THE NEXT GENERATION OF NO-TUNE TRANSVERTERS

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The current generation of No-Tune Transverter designs, all reprinted in "The ARRL UHF/Microwave Projects Manual," is approaching ten years old. The basic engineering philosophy remains sound, but many new components have been introduced during the past decade and a wealth of experience has been accumulated. This paper is a stage-by-stage re-examination of the basic no-tune transverter, with comments on what might be improved using new components or updated techniques.

INTRODUCTION

The basic philosophy of no-tune transverter design is simple--using reasonably well-matched MMIC amplifiers, reasonably well-matched diode mixers, and bandpass filters designed to be terminated in 50 ohms, a system is built up that may be reproduced without the need for optimizing each stage at the bench. The alternative to no-tune design is a modular system in which each stage is optimized using a network analyzer, signal generator, and spectrum analyzer, and then the stages are interconnected using connectors and coax. The major drawback to the modular approach is that the VSWR on the interconnecting lines is never flat. A filter connected between a "reasonably well-matched" amplifier and a diode ring mixer will not exhibit the nice passband shape seen on the network analyzer. To reproduce the performance of a prototype system, not only must each stage be identically tuned, but the lengths of interconnecting coax lines must also be identical. This is not a severe problem for a manufacturer building a production run of a hundred units, but is a real difficulty for one hundred different microwave experimenters each building a single unit using whatever test equipment might be available. In a no-tune design, the filters have many sections with relatively low loaded Q, and the gain is strategically placed for termination and isolation, so that high performance may be obtained even though none of the termination impedances are exactly 50 ohms and none of the transmission line filter elements are perfectly tuned. Since the production transverters are built from the same artwork as the prototype, performance is very uniform from unit to unit--regardless of who does the actual technician work of soldering on the parts.

The major "modules" of a no-tune transverter are the oscillator, frequency multiplier(s), mixer(s), filters, amplifiers, interconnections and packaging. Each of these will be discussed individually.

MIXERS

The current generation of no-tune transverters uses passive diode ring mixers for both receive and transmit. This remains a good choice, since diode mixers have reasonable dynamic range, good output power capability, good noise performance and the advantage of not requiring any DC power supply components. The new MMIC mixers have conversion gain, small size, low cost, and easy integration into tiny surface-mount systems. At first glance, conversion gain might seem to offer a performance advantage, but in practice it does not. Since the noise figure of an MMIC mixer is usually higher than the loss of a diode ring, conversion gain is only useful if the IF noise figure is high. A no-tune transverter uses a VHF radio as a tunable IF, so the IF noise figure is quite low. In some cases additional conversion loss is needed to avoid overloading the IF. At least one stage of receive RF amplification will be needed with either a diode or MMIC mixer to reduce the system noise figure. For transmit, up-conversion gain is clearly not needed, since the transmit IF port almost universally needs attenuation. Presently available up-conversion MMICs do not offer an output-power advantage over diode mixers, but that will change soon. Where MMIC mixers do offer an advantage is in the design of complete transceiver systems, as they permit higher IF noise figures and lower transmit IF power. The active MMIC mixer is also a good choice for remotely mounted receive converters, as for satellite applications. The conversion gain is then used to overcome the IF transmission line loss between the remote front-end and the operating position.

Once the decision is made to use diode mixers for both transmit and receive, the choice is then between commercially packaged mixers and printed narrow-band mixers with discrete diodes. Below 1 GHz there is no choice, since commercially available mixers are inexpensive and work well and the only printed mixers that will fit on a reasonable sized PC board have low performance. The new TUF package mixers from Mini-Circuits are particularly attractive. At 1296 MHz, the original no-tune printed mixer using a quadrature hybrid ring with an extra quarter wave line works very well. There is no need to replace it from a performance standpoint. It does use quite a bit of board real-estate, however, so a next generation transverter should probably use a commercial packaged mixer. The TUF-5 is a good choice.

Above 2 GHz, packaged mixers are more expensive, so the current practice of using printed transmission line hybrids and discrete microwave diodes is good from a cost standpoint. Especially at 2.4 Ghz, the cost of all components, including packaged mixers, is dropping rapidly, so a next generation 2.4 GHz engineer should examine all the options before blindly duplicating previous artwork. Above 3 GHz, packaged mixers are still relatively expensive, so at present the choice is between printed microstrip with discrete diodes and MMIC active mixers. Some low-cost MMIC mixers still work at 5.8 GHz, but the printed filter-mixer combination on the no-tune 5760 transverter offers better performance for 1296 MHz IF applications.

At 10 GHz and up, close-tolerance microstrip mixers with discrete diodes are the only low-cost choice. This is an area that could use some work.

Several of the current no-tune transverter designs directly connect the receive mixer IF port to the outside world. This was tolerable ten years ago when it could be assumed that the IF port was connected to a narrow band receive input stage, as in the ICOM 202. At the IF port of the mixer are many signals at very high levels--including a large LO component with its spurs and harmonics. VHF radios today have much wider front-end bandwidths, especially those with extended tuning range. These days it is also common to add a low-noise MMIC receive preamp between the transverter and the IF radio. The MMIC amp is usually used without any RF selectivity, so it has gain and not much dynamic range all the way up through the lower microwave spectrum. A number of undesirable results are possible, including saturating the low-noise MMIC IF amp, generating numerous mixing products between the microwave LO, IF radio LOs and local strong signals coupled in on the power supply lines, and presenting the IF receiver with strong signals at frequencies well above the design range of the input filtering. The solution is very simple--a simple capacitor input pi-network low-pass filter at the mixer IF port will reduce the levels of all the microwave signals while passing the desired VHF IF signal. The pi-network should be placed physically close to the mixer IF port and microwave components should be used, even though the "design" frequency is in the VHF range. A good approach for a 144 MHz IF is to connect the input 22 pF chip capacitor directly from the mixer IF port to ground. An example schematic and board layout are shown at the end of this paper. According to diode-ring-mixer application notes, such a network is a good choice for mixer IF port termination when the LO and RF frequencies are much higher than the IF.

FILTERS

The original no-tune transverter, described at the original "Microwave Update" conference (it wasn't called Microwave Update because it was the first one) in Estes Park Colorado in 1985, used computer designed interdigital filters hand constructed from UT-141 and hobby brass sheet. These filters worked exceptionally well, but each one took about 4 hours of skilled labor to construct. The 3456 transverter described at the conference included 6 of these filters and a number of other hobby brass modules, so it was received with admiration, but muted enthusiasm by those who contemplated duplication. Much of the effort in the following years was directed at designing reproducible bandpass filters that could be printed in microstrip form. Interested readers are directed to the numerous articles in the Microwave Update proceedings from 1986 through 1988.

While all of the various filters used in the no-tune transverters are functional and easily reproduced, they share a common flaw--they act like poor (but not nearly poor enough) antennas. Radiation loss is significant, especially in the higher frequency and higher loaded Q versions. While radiation loss is annoying, radiation pickup is far more serious. The radiated energy has to go somewhere, and where it goes is into the transverter enclosure, which is almost certainly large enough to behave as a cavity capable of supporting many modes at the filter frequency. Some of these modes will have E-Field maxima at the ungrounded ends of other hairpin filter elements--often with enough gain in between to satisfy the Barkhausen criteria for oscillation at some frequency. Of course the frequency at which the phase shift around the loop is 0 degrees depends on the cavity dimensions (when behaving in this manner, the latin "cavity dementia" is a more appropriate term), so removing the cover of the the transverter case to find out what is wrong generally "cures" the problem.

One way to deal with radiation coupling between hairpin filter elements is to operate the transverter without an enclosure. This works, as long as there are no other strong local signals that may be picked up by the exposed hairpins. Given the desirability of operating from hilltops, there are likely to be many strong local signals. A preferred method is to use strategically placed RF absorber material inside the transverter case.

Some filter topologies radiate more than others. The edge-coupled straight halfwave line microstrip filter looks suspiciously like a close-spaced yagi suspended by dielectric above ground. One of the lower radiation microstrip topologies is the off-center-tapped-grounded-halfwave used in the no-tune 5760 transverter board and the 2140-2320 MHz multiplier board, both shown in the photograph on page 3-35 of the ARRL UHF/Microwave Projects Manual.

The best solution is to use individually shielded filters. A method that works well for hairpin filters is to add a rectangular shield plate covering the entire filter about 2.5 mm above the printed board and extending about 2.5 mm beyond the edges. The perimeter of the shield is then connected through the board to the backside groundplane. The shield connections to the groundplane should be spaced closer than about 0.1 wavelength at the filter frequency. The example transverter artwork includes locations for plated-through holes to facilitate filter shielding. Adding shields over the hairpin filters of existing G-10 no-tune transverters is relatively easy, and highly recommended.

There are many other practical no-tune filter types besides microstrip--including surface acoustic wave filters for lower frequencies and ceramic dielectric resonator filters at higher frequencies.

Most of these do not lend themselves to low-volume applications, but it is possible to make use of commercially available filters for high-volume applications in or near the amateur bands.

At frequencies above 6 GHz, microstrip techniques fail for thick PC boards (the filters become very good antennas), and more conventional cavity techniques become attractive. It should be possible to print the coupling structures and use a micrometer to set the post height of coupled cavity filters using plumbing caps in a no-tune design, as long as the IF is high enough that filter tuning is not critical. A 10.368 GHz transverter with a 1296 MHz IF might be possible. Another possibility is to use printed stripline filters with compressed sandwich construction. All of these techniques are more involved than the simple printed filters used on lower microwave bands, and consequently tuned transverters using single resonator plumbing cap filters and simple tune-up using a power meter become relatively attractive.

AMPLIFIERS

The no-tune transverter described in 1985 used ceramic packaged Avantek MSA 0235 MMICs for all of the low-level microwave gain at 3.5 GHz. The early single board transverters used a wide variety of plastic-package MMICs from the MSA0104 series. The introduction of the MSA1104 in 1988 offered higher power for the lower microwave bands. The MSA0885 minimum-feedback amplifier was specified in a number of transverters, but it has generally fallen from favor in amateur designs due to stability problems. At KK7B the only silicon amplifiers used in recent designs are the MSA0686 for no-tune receiver input stages, the MSA1104 for no-tune exiter output stages and LO multiplier drivers, and the MSA0286 for everything else. The MSA0286 is the plastic package version of the MSA0235 used in the original no-tune transverter, and it has the virtues of moderate noise figure (6.5 dB), moderate gain (12.5 dB at VHF, 7.5 dB at 3.5 GHz), moderate power output (+7 dBm at 2.3 GHz), unconditional stability, and most important but often forgotten--excellent input and output match to 50 ohms across the entire usable frequency range. The MSA0286 is an almost ideal low-level, low-cost active isolator with gain. Many stability problems in no-tune transverter designs may be cured by simply replacing the offending stages with the MSA0286.

Silicon MMICs are a mature technology that has been widely described in the literature and exploited in numerous applications. GaAs MMICs are at approximately the same stage now that silicon MMICs were a decade ago. GaAs MMICs offer higher stable gain, higher power output, a higher upper frequency limit, and lower noise figures than silicon MMICs, at comparable cost. One particularly useful device is the HP MGA86576, with 23 dB gain, 1.6 dB noise figure, and +6 dBm output power at 4 GHz. The MGA86576 offers the possibility of replacing multiple stage amplifiers with single devices, building a reproducible 5.76 GHz no-tune transverter with RF gain on the board, and even building a no-tune 9.072 GHz LO multiplier for a 10.368 GHz transverter with 1296 MHz IF. There are only two areas where the silicon MSA0286 still has an edge: the input match and output match of the MGA86576 are poor at low frequencies, and its source leads must be connected to the ground plane with very short leads (0.062" FR-4 board is too thick) for stability. Since thinner circuit board is generally used at higher frequencies, the MGA86576 would seem to be the logical choice for gain in a new 3.456 GHz or 5.760 GHz transverter design.

Numerous other GaAs MMICs from various manufacturers are also very attractive. One example is the Tri-Quint TQ9132, offering +18 dBm at the 1 dB compression point from 800 to 2500 MHz with a 6 V supply. A most interesting development is the new series of integrated front-end amplifier ICs. As an example, the TriQuint TQ9205 2.4 GHz Amplifier/Switch Front-End offers a 200 mW transmitter amplifier, 3.5 dB noise figure receive preamp and TR switching all on a single IC, for direct connection between the image filter and antenna.

One trend that should be noted in microwave amplifier technology is the move to lower supply voltages and better efficiency. The limiting factor in reducing the size of personal communications equipment is battery technology, and 3 V systems are being widely developed. There are several immediate benefits to amateur microwave engineers. First, it is easy to build a very stiff 5 or 6 volt supply using a linear regulator from the amateur portable standard 12 volts. Portable equipment designed for a 13.8 volt supply (as the original no-tune transverters) may still work as the battery voltage drops to 12.0 volts, but performance is probably degrading by the time the battery is at 11.5 volts. Second, the latest generation of MMICs is designed to work over a wide range of supply voltages from 6 volts down to 3 volts, so behavior as the voltage fluctuates is usually much more graceful than the previous generation. Finally, now that it is possible to build a complete medium power 2.3 GHz SSB-CW transceiver with a total receive power drain of 60 mA at 6 volts, casual weekend grid expeditions with flashlight battery power and a mountain bike are practical. This could be the hook to catch the next generation of hot microwave experimenters.

OSCILLATORS

The current generation of no-tune transverters uses a Butler 5th overtone crystal oscillator operating near 100 MHz, followed by frequency multipliers to the desired final output frequency. There are several oscillator areas that could use some improvement. First, a number of builders have experienced some difficulty getting the oscillators to start. A good quality crystal oscillator uses only enough feedback to ensure reliable starting. If the feedback components are selected for reliable starting with a "high activity" crystal and hot transistors, some crystals on the "low activity" side of average may not start, especially if the transistors are different than the ones in the prototype. The easiest fix for an oscillator that won't start (after the obvious inspection for assembly errors and missing power supply connections) is to replace the transistors with devices with higher VHF gain. An oscillator that won't start with an unmarked genericVHF transistor will often work with a "real" 2N5179 or MRF901. It is rarely necessary to replace the crystal.

Some experimenters have discovered that raising the base bias voltage of the oscillator transistor will greatly increase the oscillator output power. This is sometimes done while trying to squeeze the last half milliwatt out of the local oscillator chain. It is easy to overdrive the crystal in the Butler-emitter-follower circuit if the base voltage is increased too much. The syptoms of crystal overdrive are reduced frequency stability followed by erratic frequency jumps of a few kilohertz and then finally crystal failure. Failure may occur after a few months of continuous operation.

The frequency of oscillation is not set by the amplitude bandpass characteristics of the crystal, but by the number of degrees of phase shift around the oscillator loop. Oscillation occurs at the frequency where the output signal is fed back to the input exactly in phase with the input signal. The reason that crystals are used to set the frequency of oscillators is that the phase versus frequency curve of a quartz crystal is exceptionally steep--on the order of 1 degree per 50 Hz for typical 5th overtone crystals near series resonance. In comparison, the phase of S21 for a good UHF transistor may only vary a few degrees every 50 MHz. Since the total phase around the loop must add to 0 (or 360) degrees, changing to a different transistor type will generally change the phase shift required from the crystal, and hence the oscillation frequency. A batch of crystals that all oscillate within 500 Hz of 90 MHz with a particular batch of 2N5179 transistors may all oscillate 500 Hz lower with a batch of transistors from a different manufacturer. This is frustrating for equipment manufacturers, and upsetting to beginning microwave enthusiasts who expect their new Downeast Microwave transverter to convert 1296.100 to exactly 144.100 MHz. Frustration over a digital dial that doesn't read correctly to the nearest 100 Hz at 2.3 GHz is the mark of a true amateur.

Far more important than the exact frequency of oscillation is the long and short term stability of the oscillator. If the LO frequency is always 23.0 kHz low at 1296, the calling frequency then becomes "144.123" on the IF radio--an inconvenience so minor it is embarrassing to even mention. For many no-tune transverters however, the dial is "about 25 kHz off," which means continuous tuning around the calling frequency to find the other station buried in the noise. Knowing the frequency offset of a microwave transverter to within a few hundred Hz over a many month period is a benefit of good long term stability.

Especially on the higher bands, and when the transverter is subject to the changing temperatures inherent in portable operation, short-term stability is also a problem. Having to constantly retune the frequency during SSB contacts is a sympton of poor short-term stability. Most no-tune transverters operated portable have poor short-term stability, and many operators have simply accepted it as a fact of life. There are a number of easy ways to improve the short term stability of a 5th overtone Butler oscillator. To a very good approximation, short-term frequency drift is entirely due to changes in the crystal temperature. Stability may often be improved by a factor of ten by simply slipping a foam packing bead (the ones shaped like the number 8) over the crystal. A more serious approach is to drill the oscillator circuit board so that the crystal may be mounted on the opposite side of the board, and then wrap the crystal in packing foam and tape it down. It is important to thermally insulate just the crystal, not the entire oscillator. The oscillator components--especially the voltage regulator and series bias resistor for the MSA1104 buffer amplifier, get very warm, and they will get even warmer if they are enclosed in insulation. Enclosing the entire oscillator and buffer circuit in packing foam is a good way to increase warmup drift and move the crystal to a less stable point on its "room temperature" specified frequency/temperature curve.

Long-term stability is more difficult to improve. As previously mentioned, raising the Butler oscillator output power by changing the base bias resistor will degrade long-term stability. Higher quality ("more expensive") crystals may be tried. The best approach to reasonable long-term stability with a 5th overtone oscillator is to run separate power to the crystal oscillator and leave it on 24 hours a day, 365 days a year, in a 72 degree room. A novel approach would be to seal the oscillator in a film can with a few fishing sinkers and bury it in the gravel of a tropical fish tank. Extra points are awarded for creativity.

The shortest term frequency stability is the kind that sounds like modulation. I will avoid the term "phase noise" because it has been misused by many authors. The effects of poor short-short-term stability are similar to the effects caused by familiar propagation paths. Medium-bad short-short-term stability (the "warblies") makes clean sine wave CW signals sound like airplane-tropo propagation, and real bad short-short-term stability makes them sound like Aurora scattered signals. Bad short-short-term stability is usually a symptom of poor Phase-Locked-Oscillator design or adjustment. Butler 5th overtone oscillator-multiplier local oscillators have good short-short-term stability up through at least 10 GHz.

With the oscillator discussion completed, a few recommendations for the next generation of no-tune transverters follows. Three new approaches will be explored: computer clock oscillators; fundamental oscillators; and the new phase-locked-loops.

COMPUTER CLOCK OSCILLATORS

Packaged, laser-trimmed computer clock oscillators are now available off-the-shelf from numerous vendors. Short-short, short and long-term frequency stability are about the same as the Butler 5th overtone oscillators--they may be using the same circuit. They cost less than the parts to build an overtone oscillator, are much smaller, and work every time. A sample of ten purchased for the example 1296 transverter resulted in frequency errors scattered within about plus or minus 25 kHz of the desired local oscillator output frequency--about the same as low-cost 5th overtone crystals. The only disadvantages are that the most commonly available models operate near 60 MHz, requiring more multiplication to reach the desired output frequency, and only certain frequencies are available off-the-shelf. Prices become attractive for custom frequencies in lots of 100 or more--but that discourages development. Fortunately, two of the most common frequencies, 60 MHz and 64 MHz, are very useful for amateur microwave applications. 64 MHz multiplies nicely up to 1152 MHz, the preferred LO for 1296 to 144 MHz converters and the common subharmonic for the bottom end of the weak-signal segment of all the amateur microwave bands. 60 MHz multiplies up to 2160 MHz, the LO for both 2304 MHz to 144 MHz transverters and 3456 MHz to 1296 MHz transverters. These are not just numbers--practical circuit board local oscillators for these two microwave frequencies have been developed, and they have some clear advantages over previous no-tune local oscillators.

FUNDAMENTAL MODE CRYSTAL OSCILLATORS

Long and short-term frequency stability may be improved by replacing the 5th overtone crystal oscillator operating near 100 MHz with a fundamental crystal oscillator operating near 10 MHz followed by frequency multipliers. As a fringe benefit, fundamental crystal oscillators may be frequency tuned a small amount without loss of stability. This permits setting the local oscillator to an exact desired frequency or using the variable crystal oscillator (VXO) as the main tuning control in the system. The disadvantage of fundamental crystal oscillator-multiplier systems is that the local oscillator spurious outputs are much closer to the desired output, and thus more difficult to suppress. The additional frequency multiplier stages also add circuit complexity. Since the additional circuitry is in the VHF range, where no-tune techniques are not available due to wide component tolerances, the multiplier stages must be tuned. A well-engineered VHF frequency multiplier stage will include some test points, where a meter may be connected and all of the previous stages tuned for maximum output. Such circuits are not "no-tune," but "easy-tune."

It is important to note that the spurious outputs of the early multiplier stages must be exceptionally well suppressed. For example, suppose that the output of the 100MHz multiplier has spurious outputs at 90 MHz and 110 MHz that are 40 dB below the desired carrier. This is expressed as "-40 dBc." This level of local oscillator spurious suppression is acceptable in simple, low-powered VHF applications like scanner receivers and hand-held radios that incorporate more filtering after the power amplifier. If this 100 MHz signal is used to drive a frequency multiplier, however, the spurious outputs will increase by 6 dB every time the frequency is doubled or 20 dB every time the frequency is multiplied by 10. After multiplying by 10 to 1000 MHz, the spurious outputs at 990 MHz and 1010 MHz will be -20 dBc. This might be acceptable in some low-performance experimental systems. After frequency doubling to 2000 MHz, the spurs at 1990 MHz and 2010 MHz are -14 dBc. If we multiply the original 100 MHz signal all the way up to 10 GHz, the spurious sidebands will be just as strong as the carrier, and the local oscillator will be useless as a local oscillator, but perhaps find some application as a comb generator. Communication theorists will also note that as the spurious levels rise, the number of sidebands also increases. Communications experimenters will recognize this as the old standard method of generating VHF FM by applying a small amount of phase modulation to a stable 6 MHz crystal oscillator and then multiplying it up to 150 MHz. All of this tutorial on local oscillator spurs is just a reminder that the casual design of microwave local oscillators is an invitation to problems. Any 100 MHz signal source that is multiplied up into the microwave range needs to be very clean. A reasonable goal for spurious sideband suppression is -80 dBc. This generally can not be achieved on a single printed circuit board without extensive shielding. Running the fundamental oscillator in a separate shielded enclosure with coax output and a separate helical resonator filter is a good approach. Balanced frequency doublers and square-wave odd-harmonic multipliers that suppress the undesired harmonics are also highly recommended. Despite the additional engineering challenges, the fundamental crystal oscillator followed by frequency multipliers and extensive filtering remains the conservative performance standard.

PHASE LOCKED OSCILLATORS

Phase-locked-loops offer an inexpensive method of improving the frequency stability of microwave oscillators. Phase-locked-loop oscillators offer numerous advantages over oscillator-multiplier systems. The most important advantages are reduced parts count, small size, frequency agility, low cost and reduced power consumption. Frequency agility is not a major concern in amateur microwave transverters, but the other advantages are significant. A phase-locked-osillator starts with a low-performance microwave oscillator with enough power at the right frequency and then improves its performance using a feedback control loop. The loop parameters determine the improvements. The feedback loop must be engineered for each application, microwave oscillator, phase detector, and reference frequency. A perfect feedback loop will generate an output signal that is a perfect replica of the reference signal multiplied up to the desired output signal. This is exactly what is accomplished by a perfect oscillator-multiplier-filter local oscillator system. There is little difference between serious, expensive phase-locked-oscillators and serious, expensive oscillator-multiplier systems except for a few applications like very-long-baseline-interferometry where using phase as a variable is not an option. The popularity of phase-locked-oscillators for fixed frequency microwave systems arises from the fact that "pretty good" feedback loops are much simpler and less expensive than "pretty good" frequency multiplier chains. The differences between a merely "good" phase-locked-oscillator and a merely "good" oscillator-multiplier are significant. The far-out spur suppression of a phase-locked-oscillator is likely to be much better than that of a low-cost oscillator-multiplier. The close-in noise performance of an oscillator-multiplier will be better than a phase-locked-oscillator. Many simple phase-locked-loop oscillators that work for FM and other wideband modes are unacceptable for SSB and CW. The famous "Auraural Local Oscillator" in the Yaesu FRG-9600 is a good example. On the other hand, a well engineered microwave phase-locked-oscillator can have excellent performance and low cost. The components are now available--what is needed is feedback control loop engineering talent and time at the bench to bring low-cost phase-locked-oscillators up to the specific performance requirements of weak-signal amateur microwave transverters.

FREQUENCY MULTIPLIERS

The no-tune transverters have all used variations of the Schottky diode frequency multiplier described in 1985 in "A Clean Microwave Local Oscillator." This frequency multiplier has become a standard, and has been duplicated and successfully adapted by hundreds of microwave experimenters. The only performance area where the Schottky diode frequency multiplier does not measure up is DC efficiency. By the time all of the MMIC class A linear amplifier stages are running, the current is typically over 100 mA from the 12 V supply. Careful choice of MMIC devices can reduce the supply current a little, but usually at the expense of stability. Since the state-of-the-art in microwave systems is being pushed by pocket-sized battery powered devices, the amateur microwave experimenter standard is unlikely to be adopted for consumer electronics.

For amateur service, however, the Schottky diode multiplier is still an excellent choice. As long as the oscillator itself is left running for frequency stability, the batteries may be saved by cutting the power to the frequency multiplier amplifiers between scheduled contacts. Since most amateur microwave work is scheduled, rather than random, this may well result in longer battery life than using a lower total-power-drain phase-locked-oscillator that must be constantly powered. Of particular significance for experimental work is the graceful failure of oscillator-multiplier systems. As device gain drops with dying batteries, the local oscillator power gradually reduces. With a phase-locked-oscillator, once the loop unlocks, there is no way to continue the contact.

If lower frequency sources are used, either fundamental crystal oscillators or packaged computer clock oscillators, then frequency multipliers that work well at lower frequencies are needed. The Schottky diode multiplier is particularly attractive in the lower microwave range where printed no-tune filters may be used to select the desired harmonic. It is less attractive at VHF, where the filter elements must be manually tuned. On the other hand, some techniques that suppress the undesired harmonics work well at VHF. The balanced Schottky diode doubler, described in detail by Hayward and DeMaw, provides excellent suppression of odd harmonics. Diode frequency doublers are particularly attractive now that MMIC amplifiers are inexpensive and widely available. One aspect of balanced diode frequency multiplier operation that has received little comment in the past is the need for good spectral purity on the drive signal. The input must be a symmetrical sine wave for proper operation. A symmetrical square wave will result in mostly DC output, with very little second harmonic. An asymmetrical drive waveform will generate odd harmonics. A very good choice for the input circuit to a balanced diode frequency multiplier is a conventional pi-network with high enough Q to offer good harmonic suppression. Unfortunately, by the time the pi-network, balanced drive transformer, diode pair, DC ground, AC coupling and MMIC amplifier are incorporated, a large number of parts has been used just to double the frequency. Since it takes three doublers just to get from a good fundamental crystal oscillator to the normal frequency range of a 5th overtone Butler oscillator, this is not a commercially viable option. It is a very clean, conservative approach, however, and is highly recommended for custom engineered high performance systems.

To generate odd harmonics, a square wave may be used. A square wave may be generated by hard-limiting a clean sine wave. A sine-to-square wave generator that has been successfully employed in a number of designs is a fast CMOS hex-inverter IC with the input inverter configured as a limiter with resistor feedback and the five remaining inverters connected in parallel. If a six volt supply is used, then the output is a 5 volt peak-to-peak square wave with good output drive capability. Simple Fourier theory may be used to predict the levels of the various odd harmonics. The power drain of the IC is determined by the input frequency. A 74AC04 hex-inverter driven at 72 MHz with the output taken at 216 MHz runs very warm. A 74AC04 driven at 46 MHz with the output at 230 MHz runs cooler, and a 74AC04 driven at 20.6 MHz with the output at 144.2 MHz runs cool. The filtering needed to select the desired harmonic depends on the output frequency range and the multiple. A two-section filter will provide reasonably clean 3rd harmonic output with fundamental HF crystal drive. For multiplying 20.6 MHz by 7 up to 144.2 MHz, a three-section, 10 MHz bandwidth filter followed by an MMIC amplifier and a two-section, 10 MHz bandwidth filter provides 45 dB spurious suppression and broad, easy tuning. At higher drive and output frequencies, the even harmonic suppression is not as good, and the filtering must be improved accordingly.

The example transverter at the end of this paper uses a 64 MHz crystal clock oscillator followed by a CMOS frequency tripler to 192 MHz. The output is then multiplied by 6 to 1152 MHz. The spurious outputs are all more than 56 dB below the desired 1152 MHz output, except for the 2304 MHz output at -30 dBc and the 3456 MHz output at -46 dBc. This oscillator is simpler and cleaner than the one in the original no-tune transverter, which uses a 96 MHz Butler 5th overtone oscillator with a Schottky diode times 6 multiplier to 576 MHz and then an unbalanced Schottky diode doubler with filters and amplifiers at 1152 MHz.

The over-driven MMIC frequency multiplier, as used in the no-tune 5.76 GHz transverter, works well as a doubler, but the output levels at higher multiples are very drive and match dependent. Small changes in power supply voltages or interconnecting coax lengths may be expected to have large effects on the frequency multiplier output level when overdriven MMICs are used as higher-order frequency multipliers.

Balanced diode doublers work well all the way up through the microwave spectrum. The same 180 degree hybrid microstrip ring normally used for mixing may be used for frequency doubling simply by reversing one of the diodes. The junction of the two diodes becomes a microwave port at twice the ring frequency instead of an IF port. A prototype 180 degree microstrip hybrid ring doubler driven with 4.536 GHz at +8.5 dBm provides 0 dBm output at 9.072 GHz with about 30 dB suppression of the drive frequency.

INTERCONNECTIONS

The present generation of no-tune transverters incorporates as much low-level circuitry as possible on a single board. This was done to make it as easy as possible to get a simple station on the air, which could then be enhanced with external receive preamplifiers and transmit power amplifiers. The number of new stations on the microwave bands using no-tune transverters is evidence of the success of this approach. There are several directions that may be taken for a next generation of transverters. One is to incorporate the receive preamplifier, transmit power amplifier, DC switching, antenna switching and IF switching all on a single board. This approach is being pursued by Steve Kostro of Downeast Microwave for his new line of microwave transverters. It has the big advantage of providing a good microwave product line that is easy to manufacture and service. A different approach is to separate the transverter functions into simpler, smaller boards, so that advanced experimenters may combine them into high-performance custom engineered systems. Both types of systems are needed: "microwave appliances" to increase band occupancy and fill the demand for microwave amateur satellite hardware; and separate modules to allow advanced microwave experimenters to continue pushing the state of the art by custom engineering portable contest stations, tropospheric DX stations and EME stations. Since Microwave Update is an advanced experimenters forum, the modular approach will be discussed.

A few guidelines for modular microwave systems will make them much easier to use. Well-matched 50 ohm ports on all of the modules make life much easier. If a port is not close to 50 ohms, the VSWR on the interconnecting coax will be high, and performance will be a function of the interconnecting transmission line length. Where possible, the ports of modules should be the inputs and outputs of MMICs, preferably types selected for low VSWR. If coax and connectors must be used between poorly match parts of the circuit--for example the ports of multiple section filters and diode mixers--the best interconnecting line length should be experimentally determined.

At higher microwave frequencies, 1 mm can make a difference of several dB.

A conflicting requirement is that all ports should be band-limited. As discussed in the mixer section, an IF port should present just the IF signals to the receiver, not a whole spectrum of VHF and microwave intermodulation products and local oscillator breakthrough. Receiver inputs should have at least a single section transmission line filter at the input connector, so that an antenna or external broad-band low-noise preamplifier may be directly connected.

Individual modules need to be enclosed and shielded. Radiation coupling can make an otherwise stable system oscillate. In low power portable operation, little gain is used at microwaves for either transmit or receive. If the gain is low, there is not much chance of unexpected oscillation. Gain has often been added to no-tune transverters in the form of well-shielded commercial surplus packaged amplifiers. As long as the no-tune transverter board is kept out of the antenna beam, such systems usually work well. For a 100 W output 2.3 GHz system constructed entirely from new components, the transmit amplifier gain needs to be about 60 dB. With 60 dB of microwave gain, shielding is not optional. Even well-engineered microstrip circuits tend to have incidental radiation coupling somewhere around -50 dB. The old rule that the reverse isolation must exceed the forward gain for stability applies here. If the individual filter sections are shielded as discussed in the filter section, there is no problem with enclosing each circuit module in its own metal box with a power supply feedthru capacitor and SMA connectors for RF. When individual modules are enclosed in metal boxes, it is relatively easy to reduce incidental coupling to below -80 dB. Well shielded microwave systems look really nice on the spectrum analyzer.

PACKAGING

The present generation of no-tune transverters is sensitive to packaging, primarily because of radiation coupling between unshielded microstrip filters. If the guidelines presented in this paper are followed, the next generation will be far less sensitive to changes in packaging. Advanced microwave experimenters should automatically question any packaging scheme that looks clever and neat. There is a natural tendency to want to duplicate neat looking things, even though they may not offer the best solution for the problem at hand. For serious home stations, rack mounting is appropriate. For a living room station, an IF rig in a walnut-and-birch cabinet with brass front panel and National dial looks very nice, especially if the microwave transverter is remotely mounted on the back of the dish. For vehicle portable operation, heavy, sturdy construction is needed, and for backpack portable, light weight is important. A backpack or bicycle portable system package may be built out of sheet styrofoam fastened together with epoxy. The styrofoam case provides excellent physical protection for the circuitry inside, and weighs almost nothing.

For advanced microwave experimenters, custom packaging is the mark of a custom engineered system. There should be no need for two amateur microwave systems to look alike.

EXAMPLE NEXT-GENERATION TRANSVERTER

A 1296 MHz transverter has beed developed using many of the ideas discussed in this paper. Figure 1 is the schematic, Figure 2 is a photograph of the "component side" of the board, Figure 3 is a photograph of the foil side of the board, and Figure 4 is the X2 FR-4 printed circuit artwork. The artwork has a copywrite--for permission to reproduce, contact the author. On the 3.5" X 5" circuit board are a 64 MHz computer clock oscillator, CMOS frequency tripler, 5 slug-tuned VHF coils with a VOM test point for tuning, a X6 Schottky diode frequency multiplier, an 1152 MHz bandpass no-tune local oscillator amplifier employing MSA0286 MMICs and shielded hairpin filters, a 1296 MHz bandpass no-tune RF amplifier using shielded hairpin filters, a TUF-5 mixer, and pi-network IF port coupling. Astute readers will note that there are only enough functions on the board for either transmit converting or receive converting. The only difference between a transmit converter and receive converter is the selection and orientation of the MMICs in the bandpass RF amplifier section. To build a complete transmit and receive converter, a second identical board is used. For transceive operation, the 64 MHz oscillator, CMOS tripler and Schottky diode multiplier are left off the second board, and a little 1152 MHz energy is picked off the first board with a capacitive probe near the output of the 1152 MHz bandpass amplifier and fed into the first filter section of the second board. The spurious performance of this transverter board, either as a receive or transmit converter, is superior to the previous generation. The component cost for a complete transceive system is similar, but the size is reduced and the oscillator does not need tuning. The 5 VHF slug-tuned coils on the CMOS tripler output are tweaked by connecting a VOM to test point "TP" and tuning for maximum voltage. If the individual boards are soldered up into separate enclosures with SMA connectors and power supply feedthru capacitors, a very high performance system may be built.

REFERENCES

The references on no-tune transverters span a decade. Many of the ideas presented here have been previously discussed in the references, and it is common for workers studying the same problems to independently reach the same conclusions. The intent of this paper has been to present as many useful ideas as possible in a compact form to the next generation of no-tune transverter engineers. It is impossible to credit the originator of each idea expressed here, as many of the sources are obscured in the free exchange of information common in the amateur microwave community.

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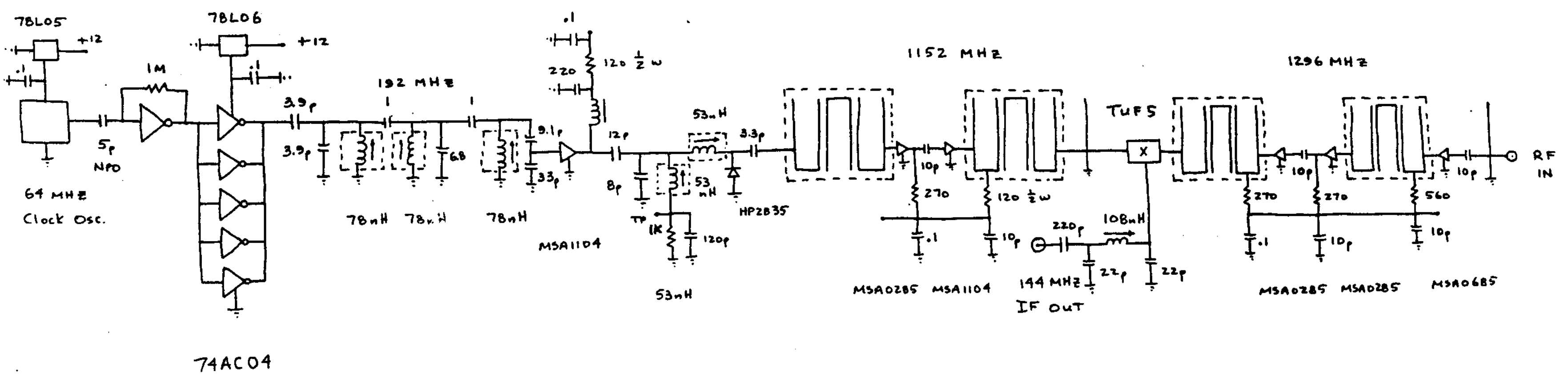

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FIGURE 1

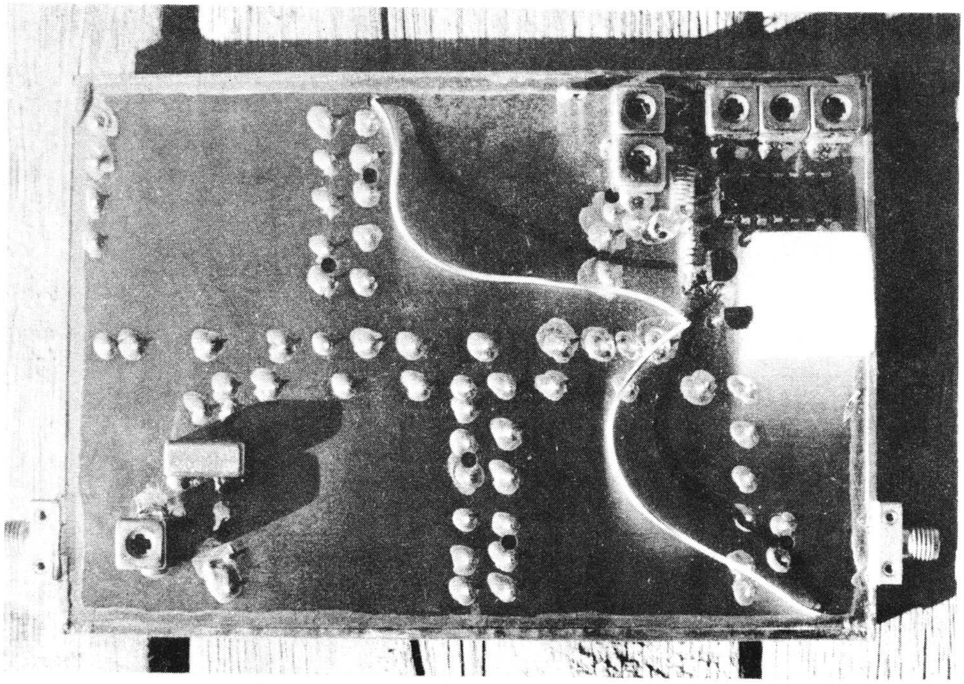

Figure 2

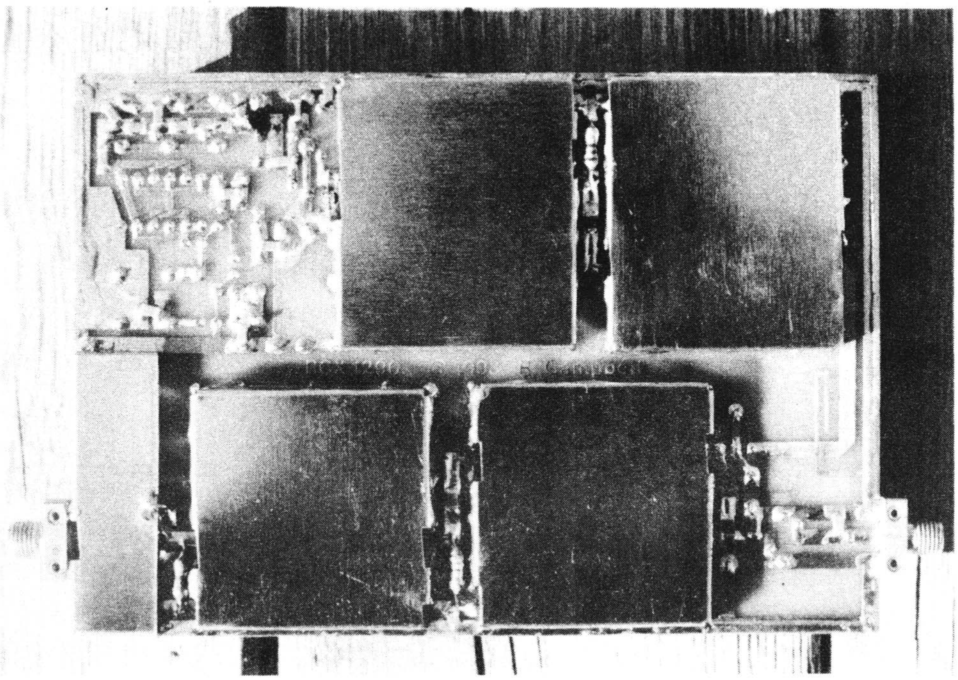

Figure 3

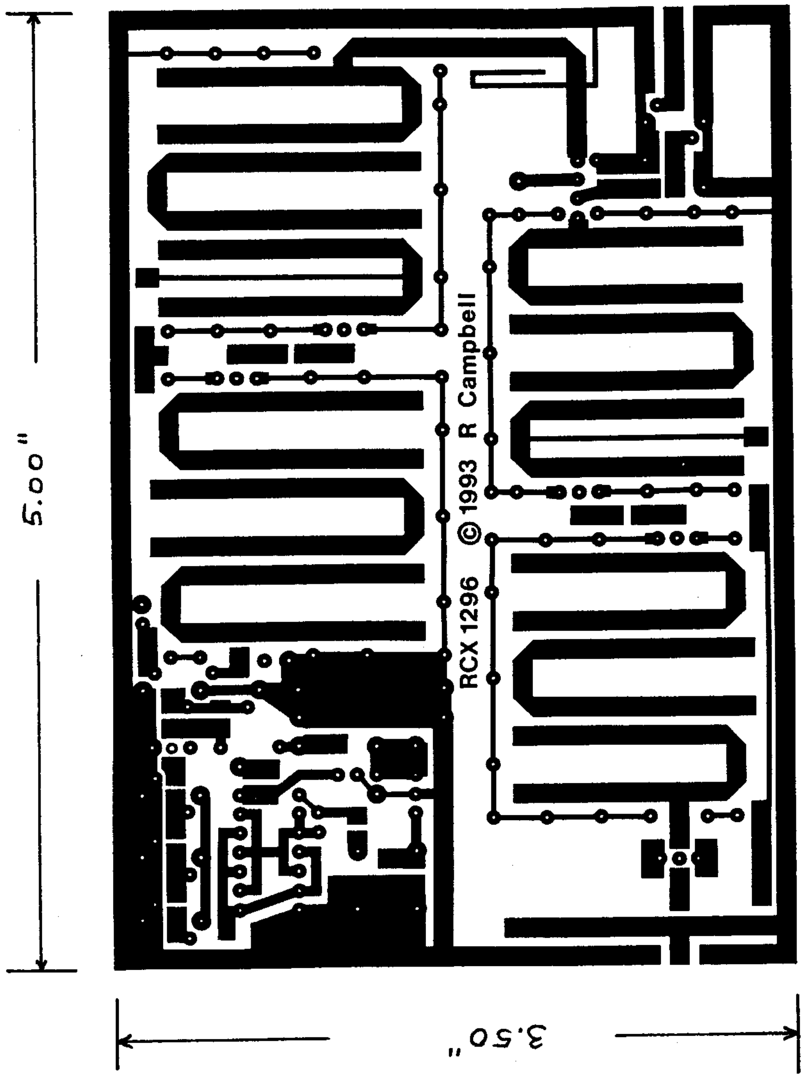

FIGURE 4