufficient to prevent FMing or loss of lock.

After you've used your source for six months or so, recheck your output frequency. Most of the variation due to crystal aging occurs within this period.