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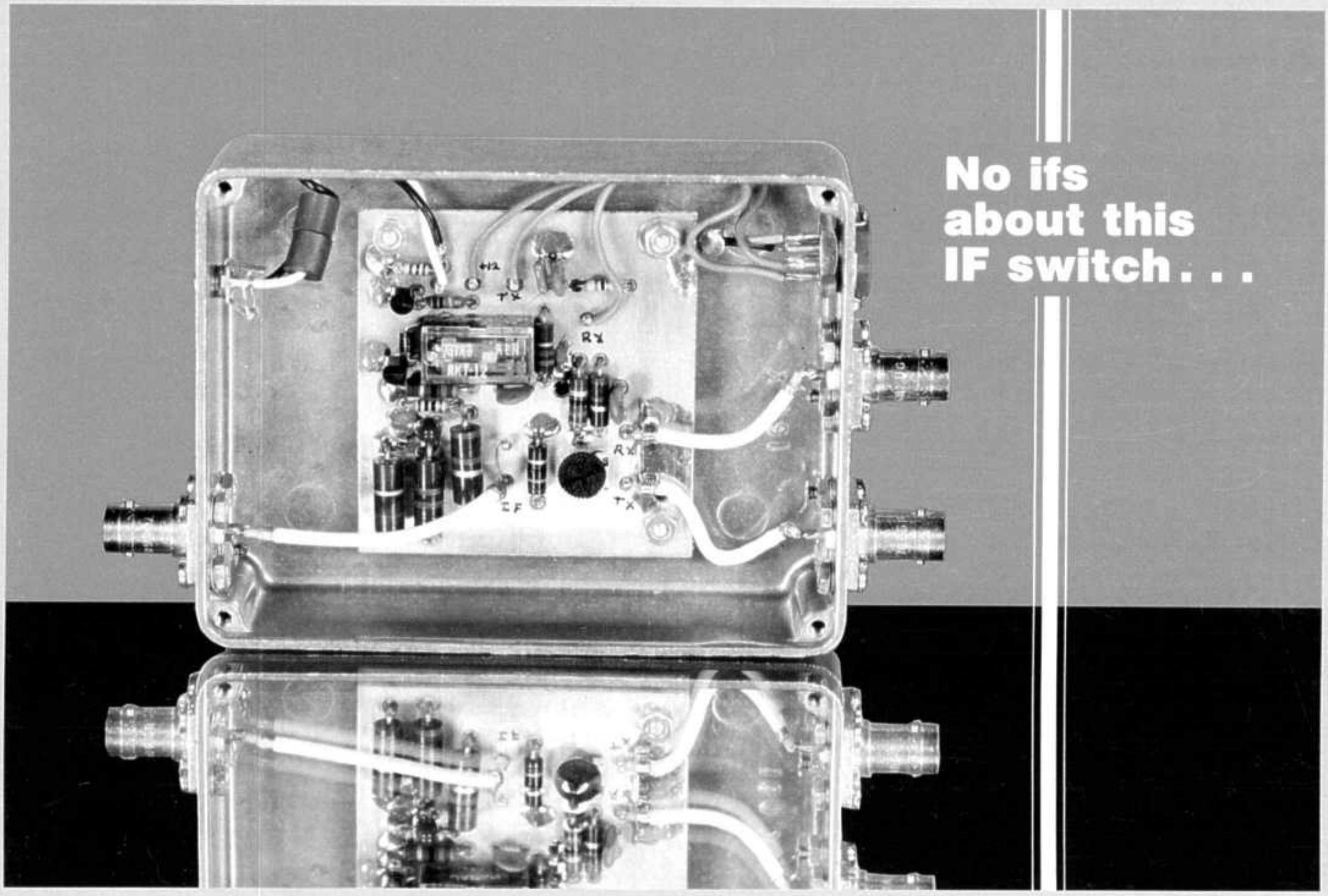

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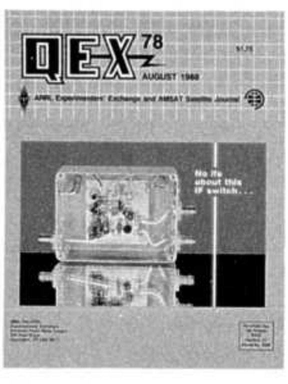

ABOUT THE COVER

Need a convenient and nearly fail-safe way of switching the output of your 144-MHz IF rig for use with a transverter? See page 3 for the details!

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Purposes of *QEX*:

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2) document advanced technical work in the Amateur Radio field
3) support efforts to advance the state of the Amateur Radio art.

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

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Empirically Speaking...

AO-13 and Beyond

As proclaimed by last month's "Empirically Speaking" headline, the AMSAT Phase 3C launch was successful. On June 15, 1988, Phase 3C, sandwiched between MeteoSat and PanAmSat, was launched by the European Space Agency aboard an Ariane 4 rocket from the Kourou Space Center, French Guiana. This event was billed by ESA as a "demonstration launch" of the redesigned Ariane space vehicle. Frankly, it was a superb demonstration as evidenced by live launch coverage on CNN and C-SPAN. C-SPAN also covered the entire Ariane flight and subsequent satellite deployment. Live television reportage of the launch was complemented by a 20-meter net that provided the Amateur Radio community with the latest satellite launch information.

Phase 3C became AO-13 when the beacon was switched on 60 minutes after separation from the launch vehicle. Although months of careful preparation and effort paid off with a successful launch, AMSAT-DL engineers still had an enourmous amount of work to do. AO-13's GTO (Geostationary Transfer Orbit) was highly ellipitical, at an inclination of near 0 degrees (almost directly over the equator) and was losing 30 to 40 km per day in perigee as AO-13 moved through the upper reaches of the F1 layer. After careful calculations, a two-step burn of AO-13's rocket motor was planned. The first, June 22, was for 50 seconds and raised the perigee of AO-13's GTO, eliminating atmospheric drag as a problem. The second, July 6, was a 5.5 minute, go-for-broke effort that completely depleted the on-board fuel supply. This second burn placed AO-13 in its near-Molniya orbit of 2,200 km (1,490 miles) by 36,000 km (22,360 miles), inclined 58°. To make a long story short, AO-13 is exactly where it is supposed to be.

The AMSAT-OSCAR story, while currently at the pinnacle of success, does not stop here. There are additional Phase 3 possibilities to explore. Two orbiting OSCARs, 180° apart, providing 24-hour coverage similar to a geostationary satellite, for instance, and PACSAT (for PACket SATellite). PACSAT was approved by the AMSAT Board of Directors in November, 1987 and the project is currently being expedited. This type of OSCAR will be placed in low earth orbit (LEO) and will have some distinct advantages: (1) PACSAT's earth station transmitter power and antenna requirements will be modest as compared to those currently being used: (2) PACSAT's "window" will be short and often, allowing rapid delivery of messages on a "store and forward" basis (think about QSLing a VK9 using PACSAT!) and (3) PACSAT's use of current satellite designs will keep R&D, build and launch costs low attracting new OSCAR sponsors. AMSAT-LU (Argentina), for example, will be sponsoring one of the first PACSAT's. (Note: The first PACSAT has a possible (not confirmed) launch date in early January, 1989.) When all the factors are considered, PACSAT certainly will open space communications to more amateur radio operators than ever before.

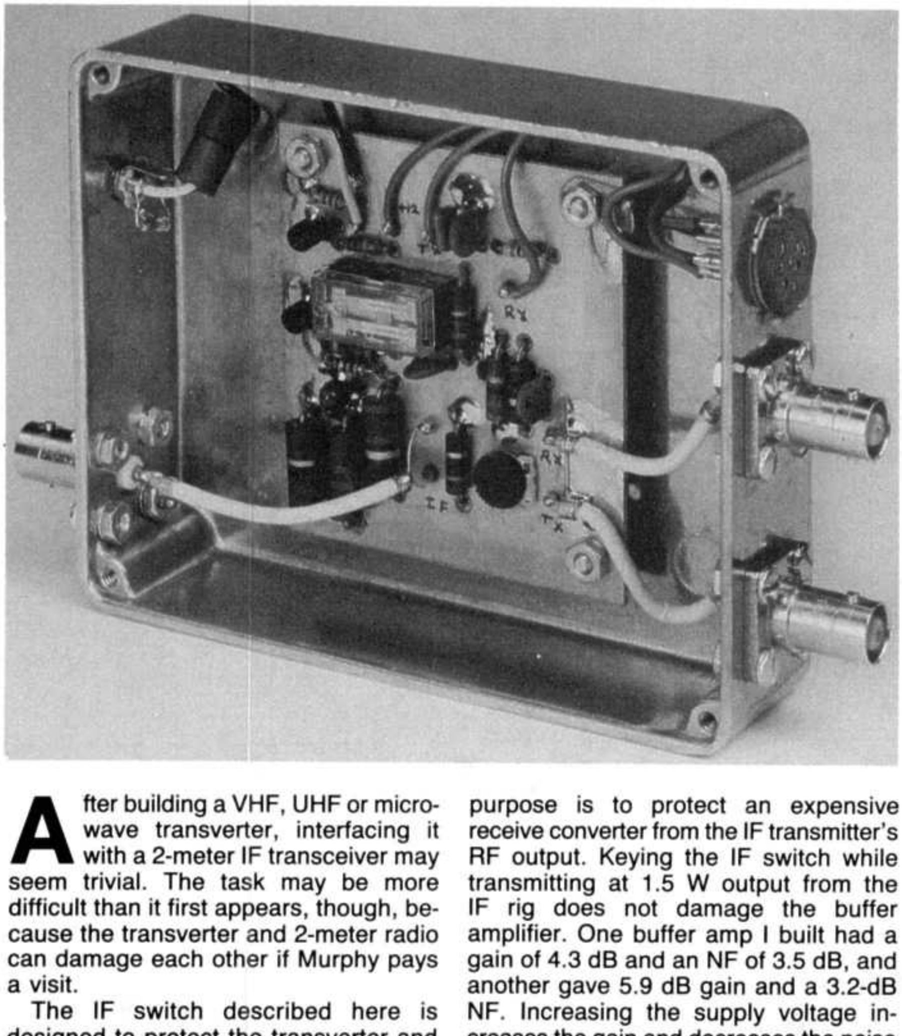
Phase IV geostationary orbit (GSO) satellites are the next major AMSAT design project. There are numerous technical and engineering obstacles to be overcome with a GSO, but development work has been initiated and is proceeding at a rapid pace.

Phase IV development plans currently detail use of the 13- and 23-cm bands for a national high-speed digital communications network and multichannel voice repeater linking. Other modes will follow the current OSCAR assignments.

The soul of AO-13 contains all the science, engineering, disappointments, nail-biting and finger-crossing of a great many AMSAT members, engineers and scientists plus the aggregate hope for the future of Amateur Radio space communications. Progress does not stop when the transponders are turned on. As usual, space development will have its associated spinoff effect. Development of commercially available 13-cm radios and antennas are needed; new thinking in terms of how "semi-intelligent" digital communications net-works can be established and efficiently managed; simple earth-station designs will be required to extend the capabilities of space communications to the average ham: The list lengthens as hams expand their horizons in space. All we in the Amateur Radio community need do is emulate the superb efforts of the AMSAT-DL Phase 3C team by contributing what we can to the development of new satellites and associated satellite technology. Glad to have you with us, AO-13, and hearty congratulations to AMSAT, and in particular to AMSAT-DL, for a job well done! —*NM1Q*

A VHF/UHF/Microwave Transverter IF Switch

By Zack Lau, KH6CP
ARRL Lab Engineer



After building a VHF, UHF or microwave transverter, interfacing it with a 2-meter IF transceiver may seem trivial. The task may be more difficult than it first appears, though, because the transverter and 2-meter radio can damage each other if Murphy pays a visit.

The IF switch described here is designed to protect the transverter and transceiver while doing the necessary TR switching. See Fig 1. Using the voltage on the keying line of the IF transceiver, K1B switches the IF rig's antenna line between the transmit and receive terminals of the transverter, and K1C switches 12 V dc. An adjustable attenuator is provided on the transmitter input side of the transverter, and a protective buffer amplifier is provided on the receive side. I selected the attenuator-resistor values in the IF switch to limit the 2-W drive signal from my ICOM IC-202 144-MHz transceiver to the −4 to 0 dBm of RF drive required by my 903-MHz transverter.[1]

The buffer amplifier, U1, does not offer great noise figure (NF) or gain. Its main purpose is to protect an expensive receive converter from the IF transmitter's RF output. Keying the IF switch while transmitting at 1.5 W output from the IF rig does not damage the buffer amplifier. One buffer amp I built had a gain of 4.3 dB and an NF of 3.5 dB, and another gave 5.9 dB gain and a 3.2-dB NF. Increasing the supply voltage increases the gain and decreases the noise figure of the amplifier, but I didn't adjust U1's bias for the higher supply voltage, as this would detract from the ruggedness and reliability of the buffer amp.

Understanding the relay driver circuit requires some insight into how transceivers and transverters work. Most solid-state transceivers use pull-up resistors to provide a positive voltage that is brought to ground to put the rig in transmit. Instead of keying the rig through a current source in the keying line (as is done in my LMW 1296-MHz transverter), why not key the system with a voltage? The advantage of voltage keying is that the system can be made fail safe—the 1-MΩ resistor, Zener diode and FET input on the keying line can't damage the radio. Current-source keying must be carefully designed for compatibility with the IF rig's keying circuit to avoid damage to the rig. Voltage keying is safer than current keying for another reason: If the connection to the rig's keying line falls off, the system is left in the (sometimes safer) transmit mode. D2 protects the FET from static discharges.

[1]Resistor values for 50-Ω attenuators are given on p 25-44 of *The 1988 ARRL Handbook*.

Construction

Fig 2 shows the PC board for the IF switch. The component side is a ground plane. After drilling the holes for the component leads, use a 1/8-inch drill to remove the foil from the holes on the component side of the board. Drill a hole through the PC board at the center of U1's mounting location to provide clearance for the body of the device, and to allow one of the ground leads of the MMIC to pass through the board. One of the MMIC's two ground leads is bent and passed through a hole in the PC board under the MMIC, so that it can be soldered to the ground-plane side of the board. (The other ground lead is soldered to the circuit ground on the trace side of the board.) Soldering one ground lead to each side of the board provides good performance: wrap-around foils aren't necessary. Fig 3 shows component placement on the PC board.

I used carbon-composition resistors at R1-R3 in my first prototype IF switch. In another unit, metal-oxide resistors worked just as well at 2 meters (the input return loss at J1 is more than 20 dB). The ferrite beads on the keying line may not be needed; they are there to prevent RFI problems.

Testing

A power meter that can accurately measure power levels of −15 dBm, and that can handle accidental-overload power levels equal to the full output of your IF transceiver, is ideal for testing the IF switch. The maximum amount of power present at the receive side of the switch is −15 dBm. The power to the transmit converter is adjustable (by means of R7) between 0 and −15 dBm (for 1.5 W of drive). This represents attenuation that's adjustable from 32 to 47 dB.

If you are wondering why I didn't just use a TR sequencer to switch the IF rig and the transverter, consider the wear that CW operation can have on those expensive microwave relays! (Using my ICOM IC-202 as an IF rig, I have to flip a TR switch during CW operation anyway!)

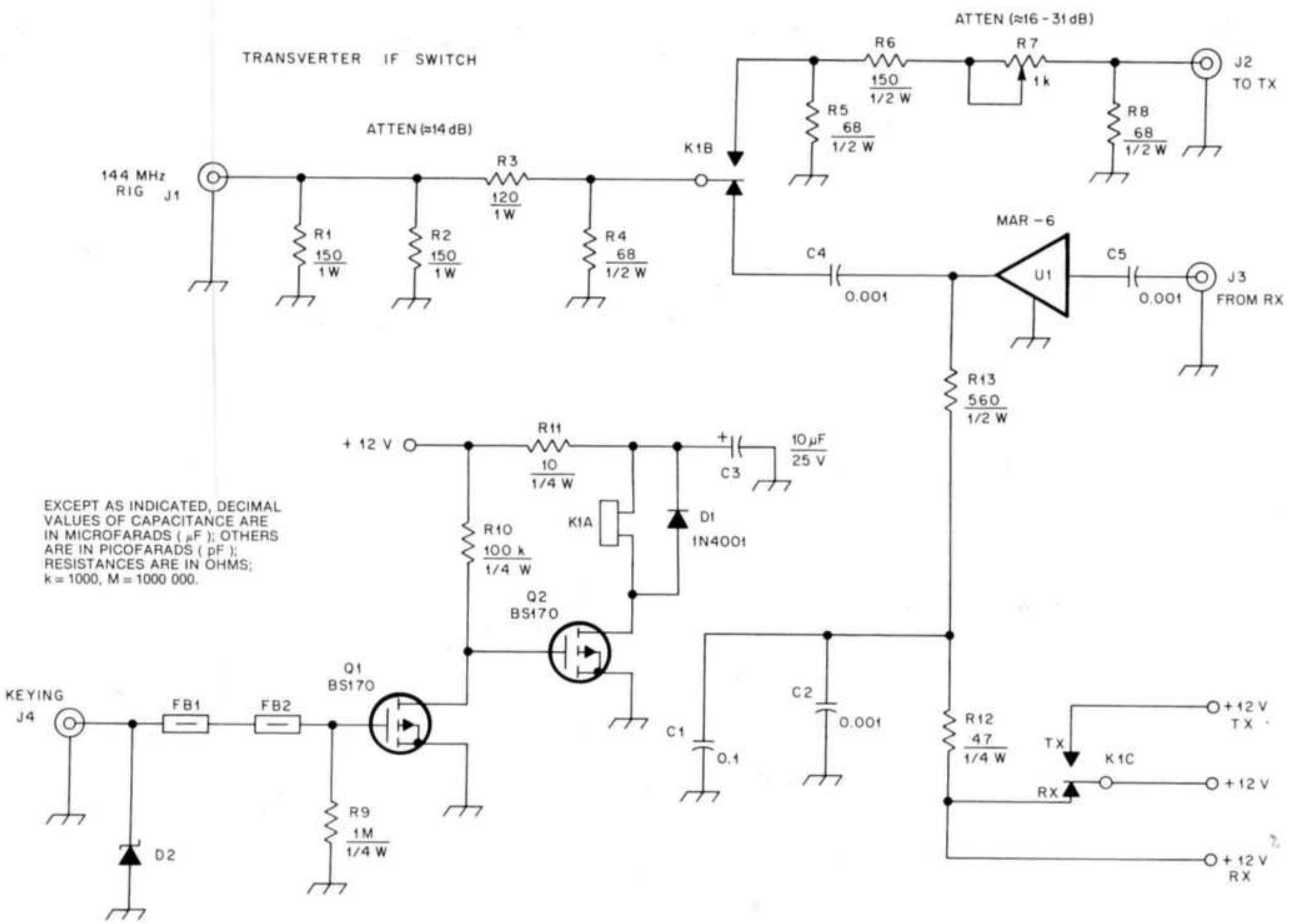


Fig 1—Schematic of the transverter IF switch. D1—1N4001 or equiv. D2—20- to 33-V, ½-W Zener diode. FB1—FB-43-801 (see text). FB2—FB-64-801 (see text). J1-J3—BNC female. K1—DPDT relay (RS 275-213). Q1, Q2—BS170 FET (available from Digi-Key, or Radio Shack® 276-2074). U1—MAR-6 or MSA 0685 MMIC.

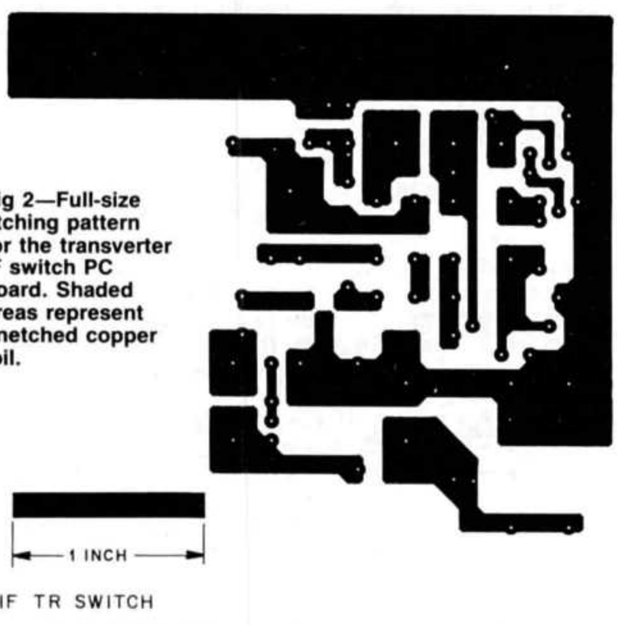


Fig 2—Full-size etching pattern for the transverter IF switch PC board. Shaded areas represent unetched copper foil.

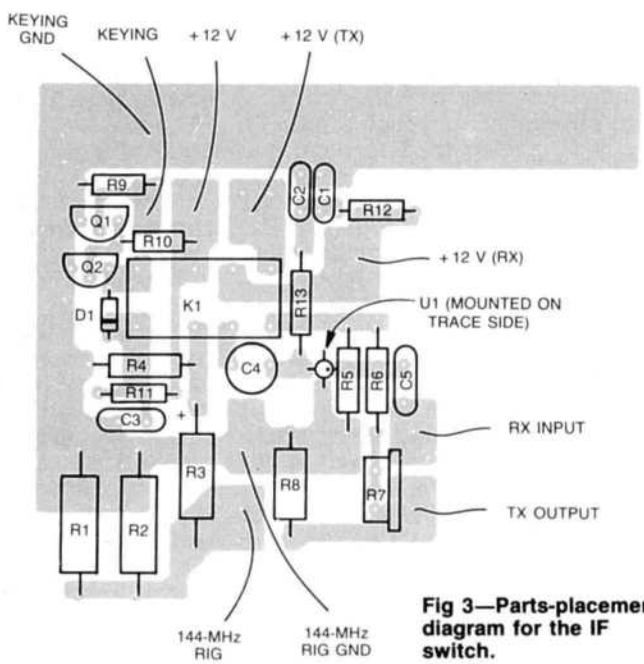


Fig 3—Parts-placement diagram for the IF switch.

High-Voltage Breakdown Tester

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An adjustable, current-limited, high voltage breakdown tester, referred to in this article as HVBT, is indispensable for testing high-voltage devices such as vacuum relays and vacuum capacitors. An HVBT facilitates testing other devices for peak inverse voltage (silicon rectifiers), collector-emitter and/or drain-source breakdown voltage (transistors) and for the presence of gas in transmitting vacuum tubes. Each of these tests can be performed accurately and *non-destructively* by using the simple circuit described in this article.

The Circuit

The HVBT circuit, shown in Fig 1, consists of an adjustable transformer, T1, that delivers 0 to 132 V ac to the primary of a 10 kV (rms) transformer, T2. A 150 W, incandescent lamp, DS1, is used in series with T2's primary to limit current in the event that the output of the supply is shorted. A full-wave voltage-doubler circuit (D1, D2, C1 and C2) delivers a maximum no-load output voltage not greater than plus and minus 15 kV with respect to ground. This gives a maximum voltage testing capability of about 30 kV if the device being tested can be isolated from ground, or ± 15 kV if it cannot.

The High-Voltage Power Supply

T2 is the heart of the HVBT power supply. The 10 kV (rms) transformer that is used in the HVBT power supply is a new, high-quality surplus unit that costs approximately $20 and is obtainable from Fair Radio Sales.[1] This transformer has a Hypersil® tape-wound, grain-oriented silicon-steel core. The windings are epoxy encapsulated. The secondary resistance of this transformer is about 39 kΩ. I estimated the current capability of this transformer as follows: The safe load resistance for any full-wave voltage-doubler circuit is about 300 times the secondary resistance of the transformer. In this case, 39 kΩ × 300 = 11.7 MΩ. T2's output voltage, under this load, will be about 26 kV (2.6 × 10,000 V). The maximum load current of 2.2 mA can be determined by using Ohm's Law (I = E/R or 26 kV/11.7 MΩ

[1]Notes appear on page 7.

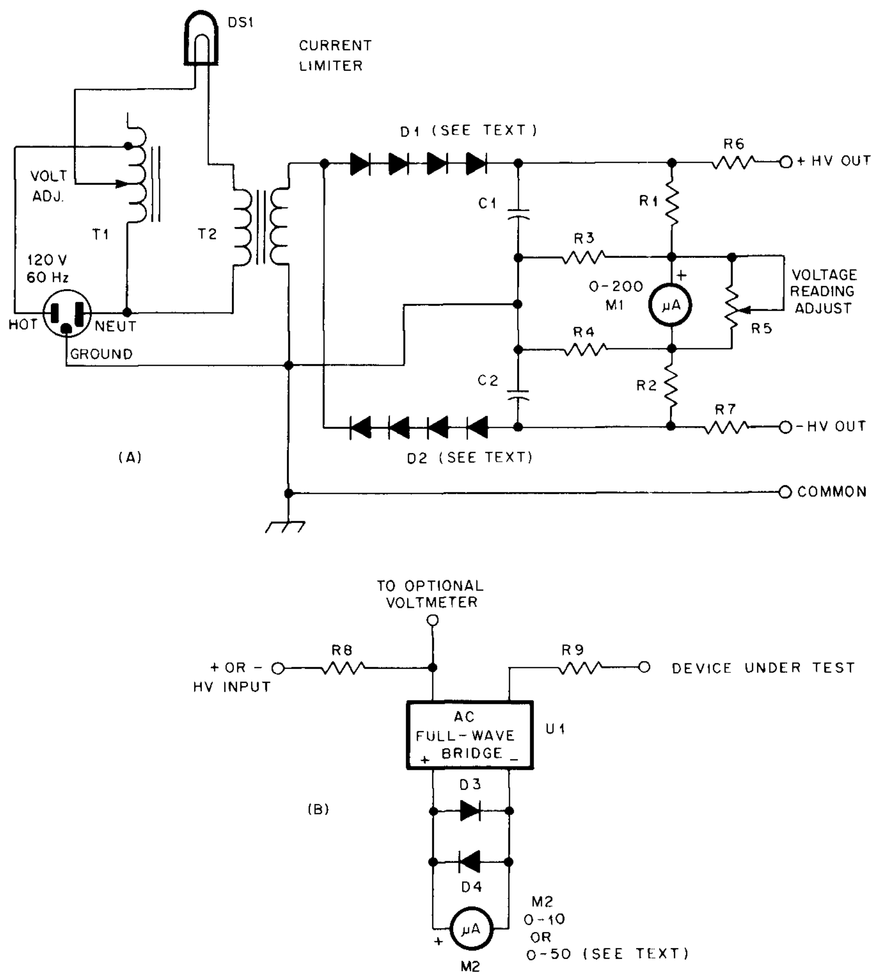

Fig 1— The high-voltage breakdown tester schematic. D1 and D2 are series connected diodes or HV rectifiers as discussed in the text. At B: Optional current limiter/detector for the HVBT — see text.

C1, C2—0.01 μF to 0.1 μF, 10-k WVDC or 15-k peak voltage rating. See text.
D1, D2—At least 35 k PIV (total) at at least 25 mA. Several CRT HV rectifiers in series will work well.
D4, D5—50 PIV, 1 A.
M1—200 μA, 4½-inch meter. Simpson no. 17593, Triplett no. 152-2219 or equivalent. Custom 0 to 30 kV scale.
M2—10 μA to 50 μA. A suitable meter can be obtained from Radio Shack®. See text.
R1, R2—55 MΩ, 5 W (Dale DC-5).
R3, R4—10 kΩ, ¼ W.
R5—5-kΩ, ¼-W multiturn trimmer.
R6, R7—100 to 500-Ω, 10-W wirewound.
R8—5 MΩ, 50 W. Difficult to find. See text.
R9—100 to 200-kΩ, 2 W.
T1—Variable autotransformer, 120-V input at 1.25 A minimum, 0 to 132 V output. See text.
T2—115-V primary, 10-kV (rms) secondary, 60 Hz. Secondary resistance should not be greater than 40 kΩ. (AMP PIN 845321-3)
U1—50-PIV, 1-A full-wave bridge rectifier.

= 2.2 mA). The total circuit current requirement is less than 0.3 mA.

Construction Notes

Because of the unusually high voltage level in the HVBT circuit, metal is not the ideal material for the HVBT cabinet. Polyurethane-varnished plywood or polyester-fiberglass board would be a good choice. Although I do not recommend it, a metal cabinet with an insulated front panel could also be used.

To minimize corona leakage, all of the HVBT solder connections that are at high potential should be formed into smooth, rounded blobs by applying large amounts of solder. Corona is very easy to check for: Adjust T1 for a supply output voltage of 30 kV, turn out the lights and watch for a blue discharge accompanied by a hissing sound and the odor of ozone. Any solder connections that contain sharp points corona.

The current-limiter detector is usually built into the same cabinet as the HV power supply. It can be built on a separate wood or plastic assembly. Care should be taken to ensure that M1, which is very near ground potential, is not placed close to M2 (the current-leakage-detector meter) which can be ±15 kV with respect to ground. The current limiter can be permanently wired to either HV polarity. It can also be wired to banana jacks on the front panel allowing positive or negative polarity connection. Note: Do not touch M2's face while the output voltage is above 2 kV. Even if you use a plastic insulated meter at M2, the meter insulation will not protect you against anything like the maximum output voltage in this circuit. M2 and R8 must be mounted on an insulated panel with no grounded conductors within 50 mm (2 inches) of either component.

D1 and D2 can be made from a string of diodes (greater than 1 kV) or from a series of multiple diode HV, low-current rectifiers like those used in CRT HV power supplies. Equalizing resistors and equalizing capacitors should not be connected across the diodes unless you plan to mix diodes with different junction capacitances, such as 6-A units in series with 1-A units. If you are using all 1-A or all 6-A rectifiers in series, equalizing resistors/capacitors are not only unnecessary, but ill-advised, as the typical ½-W resistor frequently used to equalize 1-kV rectifiers has a maximum voltage rating of only 250 V. What usually damages rectifiers is reverse current. In a string of series diodes, the reverse current will always be exactly the same in each component. Obviously, reverse current that is equal in each series diode does not require equalization!

Testing Components

The basic technique behind non-destructive voltage testing of components or devices is simple: Slowly increase the applied voltage across the device being tested until 1 to 2 μA of leakage current flows through the device. *Stop increasing voltage at that point.* CAUTION: If you continue to increase voltage beyond this point, current and power/heat will also increase and the part will probably be damaged. Do not apply voltage to the device under test any longer than necessary to read the voltmeter, M1, and the current leakage meter, M2.

Two voltage characteristics can be recorded, the *breakdown voltage* (the point at which 1 to 2 μA of leakage current flows) or the *withstanding voltage* (the point at which the breakdown voltage is decreased until no detectable current flows). There isn't much difference between the two, but be sure to note which one of the two was recorded.

Apply only half of the HVBT's 30-kV output to devices that you expect to exhibit a breakdown voltage of less than 15 kV. To do this, connect one end of the device to common. The voltage reading on M2 must be divided by 2. If the breakdown voltage of the device is expected to be less than 1 kV, it's a good idea to connect a voltmeter of appropriate test scale value from the output side of R8 to common so that the test voltage can be read more accurately without causing a false leakage current indication on M2.

Vacuum capacitors and vacuum relays should always be tested before they are installed in an amplifier or antenna tuner. I have seen brand-new devices—direct from the factory—that were defective because of tiny air leaks. A defective glass-envelope vacuum relay can be easily identified because ionized air emits a blue light visible through the glass. Ionization in a defective glass-envelope vacuum capacitor, however, is usually deep inside the capacitor's concentric meshed plates, and cannot be seen. (Note: Vacuum-variable capacitors should be tested with the plates fully meshed.)

When a vacuum capacitor fails in an amplifier containing more than one vacuum capacitor, it is very difficult to determine which one is defective without the help of a breakdown tester. (The only effective way to isolate the faulty capacitor is to replace the capacitors, one at a time, until the defective unit is found.) An HVBT is more important when the amplifier has only *one* vacuum capacitor. In one situation that I know of, an amplifier had a below-normal power output with high anode current, indicating that the vacuum capacitor was bad. The owner replaced the capacitor with a fresh unit and the problem persisted. This implied that the problem couldn't be in the vacuum capacitor, because it had just been replaced. When the owner asked me for suggestions, I asked if the new capacitor had been tested for current leakage. To shorten the story: Both capacitors were bad; the third vacuum capacitor we tested was good. A high-voltage breakdown tester would have saved a lot of time and effort in this case.

Vacuum capacitors that have been in storage for a long time may develop "whiskers." These whiskers are microscopic filaments of metal that appear to grow from the surface of the copper. This unexplained anomaly causes the capacitor's breakdown voltage to be lower than normal. It's possible, however, to burn off the whiskers over a period of several minutes by continuing to apply the test voltage at the point where no more than 5 μA of leakage current is detected. If the leakage current goes to zero after one or two minutes, increase the test voltage to the same 5 μA limit. Leave the voltage at this level for a few minutes to see if the leakage current returns to near zero. If it does, repeat the process. If the leakage current stops decreasing at a certain voltage level, stop the test. Reduce the test voltage if you hear a persistent tinking sound from the capacitor. This sound is caused by flashover in the capacitor.

Semiconductor testing is very important when building HV power supplies that contain strings of series connected silicon rectifiers. The failure of one diode can cause the failure of the other diodes in that string because of the domino effect. Diode failure can deliver ac to filter capacitors: polarized electrolytics, if subjected to this, can blow their gas vents or explode violently. Diode failure in a full-wave voltage doubler can cause similar destruction: The doubler capacitor filter not receiving current from the open diode will have reverse current forced through it by the still-functioning half of the voltage doubler circuit. Bottom line: A defective 29-cent diode, which can destroy many dollars worth of good diodes and electrolytics right in the middle of a contest, can be discovered with an HVBT *before* a project is constructed.

Testing silicon rectifiers is straightforward. Increase the reverse voltage applied to the rectifier until you detect a leakage-current of 1 to 2 μA (high current diodes can withstand more reverse current). At this point, observe the leakage-current meter. If the leakage current fluctuates erratically without voltage adjustment, the diode is defective and should be discarded. Because it is difficult to mark each good diode for leakage current, I usually sort them into labeled boxes. One box may contain diodes that test between 600 V and 799 V, another box may contain the same type diodes that test from 800 V to 999 V, and so on.

The most common manufacturing

defect that causes a diode to open after prolonged use is a too-resistive wire bond to the diode's silicon wafer. This defect is now rare, but the test for it is simple: Pass dc equal to the diode's reverse current rating through the diode in the forward direction. Connect a DVM across the diode and measure the voltage drop. A diode with a too-resistive wire bond exhibits a higher than average forward voltage drop. Note: It's normal for a diode's forward voltage drop to decrease slowly as its junction temperature increases in response to forward current.

Transistors are now made with voltage capabilities similar to those of silicon rectifiers. Bipolar transistors with collector-to-emitter voltage ratings of 1.5 kV are now common, and 5-A FETs are available with 1-kV source-to-drain ratings. Testing these devices is similar to testing silicon rectifiers except that a resistor of no more than 100 Ω should be connected from the base to the emitter (or from the gate to the source of a FET) before testing. When testing small HV transistors, keep the leakage current below 1 μA.

Air-dielectric capacitors can have internal support-insulator leakage that is easily detectable with a breakdown-voltage tester. An HVBT can also be used as an aid in aligning the plates of air-dielectric capacitor plates. Use the breakdown tester to identify misaligned plates by applying voltage and watching for arcing between capacitor plates in a darkened room.

Amplifier transmitting tubes must have a near perfect vacuum in order to function properly. The leakage current/vacuum test is made *without* applying filament/heater voltage to the tube. Apply test voltage between the anode and the cathode. A healthy vacuum will exhibit less than 5 μA of current leakage at twice the tube's rated anode voltage. Twice the rated anode voltage may sound severe, but this is actually what the tube experiences when operated in an RF amplifier (when the tube's anode current is cut off during the positive swing of the tank circuit voltage, its peak anode voltage is very close to twice the dc voltage supplied to the circuit). In plate-(anode) modulated RF amplifier service, the peak anode voltage is *four times* the anode's dc supply voltage!

Circuit Alterations

You can make changes to the HVBT circuit to accommodate available parts, or scale the voltage capability of the tester up or down. For example, R1 and R2 are 55 MΩ, 5 W units available as new surplus and they cost approximately one dollar each from Fair Radio Sales. These resistors, Dale® model DC-5, are made with a long, resistive deposited film that is made to handle high voltage. If you'd like to use a different value resistor for R1/R2, I recommend that you go higher in value rather than lower, because the power dissipated in R1 / R2 is already considerable at 55 MΩ. 100-MΩ resistors are common; resistors of this value will work well provided that M1 has a full scale sensitivity of 100 μA and that the manufacturer rates the resistors you use for 15 kV each.

It's important to note that resistors have a dissipation rating *and a voltage rating,* and that the voltage rating usually takes precedence over the dissipation rating. For example: A 100-MΩ, 2-W 5% tolerance carbon resistor has a dissipation rating which might lead you to believe that the resistor can safely dissipate 2 W. Using Ohm's Law $E = \sqrt{(PR)}$ or $\sqrt{(2W \times 10\ M\Omega)} = 4472$ V at 2 W. This may seem safe, however, the resistor is limited to 500V maximum. Using $P = E^2/R$ ([500 V × 500 V] ÷ 10 MΩ) the practical dissipation for this resistor at 500 V is 0.025 W (1/40 W). Be sure to read the fine print when purchasing resistors!

R8

Probably the hardest part to find for the HVBT project is the HV current limiter resistor, R8 (5 MΩ, 50 W). The actual value of R8 is not critical, and any value between 2 MΩ and 6 MΩ will work well provided that the resistor can handle sudden overloads. If you adjust the HVBT's voltage control carefully, the actual voltage drop that appears across R8 will seldom amount to more than 20 V. R8's voltage rating is put to the test when the device being tested shorts out. This causes the voltage across R8 to briefly rise to as much as 25 kV until the C1 and C2 discharge and DS1 begins to limit the current.

If you cannot locate a single resistor for R8, there is a commonly available spiral-film resistor that has the right properties. These Sprague Q-line™ resistors have a tolerance of 2% and a dissipation rating of 2 W. You can wire 25 200-kΩ Q-line resistors in series on perfboard to make R8. A single Dale DC-5 resistor can be used at R9, which isolates the charged capacitance of M2 from the device under test.

C1 and C2

These capacitors are occasionally subjected to 15 kV in this circuit, so capacitors with a 15 k WVDC rating would seem to be necessary. Manufacturers use different rating methods and different test voltages, though. It's possible to use 10 k WVDC capacitors in place of the much harder to find 15 k peak voltage capacitors. An example of this would be Sprague's 600-V Vitamin-Q™ capacitor,[2] which will not fail until no more than six times its working voltage (3600 V) has been applied to it (short term) at room temperature. This is partly due to the fact that the capacitors are rated for continuous-duty, high-ripple current, high temperature operation at their working voltage rating. Sprague will probably disagree with me, but I believe that for low-ripple-current service, in a breakdown tester operated at normal room temperatures, for example, this type of capacitor can be used at 1.5 times its rated working voltage with no difficulty. (I have been using a pair of 10 k WVDC Sprague capacitors in my breakdown tester for years with no problems.) Remember that not all capacitor manufacturers build and rate their products so conservatively. However, if you have a pair of 10 k WVDC capacitors that you would like to use in this project, they are certainly worth a try. The HVBT circuit has built-in current limiting so if C1 or C2 shorts out no serious damage will occur. If a capacitor does short, you will hear a tick and DS1 will light.

If you use tubular plastic dielectric capacitors with screw terminals, put a stop nut between the capacitor and the mounting surface. The stop nut keeps the screw, which is usually soft-soldered to the end cap, from being pulled out of the capacitor when the fastening nut is tightened. Four nuts (usually no. 8-32) are required to properly mount each capacitor.

Operating Notes

Take reasonable care when operating this equipment. Although the steady current capability of the tester is quite limited, the charged filter capacitors can deliver a substantial shock to a careless operator. After you turn down the test voltage at the conclusion of a test, observe the voltage indicated by M2. Wait until the filter capacitors are fairly well discharged before disconnecting the device under test or reaching into the tester.

When testing with voltages above 8 kV, it is possible to have a false leakage current indication caused by sharp points on the leads between the HVBT and the device under test. These sharp points cause corona discharge, which draws current through M2. To check for this condition, simply disconnect the device under test (remember: safety first!!) but leave the test leads connected to the HVBT. Next, increase the voltage to the level at which corona was suspected. If leakage occurs, M2 will indicate how much. A repeat performance in a darkened room will indicate where to start smoothing things up with a file.

NOTES

[1]Fair Radio Sales, 1016 E Eureka St, PO Box 1105, Lima, OH 45802.

[2]The Vitamin-Q capacitor has a high purity mineral oil dielectric that does not contain PCBs. The WVDC rating for oil-filled or plastic-dielectric capacitors is usually about six times the peak volt (PV) rating. PV ratings are only for intermittent duty with low ripple current. This does not apply to electrolytic capacitors.

Automatic Antenna Controllers

By Peter Prendergast, MD, KC2PH
828 Westwood Drive
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In the 1960s, one justification for putting a man on the moon was that the technology developed towards that end could be applied to everyday living. In the 1980s, the same logic holds true for the amateur satellite program. The technology that has, in recent times, been applied only to satellites can now be used for many applications in the average ham shack. A good example of this is the automatic antenna-tracking interface/controller. These devices have many uses beyond satellite tracking. On a daily basis in my own station, I use a controller to assist in delivery of mail via packet radio, provide distant stations with access to my PBBS, and (of course) to track satellites. The uses of a controller are limited only by your imagination.

Listed and discussed below are six commonly available systems. Each has a unique place in the amateur environment. I have reviewed them all, and the key characteristics of each that are useful to know before purchasing a system are outlined.

The Kansas City Tracker

The KCT is best suited for the individual who has an IBM® PC (or clone) and Kenpro KR-5400A/5600A rotators. In my opinion, this system represents the "Rolls Royce" of controller/trackers for use with this hardware setup. All circuitry is mounted on a high-quality half-size, plug-in PC board. The KCT is bus driven, and its defaults can be set from the keyboard, which makes installation a snap. The software to run the tracker is RAM resident. You can load as many future satellite passes as will fit in the available memory: The control system quietly positions the antenna for you at the appropriate times. Having used this controller to follow FO-12, the best way to describe the effectiveness of the system is *breathtaking!* A few other notes about the KCT: The software does not inhibit the performance of your computer, and is very user friendly. More than one of these boards can be installed in the system simultaneously. Price class: $160, including software. Contact Brooks Vanpelt, KB2CST, 41 Acadia Dr, Voorhees, NJ 08043, tel 609-751-1018 for more information.

The ARRL Automatic Antenna Controller

The ARRL automatic rotator controller is a stand-alone smart controller.[1] It does not require connection to a computer during operation. Using an on-board microcomputer, the system is capable of being loaded (by means of an external computer) with 1096 antenna-aiming points. Depending on the satellite, that can mean as much as several weeks of antenna-position control. Once the aiming points are loaded, the controller can be unhooked from the computer and used until the antenna-pointing data is exhausted. The controller can also be operated by means of a built-in keypad and LED display, but antenna-aiming information cannot be saved in memory using this method. RS-232-C and TTL serial ports are available. The TTL serial port gives even C64 users access to the storage capabilities of this highly sophisticated controller. The ARRL Automatic Antenna Controller comes as a kit that is easy to assemble.

I use the ARRL controller in my station, and have found it to be highly accurate and handy for DX work. This system is ideal for those who have huge antenna systems, because it allows you to provide external power supplies for your rotators. (Those with large EME arrays will really appreciate this one!) Software to calculate positioning points for this system is available from AMSAT, PO Box 27, Washington, DC 20044, tel 301-589-6062 (order Quiktrak by N4HY), or from Silicon Solutions, Inc, PO Box 742546, Houston, TX 77274-2546, tel 713-661-8727 (order Graftrak). ARRL controller price class: $220. Distributed by A & A Engineering, 2521 W La Palma, Unit K, Anaheim, CA 92801, tel 704-952-2114.

Mirage Tracking Interface

This control system is a joy to use. Complete installation takes all of about five minutes. To use the MTI, you'll need an IBM PC (or clone) with at least one serial port. The software provided is a version of Graftrak. With the MTI and this software, you have the ability to control your antenna, enjoy Graftrak's spectacular graphics, and control the radio's frequency while tracking just about anything in the sky. The MTI is somewhere between a smart and a dumb controller. It does require that the computer be connected, but using Quicktrak from AMSAT allows you to control antenna position *and* use another serial port at the same time. Graftrak, by comparison, allows antenna and radio control, but does not allow you to use the computer to run other programs at the same time.

For the Mode JA enthusiast, this system is very handy; when using either Graftrak or Quicktrak with the MTI, not only is the antenna positioned, but Doppler shift is compensated also. Price class: $550. For more information, contact Mirage/KLM Inc, PO Box 1000, Morgan Hill, CA 95037, tel 800-538-2140 (outside California) or 408-779-7363.

Phase IV Systems Controller for the C64

This system consists of a plug-in card for the C64 that controls any type of rotator. The package includes a very clear instruction manual and software. This controller isn't as precise as most other systems, but it does come within 5° or so. (Considering the fact that most antenna systems have a 3-dB beamwidth in excess of 20°, an error of 5° is generally unimportant.) The Phase IV system does not use position-sense feedback from your rotator, but instead calculates antenna position by the length of time for which the rotator has been energized. There are some potential problems with this system: If your rotator were to stall or bind, the system may damage the rotator and/or lose calibration by attempting to continue turning the rotator. In other respects, the Phase IV system is cleverly designed, cost effective and easy to use. Price class: $149. Contact Phase IV Systems, Inc, 3405 Triano Blvd, Huntsville, AL 35805-4695, tel 205-535-2100 for more information.

Encomm KR-001 Controller

This system provides another method

(continued on page 14)

[1]Notes appear on page 14.

Correspondence

Dodd/Lloyd Program Corrections

Since "Measurement of Antenna Impedance" appeared in the November 1987 issue of *QEX*,[1] considerable discussion has taken place and been published. The following briefly addresses some of the criticism.

The main concern was the setting of ER(B) to 50. It was never used to estimate SWR as some have suggested. Measurement of SWR, if required, can be done with a SWR meter. I guess I have been conditioned by the W8CGD system of extracting results using graphics directly (in ohms), as described in the original article. I feel, however, that ER(B) has to be set to something, therefore I standardized the voltage across ER(B) to 50 (5 volts) for the following reasons. (The modified program supplied by Steve Lund, WA8LLY, is useful if problems are encountered setting ER(B) to 50.)

(1) So that the lowest voltage readings will remain in the linear portion of the RF voltmeter curve. (Set ER(B) to .05 volts and see the difference!) Tom Lloyd comments: "The real part of the solution is a function of EA(A), ER(B) and ECZ(C) only. Any change in the values of EC(D) or EZ(E) will not affect the value of RES. Similarly, the imaginary part of the solution is a function of ECZ(C), EC(D) and EZ(E) only. This means that *j*X does not respond to changes in the values of EA(A) and/or ER(B)".

(2) If ER(B) is not set to 50, then it must be entered into the program during run time. This is not an overly complex problem, but it creates more work when entering large amounts of data. Tom Lloyd comments: "This is the result of the points raised above. The method takes no account of a 'cocked hat' situation where some mean value needs to be evaluated. I do not need to tell you that a condition of no cocked hat is a rarity, albeit a welcome one. In other words, the method assumes that all the readings taken are accurate and does not make allowance for errors. These algebraic equations can be misleading unless one is very confident of the accuracy of the data used."

(3) I have planned to use this program to read the results directly to a BBC computer fitted with a PD7002C A/D chip. The PD7002C chip is a four channel device so one of the readings will have to be standardized. Using an analogue/digital converter such as the PD7002C would reduce power drift errors because all parameters would be measured very quickly. At the same time it would remove any human error that may occur when the data is transferred from the voltmeter.

Tom Lloyd also produced an analysis of the effect of inaccurate data using the trig and algebraic methods. His comments follow:

"If we process faulty data by drawing a graph then a cocked hat will usually be produced. It is an accepted practice to take a solution to lie at the 'center of gravity' of this area.

"The trig method uses this principle to find the solution in such cases. If the size of the cocked hat is small, then its sides, which are formed from arcs, can be approximated as straight lines. The trig method assumes them to be so in order to find the position of the impedance plot and finding the errors in its position resulting from errors in the data. If the data errors are large, then a large cocked hat will result and the approximation to straight lines will no longer be valid. The calculated impedance will not be exactly at the center of gravity of the cocked hat. In cases where the trig method leads to faulty solutions, the errors are rarely gross and are of the right order. These results, although in error, will still provide useful information.

"To summarize: If the supplied data are ideal, then the two results will produce identical solutions. When the data are not ideal, the errors in the algebraic solutions are greater than those from the trig solutions. However, in such cases the trig solutions indicate their reliability; no such indications are given by the algebraic solutions."

Tom and I received a letter from a *QEX* reader in Seattle, Washington, enclosing a diagram which showed that the TOM program would give an erroneous result if the negative reactive component was greater than the resistive component, ie, S< >P-Q. (Note: The reader did not include his name and address—we would like to thank him for his contribution.) Tom has produced a line of code to correct this:

$$\text{IF } (L+Q) > (PI/2) \text{ then } S = (2*PI) - (P+Q) \qquad \text{(Eq 1)}$$

This line should be inserted as line 175 into TOMS, line 313 of TOM and line 435 of TOMFD. Note that a PI symbol is used with the PB110 calculator instead of the PI function.

In conclusion, our objective in writing this article was to provide a practical and simple method of measuring antenna impedance. We realize that simplicity and accuracy are relative in this context. We are hoping to improve the design of the impedance box, using information supplied by WD8KBW, to see if this technique can be extended to VHF. —*Peter Dodd, G3LDO, 37 The Ridings, East Preston, West Sussex, BN16 2TW England, United Kingdom.*

More Impedance-Measuring Hybrid Comments

Neither the original article (November 1987) or the subsequent discussion published in *QEX*[2] cover some important points about the W8CGD method and the Dodd/Lloyd computer program.

Two of the most important points are as follows: First, the Dodd/Lloyd computer program contains a glitch that causes it to return an incorrect answer 30% of the time based on the examples provided by Dodd/Lloyd. This appears to result from not considering the multiple values of the inverse trig functions. Second, the method can be extremely sensitive to errors in experimental measurements. This sensitivity is in part dependent on the choice of the reference reactance (C_6 in the article, X_D in the following discussion).

The five measurements obtained with the W8CGD impedance measuring circuit each define an equation relating the four unknowns:

$$A^2 = |i|^2 [(R_L + R_B) + (X_L + X_D)^2] \qquad \text{(Eq 2)}$$

$$B^2 = |i|^2 R_B^2 \qquad \text{(Eq 3)}$$

$$C^2 = |i|^2 [R_L^2 + (X_L + X_D)^2] \qquad \text{(Eq 4)}$$

$$D^2 = |i|^2 X_D^2 \qquad \text{(Eq 5)}$$

$$E^2 = |i|^2 (R_L^2 + X_L^2) \qquad \text{(Eq 6)}$$

where $|i|$ is the magnitude of the unknown current through the circuit.

Five equations in four unknowns, $|i|$, X_D, R_L, and X_L, can be solved in five different ways, each using a unique combination of four of the measurements. As the equations are generally quadratic, in general two sets of solutions will be obtained from each quad of equations. Provided the reference reactance, X_D, (C_6 in the article), is properly chosen, the correct set from each solution can be selected by: (1) requiring that the unknown resistance be positive; (2) picking R_L and

[1]Notes appear on page 10.

X_L from a set having a single, signed value of X_L; and (3) selecting from the remaining sets according to the magnitudes and signs as determined from the first two selections.

If the equations were exact, with no error in the measurements or in the value of the reference resistor, and no stray impedances or admittances, a unique solution for every quad of equations would be obtained. In practice, however, defects are always present in greater or lesser degree and different values for the unknowns will be obtained from each quad. Sometimes more than four solutions will be found clustered together. This can make selection of the correct solution difficult. In this respect, the data reduction program[3] sometimes selects an incorrect solution, causing large errors in the calculated impedance. For example, such an error occurred in the first three lines of Table 1 and the first four lines of Table 3 of Reference 1.[4]

Dodd/Lloyd calculate three of the five possible solutions. The following table gives the five correct solutions for the first line of data in Table 1 of Reference 1.[5]

R_L	X_L	$\|Z_L\|$	X_D	i	Data Set
48.16	−106.63	117.00	−47.00	0.1000	BCDE
52.74	−104.44	117.00	−47.00	0.1000	ABDE
50.68	−105.82	117.33	−47.00	0.1000	ABCD
46.96	−103.99	114.10	−45.83	0.1026	ADCD
50.68	−105.45	117.00	−47.36	0.1000	ABCE
49.85	−105.26	116.47	−46.84	0.1005	(Average values)

These average values are not greatly different from those obtained by Dodd/Lloyd[6] (after error correction). This is because the data set is remarkably consistent, approaching the case of mathematical exactness where any solution procedure returns the same values for the unknowns. This will not be true in general, and larger differences between the average of all five solutions and the average of a subset can be expected.

The accuracy achieved by using this method of impedance measurement can be good, but it requires careful construction and use. For example, the measured voltages appear in the solutions generally as squares. When an experimental measurement is squared, the uncertainty of the square is at least twice the uncertainty of the measurement. The solution process then takes differences between the squares of the measured voltages. These differences have still greater uncertainty than the squares. The result is that the uncertainty of the calculated impedance can be many times larger than the uncertainty (error) in the measured voltages. Error multiplication by a factor of 2 is about the best that can be expected. Error multiplication by factors of 5 or greater can be encountered; theoretically, the error multiplication can be infinite in unfavorable cases.

Error multiplication is largely a function of the relation between the reference impedances and the unknown impedance, and it can be minimized by careful selection of the reference impedances. The reference reactance, X_D, should have the opposite sign of the unknown reactance and should have a magnitude comparable to the unknown reactance. The reference resistance, R_B, is less critical, but it should not be larger than the reference reactance. If the reference reactance must have the same sign as the unknown reactance, its magnitude should be large compared with the unknown impedance. This conflicts with the need to measure the voltages accurately so a compromise is necessary.

When errors in the measured voltages are random, the solution averaging process tends to reduce the error in the calculated impedance. Averaging, however, does nothing to reduce errors caused by voltmeter calibration errors. The requirement of the voltmeters is that they be accurate relative to each other; absolute accuracy is not required, only that the ratio of any two voltages be accurate. Peak-reading diode voltmeters, at sufficiently high measured voltages, follow a relation of the form

$$E_{in} = k(E_{out} + f) \qquad \text{(Eq 7)}$$

where k and f are constants. Provided all voltmeters have the same constants (matched diodes) the correct ratio of two voltages read as E_1 and E_2 is

$$\frac{E_1 + f}{E_2 + f} \qquad \text{(Eq 8)}$$

If the additive correction is ignored, significant errors can be introduced—especially if the measured voltages are low. For voltmeters built with 1N34A diodes, f is about 0.01 to 0.15 V and the output voltage must exceed about 3 volts for the voltmeter to follow the above relation. Both k and f are larger when silicon diodes are used.

For the best accuracy, the voltmeters should be calibrated. This can be done at AF where a DMM can be used as a standard if the voltmeter capacitors are increased to a value suitable for the frequency used.

Ordinary oscilloscopes are not suitable for measuring the voltages because of the large shunt capacitance (typically 20 pF) of oscilloscope probes.

Calibration of the reference impedances is also helpful in achieving the best accuracy. This can be done by measuring a good resistive termination. If you are at all serious about measuring impedances, you should acquire the best termination you can. The accuracy of your measurements can never be better than that of your standard.

Construction practice for good accuracy requires assuring that no stray electric or magnetic fields couple to the voltmeter circuits, that there are no unmeasured impedances in the circuit and that stray capacitance to ground be minimized or compensated. The arrangement of ground connections must be carefully thought out to avoid mutual impedances in the measuring circuits. —*Albert Weller, WD8KBW, 1325 Cambridge Blvd, Columbus, OH 43212*

Notes

[1] P. Dodd and T. Lloyd, "Measurement of Antenna Impedance", *QEX*, Nov 1987, pp 6-9.
[2] P. Dodd, S. Lund and S. Jenkins, Correspondence, *QEX*, May 1988, pp 3-5.
[3] See note 1.
[4] See note 1.
[5] See note 1.
[6] See note 1.

Bits

AOR Miniature Mobile Scanner

AOR, Ltd of Tokyo, Japan has introduced a frequency-synthesized, miniature mobile scanner with keyboard control. The AR160 covers 29-52, 136-174 and 436-512 MHz. The radio weighs 25 oz and measures 1.5 × 4.62 × 6.5 inches (HWD).

Twenty keys on top of the receiver are used for frequency selection. The front panel supports LEDs that provide bright, read-at-a-glance information and serve to indicate frequency and status information. The squelch and volume controls, and six other control keys are also mounted on the front panel.

Each AR160 is supplied with a fused dc power cord, a telescopic whip antenna mobile-mounting bracket and hardware, and an ac-to-dc converter for convenient indoor use. Price: $189. For further information, contact Ace Communications Monitor Division, 10707 East 106th St, Indianapolis, IN 46256, tel 317-842-7115, FAX 317-849-8794. (Ace Communications is a wholly owned subsidiary of AOR.) —*Paul K. Pagel, N1FB*

>50
Focus on technology above 50 MHz

By Bill Olson, W3HQT
Box 2310, RR 1
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Electromagnetic Waves and Antenna Polarization

In this month's column, I'll begin a series on antenna fundamentals. As there has been a great deal of excellent antenna information in the literature lately, to avoid duplication of the work of others, I'll discuss some areas that haven't been covered or need some clarification. The increased interest in satellite communications following the launch of AO-13 has caused greater interest in the subject of antenna polarization, because of the special polarization problems inherent in space (and especially satellite) communications. I'll define linear polarization (LP) and circular polarization (CP), and discuss the applications of each.

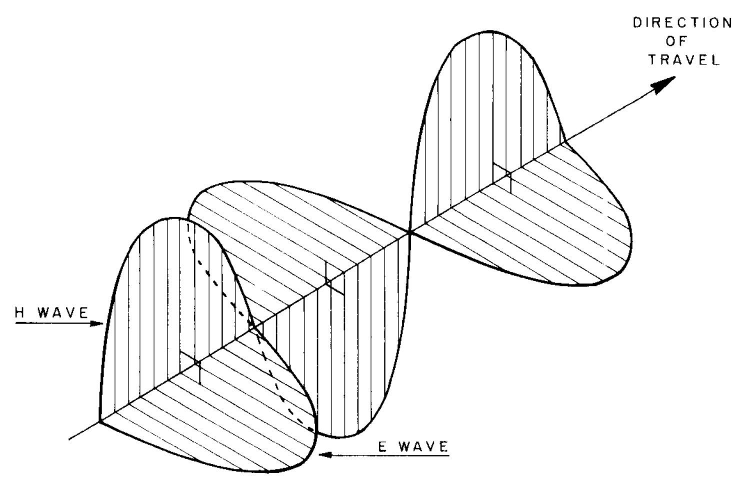

Fig 1—In a linearly polarized (LP) wave, both the E and H fields are confined to perpendicular planes, one axis of which is the direction of travel of the wave. Shown here is a horizontally polarized wave (the E field is in a plane parallel to the Earth's surface).

Electromagnetic Waves

Anyone who sat in physics class and studied electricity and magnetism, or anyone who's ever struggled with Maxwell's equations, knows that electric and magnetic fields are very closely related (that is, you can't have one without the other). We don't need to get theoretical here, just remember that a radio wave (one form of electromagnetic wave) has an electric-field component (*E wave* or *E field*), and a magnetic-field component (*H wave* or *H field*) associated with it. These two fields are always pependicular to each other in space, and travel at the speed of light away from the source (the radio antenna, in this case). See Fig 1.

Linear Polarization

In a linearly polarized wave, the E and H fields each stay in separate planes, one dimension of which is the direction of travel of the wave. A horizontally polarized wave has its E field in a plane parallel to the Earth's surface. A vertically polarized wave has its E field perpendicular to the Earth's surface. If, in the direction of maximum radiation, the E field coming from an antenna is parallel to the Earth, the antenna is said to be horizontally polarized. If the magnetic field is parallel to the Earth's surface, the antenna is vertically polarized. The planes containing the E and H fields are commonly referred to as the *E plane* and the *H plane*.

Linearly Polarized Antennas

Most hams are familiar with linearly polarized antennas. The horizontal half-wave dipole is an example of a linearly (horizontally) polarized antenna. The ¼-wave monopole or groundplane antenna is also an example of a linearly (vertically) polarized antenna. The Yagi-Uda array (the most familiar beam antenna) is a linearly polarized antenna, and can be oriented either horizontally or vertically.

For maximum response, the E plane of a linearly polarized receiving antenna must be aligned with the E field of a linearly polarized transmitted wave. In simple terms, verticals work best with other verticals, and horizontal antennas work best with other horizontal antennas.

On the HF bands, where most communication is via ionospheric propagation, antenna polarization is not as important as it is on the VHF (and higher) bands. On HF, a particular polarization is chosen more for lowering noise pickup or for a particular angle of radiation than for matching the polarization of other stations. When signals travel through the ionosphere, polarization changes in a way that is very difficult to predict. Returning ionospheric signals usually come back in all polarizations. On VHF—especially the higher bands—we are much more concerned with polarization compatibility, because most paths are non-ionospheric. We all know why the FM operator with a vertical antenna has trouble hearing stations using horizontally polarized antennas on SSB. There can be as much as 30 dB attenuation when antennas are cross polarized, although the number is usually more like 10 to 20 dB, because most antennas have some response to signals not polarized in the predominant sense of the antenna. In a cross-polarized antenna situation, received signals often peak off the *side* of the receiving antenna, where there is sometimes a better response to the received-signal polarization than there is off the front of the antenna.

Space Communications and Circular Polarization

When satellite communications came along, all of a sudden we began trying to communicate through a *moving object*. Spacecraft usually rotate and change orientation with respect to the Earth. On satellite and EME paths, signals also experience what is known as Faraday rotation, where signal polarization rotates as it goes through the ionosphere. In addition, there is the problem of polarization reversal caused by the solid geometry of the reflecting surface. Circular polarization is often used to overcome these problems. A circularly polarized antenna aligned to receive an incoming signal responds equally to all polarizations as the signal is rotated through 360°.

How does circular polarization work, and what are the trade-offs? Basically, the electric- and magnetic-field components of a circularly polarized wave rotate through 360° *in space* for each cycle of the wave. CP waves are either right-hand circularly polarized (RHCP) or left-hand circularly polarized (LHCP) depending upon which way (clockwise or counterclockwise) the fields rotate in space. The E and H fields are still perpendicular to each other at any point in time. The easiest way to visualize the generation of a CP wave is to look at two separate linearly polarized waves on the same axis and 90° apart, (one vertical and one horizontal axis) and 90° out of phase. See Fig 2. The resultant electric and magnetic fields are the vector sums of the two waves, which rotate in space once each cycle.

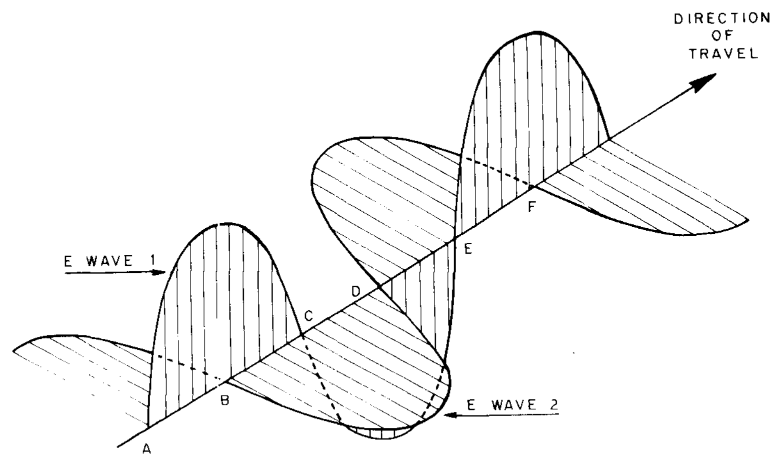

Fig 2—Two E-field waves, 90° apart in space and 90° apart in time. At point B, wave one is positive and wave two is zero magnitude. At point C, wave two is positive and wave one is zero magnitude. At D, wave one is negative, and wave two is zero magnitude. The vector sum of waves one and two (let's say they're both E-field waves) is continually rotating around the axis. (The H-field waves are there too, 90° apart from their respective E-field waves at any given time, but are omitted here for clarity).

Circularly Polarized Antennas

The most commonly used CP antenna in amateur communications is the crossed-element Yagi array. The operation of such antennas is fairly easy to visualize. Here, two linearly polarized antennas are orthogonally polarized on a common axis (the boom), and are either electrically fed 90° out of phase, or offset 90° axially (along the boom) in space. The resultant CP wave is either RHCP or LHCP, depending upon phasing. Polarization sense can be changed easily by adding or subtracting 180° from the phase of the RF fed to either set of elements.

Another circularly polarized antenna in fairly common use is the axial-mode helix. This antenna—a cylindrical, multi-turn helix approximately one wavelength in diameter—is fed over a ground plane. When the radiation from the helix wire is added vectorially, a CP wave with the direction of maximum radiation along the axis of the helix is formed. This antenna exhibits either RHCP or LHCP, and the sense can only be changed by winding the helix to produce the opposite sense.

Both of the above-mentioned CP antennas are only truly circularly polarized in the direction of maximum gain (the direction in which they are pointed). In other directions, they can have linear, elliptical (of which circular polarization is a special case), or even circular polarization of the opposite sense. Also, for maximum effectiveness, a CP antenna of the same sense (either RH or LHCP) must be used at each end of a communication path. Reversing the sense at one end will result in 20 to 30 dB of attenuation.

If a CP wave is reflected (off the moon or another passive reflector, for instance), the sense is reversed. This is why on 1296 and 2304 MHz, EME stations, by convention, transmit using RHCP antennas and receive using LHCP antennas.

A linearly polarized antenna receiving a circularly polarized signal is properly polarized for only half of each RF cycle. Remember that the E and H fields in a CP wave are continually rotating from vertical to horizontal to vertical, and so on. Because of this, there is a 3-dB degradation when using linear polarization on one end of a path and circular polarization on the other. As long as the wave stays truly circularly polarized, however, the axial orientation of either antenna (specifically, the polarization of the linearly polarized antenna) will have no effect on signal levels. That is, there is no difference between vertical and horizontal polarization at your end as long as you're communicating with a station using a CP antenna. Because high-gain CP antennas are relatively difficult to build in a small space, especially on 1269 MHz and above, many amateurs have chosen to use linearly polarized antennas with an extra 3 dB of gain for ground-station antennas on these bands. Results on AO-10 Mode L seem to prove the validity of this approach.

The idea here is to eliminate fading caused by crossed polarization resulting from spacecraft tumbling, Faraday rotation and/or ground-station location. The optimum antenna system would use CP antennas *on the spacecraft and on* the ground, with the ground station being able to switch polarization sense or be able to switch to linear polarization, depending on spacecraft orientation. With a CP antenna on the spacecraft, a linearly polarized antenna on the ground gives good results most of the time. (Reliable communications most of the time is pretty good, compared to what we usually have to live with on the HF and VHF bands!)

Conclusions

I hope this information clears up some misconceptions about antenna polarization. It is a very complex subject, but you don't need to know everything about the subject to put together an effective station. By the time you read this, AO-13 should be operational (knock on wood), and the "bird watchers" among us should experience more practical examples of above-mentioned polarization effects.

Oh yes, one more thing: Loop Yagis are *not* circularly polarized antennas! They're just long quads with circular elements. Of course, two loop Yagis can be phased to provide circular polarization, just like any other linearly polarized antennas.

See you all next month, and thanks for all the nice comments. Please send me your ideas for upcoming columns!

Components

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MORE ON TOROIDS

After the June column appeared, I received a letter from Jack Althouse, K6NY, of Palomar Engineers. Palomar offers a wide selection of Micrometals iron-powder toroidal cores. They also have a selection of ferrite cores from Fair-Rite Company.

The Fair-Rite products are of special interest if you are experiencing any problems with EMI/RFI. For fighting EMI/RFI, Palomar stocks ferrite beads, split beads (for round and ribbon cables), and ferrite rods.

According to Jack, Palomar has no minimum order. For a copy of Palomar Engineers' catalog, drop them a line at PO Box 455, Escondido, CA 92025. You may also want to request their *RFI Tip Sheet.* I'm sure that they would appreciate an SASE along with your request.

TELEDYNE CMOS SWITCH-MODE POWER-SUPPLY CONTROLLER

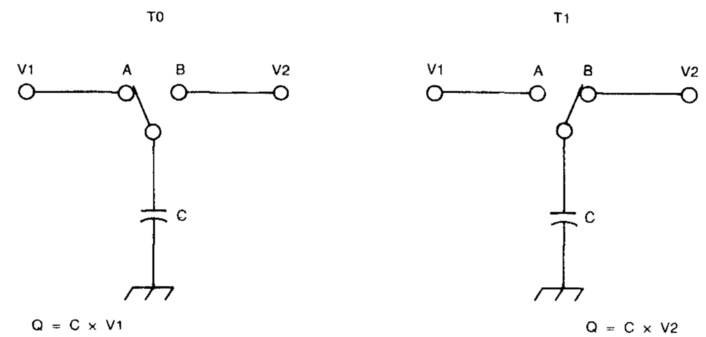
Teledyne Semiconductor has introduced what they claim to be the world's first CMOS current-mode controller for switching power supplies, the TSC170. The CMOS architecture requires only 3.8 mA maximum supply current. By contrast, bipolar circuits require *five times* that amount of current! In addition to being better-suited to low-power applications, the CMOS circuitry produces less heat.

The circuit features pulse-by-pulse current limiting, inherent feed forward, simple loop compensation, and an output stage that's optimized to drive power MOSFETs. The TSC170 uses a fixed-frequency, peak-current-terminating architecture. The clock starts the cycle, and when the peak current reaches the value set by the reference voltage, the cycle is terminated until the next clock pulse. Thus, the system appears as a constant-current source.

For information on the TSC170 CMOS SMPS Current Mode Controller, write to Teledyne Semiconductor, PO Box 7267, Mountain View, CA 94039-7267.

SWITCHED-CAPACITOR FILTERS

Switched-capacitor filters (SCFs) have dramatically changed the implementation, quality, and versatility of high-pass, low-pass, bandpass, and notch filters.[1]



$$I = \frac{Q(T0) - Q(T1)}{T1} = C(V1 - V2)F$$

$$R = \frac{V}{I} = \frac{(V1 - V2)}{I}$$

SUBSTITUTE FOR I:

$$R = \frac{(V1 - V2)}{C(V1 - V2)F} = \frac{1}{C \times F}$$

NOTE THAT R IS INDEPENDENT OF THE VALUES OF V1 AND V2

Fig 1—Illustration of the switched-capacitor effect. In the equations, I represents the net current flow from point A to point B, Q is the charge on the capacitor, F is the switching frequency and R is the resultant resistance. (See text and notes 1 and 2.)

Still, not a great deal has appeared in the press—probably because analog circuits aren't as "sexy" as digital circuits. Quietly, though, SCFs have made possible single-chip DTMF decoders and multiple-pole filters at astonishingly low prices.

Switched-Capacitor Theory

Integrating an RC network onto an IC chip is not an easy task. Capacitors become unwieldily large at about 100 pF, and resistors consume relatively large areas for even low-resistance values. Because small capacitors are easily integrated, it would be handy to use only capacitors to achieve the same result as an RC network. This is exactly how switched-capacitor circuits work.

Fig 1 illustrates the switched-capacitor effect. At T0, the voltage source, V1, connects to the capacitor and begins to charge it. The full charge on the capacitor is Q = C × V1. At T1, the voltage source supplying the capacitor is switched to V2, and the capacitor charges toward the value Q = C × V2. By using the equations given in Fig 1, the mathematics show that the resistance value (R) may be expressed as 1 / C × F, where F is the switching frequency. *[For more information on SCF theory and application, see the March and July 1984 issues of* QST[1,2]*—Ed.]*

So, by using a clock and a capacitor, a large resistance value can be electrically simulated. The value of the resistor is inversely proportional to the value of the capacitor, and more importantly, it's inversely proportional to the clock frequency. The switched-capacitor effect is the underlying principle used in the ICs to be described, and in such circuits as the SSI DTMF decoder chip, which was featured in the April "Components" column.

MAXIM's MAX260 UNIVERSAL FILTERS

In addition to having an interesting name for their company, Maxim offers a family of universal switched-capacitor filters. The family comprises three ICs, the MAX260, MAX261, and MAX262.

Each member of the family has two 2-pole filters. The filter's center frequency,

[1]Notes appear on page 14.

its Q, and its operating configuration are programmable. The MAX260 has a center-frequency range of 7.5 kHz. The MAX261 and MAX262 are high-Q filters; the MAX261's center-frequency range is 0.01 Hz to 30 kHz, and the MAX262 can be programmed with center frequencies from 0.01 to 75 kHz. Using a Q of less than two for the MAX262, allows the center frequency to be as high as 100 kHz.

The IC pricing ranges from $6 to $15 in 100-unit quantities. For more information on the MAX260 product line, contact Maxim Integrated Products, 510 N Pastoria Ave, Sunnyvale, CA 94086.

MICRO LINEAR ML2111 UNIVERSAL FILTER

The ML2111 is similar in function to the Maxim products. The center frequency range for this filter is 150 kHz, and its accuracy is ±0.4% per filter (two filters per package). For high Qs (above 20), the center-frequency range is reduced to 100 kHz. A lower-frequency version is also available as the ML2110. This part is specified to 30 kHz.

The ML2111 and ML2110 are pin-for-pin replacements for one of the first switched-capacitor filters, the National Semiconductor MF10. The ML2111 is $6.95 in 100-unit quantities, and the ML2110 is $3.95 in quantities of 100. Good news for experimenters, though: Al Tremain at Micro Linear told me he will sell small quantities directly, and the price will be in the $20 range. Contact Al at 2092 Concourse Drive, San Jose, CA 95131, or phone 408-433-5200.

I have built circuits with both the ML2111 and the MAX260. Both parts are easy to use and perform well. They work extremely well at audio frequencies, and can be cascaded to form n-pole filters. If you haven't experimented with switched-capacitor filters, these are ideal parts to start with.

Notes

[1]R. Shellenbach and F. Noble, "Switched-Capacitor Filters—An Emerging Technology for Amateur Radio Use," *QST*, Mar 1984, pp 19-25.

[2]R. Shellenbach and F. Noble, "Digital Switched-Capacitor Filters—A Practical Construction Project," *QST*, Jul 1984,pp 11-15.

Automatic Antenna Controllers

Continued from page 8.

of controlling Kenpro KR-5400A/5600A rotators with a C64. Run by software available from AMSAT, this combination provides very precise antenna positioning. The KR-001 is a dumb controller, and requires constant computer control. (One scenario suggested by Encomm is to purchase a C64 and dedicate it to antenna control. This way, you can capitalize on the memory available in the C64 by storing lots of antenna-positioning data, and free your second computer for other applications.) The KR-001 can be installed very quickly. All the necessary cables are provided, and the software is very user friendly. Price: $149.95. Encomm, Inc, 1506 Capital Ave, Plano, TX 75074, tel 214-423-0024.

AUTOTRAK

Neil Hill, K7NH, has developed a rotator-controller board that can be used with C64 or Timex 1000 computers. AUTOTRAK is available in three forms: PC board only, a complete kit of parts including the PC board, or a fully assembled and tested version. AUTOTRAK works with any of the Kenpro rotators, takes advantage of the computer port on the KR-5400A/5600A series, and works with several other rotators, including the Alliance HD73. AUTOTRAK's aiming precision is good, and is limited only by the software and rotator you use. Software and delivery charges are included in the prices. Software updates are available through the manufacturer. Price class $20 to $129, depending on version. For more information, contact Neil Hill, K7NH, 22104 66th Ave W, Mountlake Terrace, WA 98043.[2]

Summary

Applications of these controllers range from simple satellite tracking to sophisticated low-band contest antenna-system operation. Once you own an automatic rotator controller, you will wonder how you lived without it!

Notes

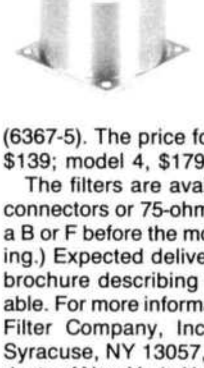
[1]J. Bloom, "The ARRL Microcontroller," *QST*, Jul 1986, pp 14- 19, and J. Bloom, "An Automatic Antenna Controller," *QST*, Sep 1986, pp 40-46.

[2]N. Hill, "A simple rotor interface board for the C-64 and the VIC-20," ham radio, Dec 1987, pp 10-27).

Bits

Tunable Notch Filters for 30 to 900 MHz

The Microwave Filter Company, Inc has introduced the 6367 series of tunable notch filters. These filters cover an approximate 2:1 frequency range with an adjustable 3-dB bandwidth. The adjustable bandwidth of these filters allows suppression of an unwanted carrier with minimum impact on adjacent frequencies. Minimum notch depth is 13 to 30 dB, depending on the model.



Tuning ranges for the various models are: 30-50 MHz (6367-1), 50-112 MHz (6367-2), 88-216 MHz (6367-3) 216-450 MHz (6367-4) and 450-900 MHz (6367-5). The price for models 1 through 3 is $139; model 4, $179; model 5, $169.

The filters are available with 50-ohm BNC connectors or 75-ohm F connectors. (Indicate a B or F before the model number when ordering.) Expected delivery time is two weeks. A brochure describing the filter series is available. For more information, contact Microwave Filter Company, Inc, 6743 Kinne St, East Syracuse, NY 13057, tel 800-448-1666. Residents of New York, Hawaii, Alaska and Canada can call 315-437-3953 collect. —*Paul K. Pagel, N1FB*