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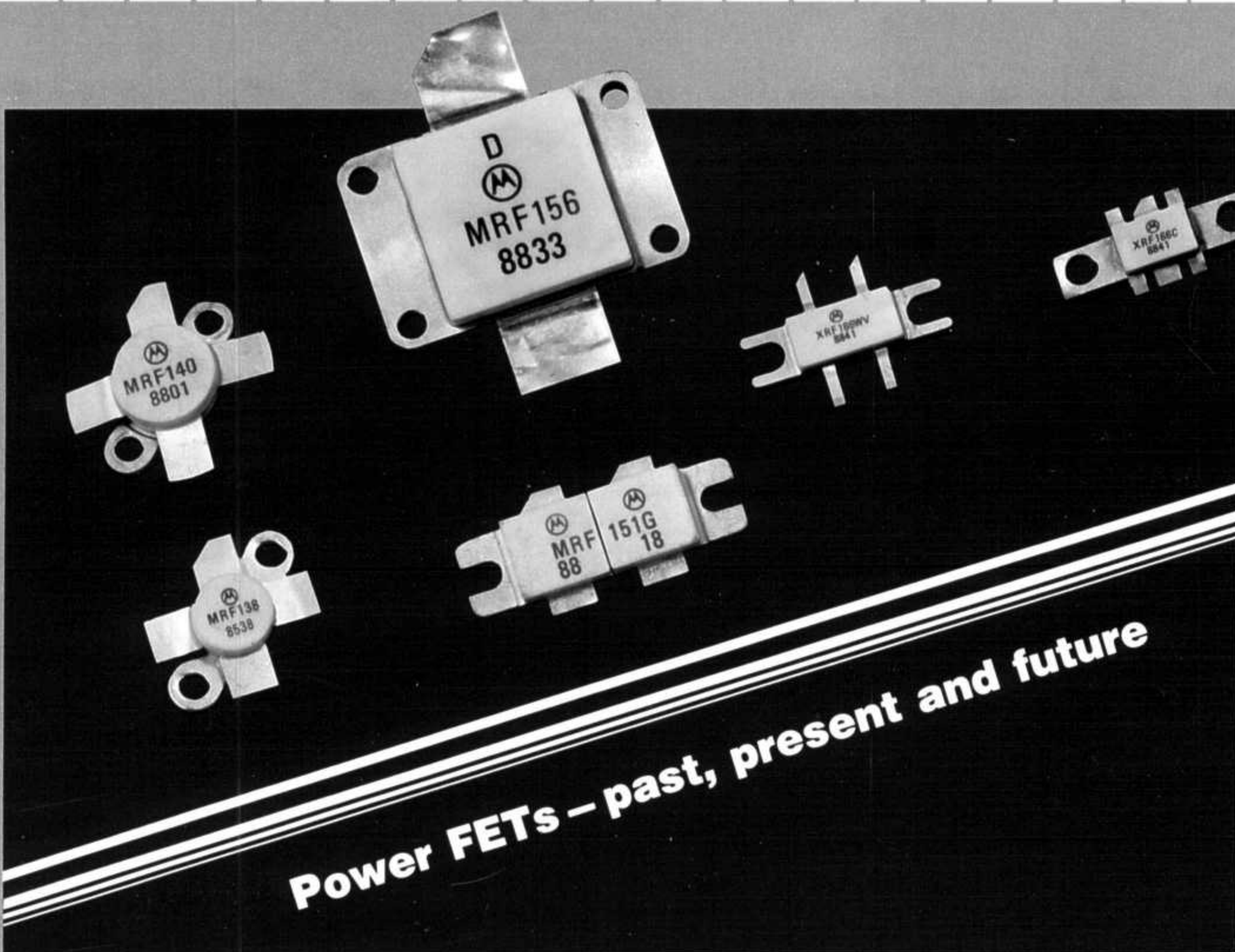

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COLUMNS

ABOUT THE COVER

A present-day power FET family portrait. These Motorola™ power FETs have power-handling capabilities ranging from a few watts to more than 100 W.

Purposes of *QEX*:

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

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

Both theoretical and practical technical articles are welcomed. Manuscripts should be typed and 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*.

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

Empirically Speaking. . .

A Digital Weekend

The ARRL Committee on Amateur Radio Digital Communication met on December 10-11 at ARRL HQ to discuss a number of topics related mostly to packet radio. Vice Director Thomas Comstock, N5TC, resigned as Board Liaison, as he will no longer be attending Board Meetings. Tom gained the respect of both the packet and satellite troops during his time on the committee. As his replacement, President Price appointed Director Frank Butler, W4RH, an EE with extensive research, development and testing experience during his career as a civilian employee of the US Air Force. So, the present committee lineup is:

Paul L. Rinaldo, W4RI	Chairman
Frank Butler, W4RH	Board Liaison
Terry L. Fox, WB4JFI	Member
Lyle V. Johnson, WA7GXD	Member
Philip R. Karn, Jr, KA9Q	Member
Doug Lockhart, VE7APU	Member
Eric L. Scace, K3NA	Member

The committee discussed comments received from several people concerning changes to AX.25 version 2.0 proposed at the 7th ARRL Amateur Radio Computer Networking Conference in early October. This review was facilitated by Eric Scace's SDL (System Description Language) diagrams, which are much more explicit than the narrative descriptions and state tables originally published for version 2.0. Also, the committee had a good idea of what works and doesn't work in the proposed version 2.1, as Phil Karn had coded the revised protocol. A few juicy bugs were crawling around, but keen eyes and unforgiving computers have them cornered. Implementers are invited to review the proposed changes and send comments to Chairman, ARRL Digital Committee c/o ARRL HQ.

The committee discussed planning for a new project to stimulate research and development of new modems and protocols for HF packet radio. While the average amateur terminal node controller (TNC) works on HF, transmission impairments (selective fading, interference and noise) cause more retries than acceptable. This problem has plagued amateur packet radio and is also keenly felt by the Federal Government. The Federal Emergency Management Agency (FEMA) has held two conferences to assess what's available and how to improve the reliability of HF packet radio. Because hams are on the spot wherever emergencies happen, FEMA needs to stay compatible with amateur packet radio by means of equipment installed in their vans and fixed installations. It doesn't just stop there. It hasn't escaped FEMA that amateur packet radio equipment is an order of magnitude cheaper than what the Government has, or will, come up with. So, if hams can make HF packet radio work, the Government would like to buy a bunch of copies.

The catalyst in the HF packet project is a new ARRL Starr Technology Fund with an initial amount of $3,000 donated by the Anne and Jacob Starr Foundation. We hope to attract additional funding in the near future to raise our chances of a successful outcome. One approach discussed by the Digital Committee was a design competition with awards for the best modems and protocols. After thorough discussion, the committee agreed that *competition* should not be the main thrust of the project—*cooperation* must be. Accordingly, work will start on developing a project plan to (a) define design goals, (b) make essential information, such as bibliographies, available those interested in proposing, (c) request proposals, (d) select the most promising approaches, (e) encourage cooperation and exchange among the participants, (f) arrange for testing, and (g) judge the final results. Stay tuned for details.

The Digital Committee reviewed the lessons learned from the HF automatic operation special temporary authority (STA) that has been running the past two years. This information will be provided in a report to the FCC. In addition, the committee decided on recommended provisions to make HF automatic packet operation permanent in the Amateur Service rules. These recommendations are to be forwarded to the ARRL Board of Directors for their consideration.

Amateurs are invited to provide inputs to the Digital Committee concerning packet radio or other forms of digital communication. Address correspondence to Chairman, Committee on Amateur Radio Digital Communication, at ARRL HQ. —*W4RI*

RF Power FETs
Their Characteristics and Applications—Part 1

By H. O. Granberg, K7ES

As early as the mid '60s, low-power (5- to 10-W) RF FETs were available from one US manufacturer. These were depletion-mode JFETS with fairly coarse geometries and only a few paralleled source sites. Operation of these devices in amplifier service was only possible up to 30 to 50 MHz because of low device power gains. Because suitable RF packages were not yet developed, the FETs were housed in stud-mounted, hermetically sealed headers similar to the present-day TO-59, which probably contributed to their poor RF performance. These early power FETs were designed for use in switching and high-current chopper applications, although some RF data was given in the data sheets.

During the following decade, power FETs were not a high-priority design item for any manufacturer until the Japanese came out with VMOS technology (using a vertical-channel structure). This allowed a denser die layout, permitting a larger number of source sites (individual small FETs) to be condensed into a given die area. This was the breakthrough in the development of RF power FETs—a year or two later several manufacturers had devices on the market using V-groove technology. Although V-groove FETs had excellent operating characteristics, such as low input, output and feedback capacitance and a low saturation voltage, they were difficult to manufacture. Etching the V-groove requires a special crystal-oriented starting material and the use of hazardous chemicals. In addition, it was a major problem to obtain uniform coverage of the oxide and metal layers in the bottom of the sharp V-groove. The successful iteration solved the latter problem by only partially etching the groove, leaving a flat bottom (see Figs 1 and 2). This solved the oxide-metal coverage problem, but the process was still difficult to control. There are some RF power FETs still on the market using this process, such as the ISOFET (see Fig 3), which is actually an inverted VMOSFET (with its source in the groove). ISOFETs have excellent RF characteristics, mainly in the form of low feedback capacitance, but all the difficulties in etching the V groove, and other problems such as excessive step-metal coverage, are present. The process does not lend itself to mass production.

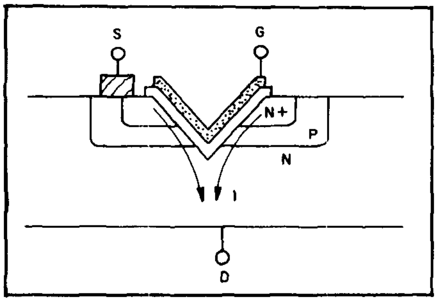

Fig 1—Early VMOSFET with sharp V-groove. This process was used from 1975 to 1976.

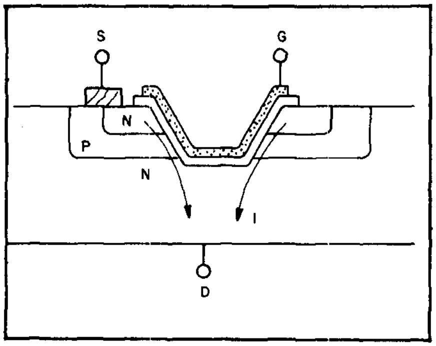

Fig 2—Modified flat-bottom VMOSFET. This process was adopted in 1979 and is still used today (mostly for limited-volume, spare-parts runs).

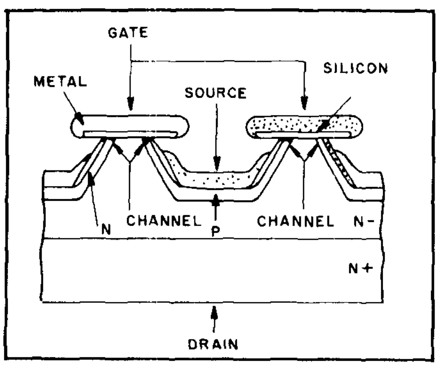

Fig 3—The ISOFET die structure, showing the source of the V-groove. Note the possibility of poor metal coverage because of shadowing effects (used 1975-77).

These problems led to the development of the DMOS process (see Fig 4), which is used by most manufacturers today. DMOS (double-diffused MOS) is a lateral process resulting in a vertical-channel structure similar to the V-groove design. Although the process is easier than VMOS, it requires more wafer processing than bipolar transistors, which along with the requirement for more die area per watt, makes the FET more costly. Vertical-channel MOSFETs, including the POLYFET of Fig 5—regardless of their trade name—are in the DMOS family, although only the one shown in Fig 4 is commonly called such.

The POLYFET, which is probably the newest variation on the DMOS structure,

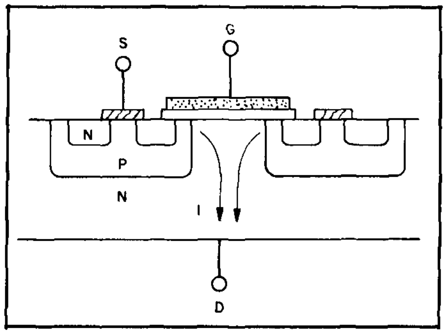

Fig 4—The basic DMOS structure. With slight modifications, this structure is used by most manufacturers for switching and RF applications.

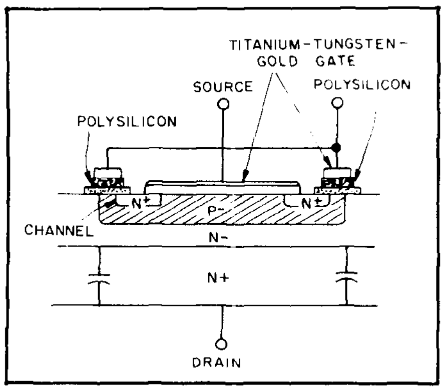

Fig 5—The POLYFET, a DMOS derivative, uses a planar process, but is difficult to implement because of its geometry and metal pattern (1985).

also uses a planar process. The name comes from the use of polysilicon gate contacts (although most other MOSFETs use them as well). The gate and source are reversed from the standard DMOS configuration, which results in a potentially shorter channel and finer geometry. The gate fingers can be made so narrow that gold metallization, with its increased deposition resolution (compared to aluminum), becomes necessary.

In addition to vertical-channel RF power MOSFETs, there are devices available from at least one Japanese manufacturer that have lateral channels (see Fig 6). Their most important characteristics are extremely low feedback capacitance and the location of the source contact in the bottom of the die.

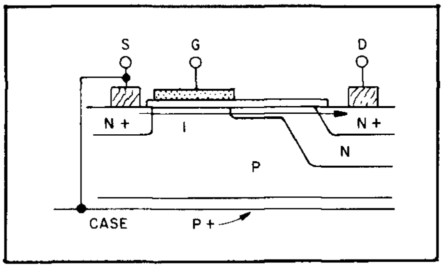

Fig 6—This lateral-channel structure uses an all-planar process and resembles multiple small-signal MOSFETs connected in parallel (1980).

The process is more costly, as about 50 percent more die area is required for a given power rating. This process can also be thought of as DMOS because double diffusion is employed. Some of the first lateral-channel devices were introduced by Hitachi in 1980.

There is another type of RF power FET available from a few manufacturers. It is called the SIT (static induction transistor), a majority-carrier device (like all FETs) that resembles a depletion-mode JFET in that it requires a negative gate potential to deplete the channel. The current flows through the bulk material and not through the surface area as in MOSFETs. This, as well as lack of a gate-oxide layer, results in higher radiation hardness than most other FETs. As shown in Fig 7, the SIT also features a vertical channel that is deeper than the MOSFET's, making the current saturation more controllable by material resistivity than the thickness of the epitaxial layer. This makes rugged, high-voltage devices easier to implement—one of the advantages of the SIT. Another advantage of SIT is the gate's insensitivity to static charges. This property comes from the SIT's thick, puncture-resistant, gate-oxide layer. Despite its advantages, the SIT has not shown high-frequency performance comparable to the MOSFET.

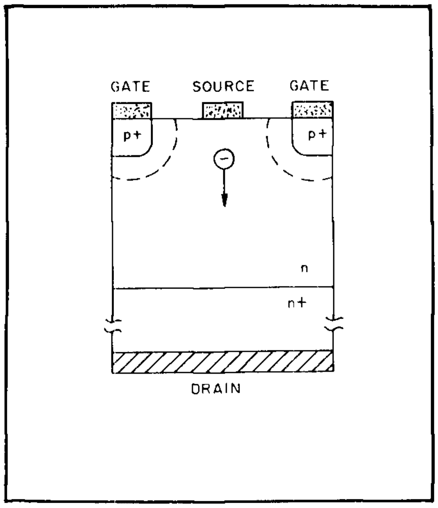

Fig 7—In the SIT, actually a short-channel JFET, current flow takes place vertically between the source and the drain. The SIT was introduced by Mitsubishi in 1976.

The DMOSFET

Since the vertical-channel DMOS is the most widely used planar process for RF and switching power FETs, the discussion and data given will focus on it. Devices with trade names such as TMOS, HEXFET and SIPMOS are known to use this process with variations in geometry, cell structure, and die layout. The DMOS structure, shown in Fig 4, was developed in 1977-79 for use in low-frequency switching FETs, and has been adopted by practically every manufacturer. Motorola (TMOS™) was a pioneer in developing the DMOS process for RF power amplifier applications. The first devices were intentionally made with coarse geometries to improve wafer yields. This resulted in higher reverse transfer or feedback capacitance (C_{RSS}) when compared to most other processes. The large-die devices were usable at VHF; the smaller ones, although power gains were adequate at UHF, were prone to instabilities.

Let's examine some of the differences between RF and power-switching FETs. In FETs designed specifically for RF, the die geometry is usually finer (larger ratio of the gate periphery to the channel area) than in switching power FETs. This reduces device capacitance. Further reduction is achieved by splitting the die into multiple cells (groups of source sites and gate fingers), where the gates and sources are usually connected in groups of two or four to the common package terminals by individual bonding wires. Some of the larger dies may have up to 36 cells, each of which consist of approximately 70 FETs, making a total of more than 2500 FETs per device.

In switching power FETs, the connections to the numerous source sites and gates are made with a metal pattern on the die surface, which allows the use of single large-diameter bond wires for the source and gate contacts. The increased metal area results in increased MOS capacitance, which is reflected in the device's input and feedback capacitance. The transconductance (g_{FS}) of a MOSFET is a measure of its electrical size. Thus, a good indication of its high-frequency performance can be obtained by comparing the capacitance values (especially C_{RSS}) of devices with similar transconductances.

Another parameter worth mentioning is gate resistance. Most modern power FETs use a gate structure of polycrystal silicon, which has a bulk resistance comparable to carbon. It is also used as a conductor between the metal pattern and each individual gate finger. In RF power FETs, each gate is fed through a separate contact having a resistance of approximately 0.1 ohm. In switching power FETs, the polycrystal silicon is applied in a sheet form in a separate layer, but the distance between the metallization and the farthest gate area still results in a gate resistance that's at least 30 to 40 times greater than a die of comparable size. In high-frequency applications, the high gate resistance permits a part of the drain-source RF voltage to be fed back to the gate (because of the C_{RSS}) at amplitudes that can rupture the gate-source oxide layer. The rupture first occurs in the far end of the die, away from the gate terminal. Since the gate resistance is integral to the FET die, external limiting or clamping circuits are of no help. The gate of a MOSFET is the most sensitive part of the device, and can be permanently damaged by static charges during handling. Although the larger (higher-power) devices because of their high gate-source capacitance are less susceptible to static electricity, proper handling precautions should be used.

RF Power MOSFET Characteristics

Even though the cost of RF power MOSFETs is higher than BJTs (because of the larger required die area and more difficult wafer processing), they are becoming increasingly popular in new designs. MOSFETs offer certain advantages, which in most cases make the design simpler and more compact. These advantages include higher input impedance (in all circuit configurations), gain control by varying the dc gate-bias voltage, and immunity to thermal runaway. They do have some disadvantages, however, such as higher saturation voltage, which makes low-voltage operation less feasible than with BJTs. Power

MOSFETs are also susceptible to static-electricity punch-through of the thin gate-oxide layer. The gate-oxide layer is also sensitive to radiation, which induces a negative shift in the gate-threshold voltage. This characteristic makes the MOSFET more vulnerable in this respect than JFETs or BJTs. Other characteristics of RF power MOSFETs and BJTs are shown in Table 1.

In N-type material, electrons are majority carriers, whose mobility is superior to that of holes in P-type material. Practically all RF power FETs are of the N-channel type for improved high-frequency performance. Most are enhancement-mode types, (the device is cut off at zero gate voltage, and the gate must be biased to a positive voltage with respect to the source for drain-source current to flow). Other RF FETs, such as SITs and GaAsFETs, are depletion-mode devices and must be turned off by a negative bias.

Table 2 explains the importance of the various MOSFET parameters and compares them to the more-familiar BJT parameters. One important parameter not listed is thermal stability. A MOSFET is almost always biased for some level of idle current, while a BJT, with its low base-emitter voltage drop, must be biased only for linear operation. The forward voltage variations in a base-emitter junction are 1 to 2 mV/°C, and always have a negative temperature coefficient. The gate threshold voltage of a MOSFET also has a negative temperature coefficient of about 1 mV/°C (see Fig 8), but only to a certain current level, whereafter the bulk resistance of the FET, which has a positive coefficient, becomes dominant. This also causes the FET's g_{FS} to decrease with increasing temperature. These two characteristics protect the FET from thermal runaway, and help prevent overdissipation under mismatched load conditions. (This is not always true with

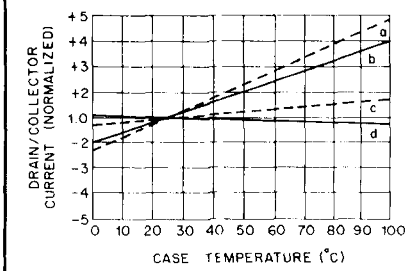

Fig 8—Drain and collector idling currents versus temperature at constant gate or base voltages. Dashed lines represent BJTs; solid lines represent power MOSFETs.

Table 1

Comparison of RF power MOSFET and bipolar transistor characteristics when used as RF amplifiers.

Characteristic	*Bipolar*	*MOSFET*
$Z_{in}R_S/X_S$ (2.0 MHz):	3.80 -*j* 2.0 ohms	19.0 -*j* 3.0 ohms
$Z_{in}R_S/X_S$ (150 MHz):	0.40 + *j* 1.50 ohms	0.60 -*j* 0.65 ohms
Z_{ol}(Load Impedance):	Nearly equal for each transistor, depending upon supply voltage and power output.	
Biasing:	Not required, except for linear operation. High-current (I_Ch_{FE}) voltage source necessary.	Some gate bias always required. Current source, such as resistor divider is sufficient.
Linearity:	Low-order distortion depends on die size, geometry and h_{FE}. High-order IMD is a function of type and value of emitter ballast resistors.	Low-order distortion worse than bipolars for given die size and geometry. High-order IMD better due to lack of ballast resistors and associated nonlinear feedback.
Ruggedness:	Usually fails under high-current conditions (overdissipation). Thermal runaway and secondary breakdown possible.	Overdissipation failure less likely, except under high-voltage conditions. Other failure modes: Gate punch-through and exceeding BV_{DSS}.
Stability:	Instability mode known as half f_o troublesome because of varactor action of base-emitter junction diode. Lower ratio of feedback capacitance versus input impedance.	Superior stability because of lack of diode junctions and higher ratio of feedback capacitance versus input impedance.
Advantages:	Wafer processing simpler. Low collector-emitter saturation voltage makes low-voltage operation feasible.	Input impedance more constant under varying drive levels. Better stability, reduced high-order IMD, easier to broadband. Devices and die can be paralleled. High-voltage devices easy to implement.
Disadvantages:	Low input impedance with great reactive component. Internal matching required to reduce Q. Input impedance varies with drive level. Devices or die cannot be easily paralleled.	Larger die required for comparable power level. Nonrecoverable gate breakdown. High drain-source saturation, which makes low-voltage, high-power devices less practical.

Table 2
Comparison of bipolar and power MOSFET dc parameters.

"Equivalent" parameters Bipolar	MOSFET	Description
BV_{CEO}	BV_{DSO}	Not specified or measurable with MOSFETs. In case of low gate-source leakage, the gate can charge to voltages exceeding the punch-through voltage rating.
BV_{CES}	BV_{DSS}	Normal method of measuring MOSFET breakdown voltage. It refers to the maximum drain-to-source voltage the FET can withstand with the gate dc biased or at the same potential as the source.
BV_{CBO}	BV_{DGO}	Not specified or measurable with MOSFETs. Gate-source rupture voltage would be exceeded.
BV_{EBO}	V_g	Not specified or measurable with MOSFETs unless done carefully at low current levels. Gate rupture can be compared to exceeding a capacitor's maximum voltage rating.
V_b (forward)	V_g(th)	Not specified or necessary in most cases for BPTs. For a MOSFET, this parameter determines the turn-on gate voltage, and must be known for biasing the device.
I_{CES}	I_{DSS}	Drain-source leakage current with gate shorted to source. BPT and FET parameters are equivalent, and normally only refer to wasted dc power reliability.
I_{EBO}	I_{GS}	Not normally given in BPT data sheets, but important for MOSFET biasing. Both affect their associated device's long-term reliability.
$V_{CE(SAT)}$	$V_{DS(SAT)}$	Not usually given in BPT data sheets, but important in certain applications. With power MOSFETs this parameter is of great importance. The MOSFET numbers are larger than those for BPTs and are material and die-geometry dependent.
h_{FE}	g_{FS}	These are parameters for low-frequency current and voltage gain, respectively. In a MOSFET, the g_{FS} is more an indication of the device's electrical size. To a certain extent, it depends on processing.
f_T	(f_T)	Unity current or voltage gain frequency. Not given in many MOSFET or BPT data sheets. The value can be two to five times greater for the MOSFET, for equivalent geometry and electrical size (see Fig 1). The figure of merit for a MOSFET is usually considered to be the ratio of the gate-source capacitance versus the g_{FS}, but other parameters such as the R_{DS} (on) have some effect on the figure of merit.
G_{PE}	G_{PS}	Power gain in common-emitter or common-source configuration. This figure is roughly the same for both types of devices; it's normally regarded as current gain for the BPT, and voltage gain for the MOSFET. At lower frequencies, where FET gain is extremely high, the number may merely be an indication of how much stable and usable gain is available.
C_{IB}	C_{ISS}	Base-to-emitter or gate-to-source capacitance. Rarely given for BPTs. In RF power FETs, the C_{ISS} has a greater effect on the gate-source impedance. In fact, if stray inductances from the metal die patterns wire bonds and the transistor package were absent, the gate impedance would be be a pure capacitive reactance. The C_{ISS} consists mostly of die MOS capacitance, whereas the C_{IB} of the BPT is a combination of MOS and diode-junction capacitance. Since the diodes are forward biased during one half of the cycle, and reverse biased during the other half, the base impedance is largely drive-level dependent.
C_{OB}	C_{OSS}	Collector-to-emitter or drain-to-source capacitance. Both are usually specified, and are approximately equal in value for a given device rating and voltage. Both are combinations of MOS and diode capacitance. Each affects efficiency, since this capacitance must be charged and discharged at the rate of the operating frequency.
C_{RB}	C_{RSS}	Collector-to-base or drain-to-gate capacitance. Rarely specified for BPTs. Normally referred to as the feedback capacitance for MOSFETs, considering the reduced gate-source capacitance and superior high-frequency performance. At low frequencies, the C_{RSS} provides a 180°-out-of-phase feedback to the gate, which can turn into positive feedback at high frequencies, depending upon stray inductances and the C_{ISS}. The results will be noticed as parasitic oscillations unless the C_{RSS} is low or the resonance fall outside the device's frequency capabilities.

devices operated at high supply voltages.)

Despite the RF power MOSFET's cost-related drawbacks, its advantages still make it the choice over BJTs in many applications. At VHF and UHF, the high gate-input impedance and the high power gain of the MOSFET make possible the design of broadband amplifiers with simpler input-matching networks. Since the gate-source impedance remains capacitive to much higher frequencies, it makes internal matching networks unnecessary, at least up to VHF, for devices up to the 100- to 150-W power class. On the other hand, VHF bipolar transistors with power ratings of 50 W and higher commonly employ internal matching networks that transform the die impedance to a practical level.

In general, at frequencies below VHF, the input and output impedance-matching techniques for FETs and BJTs are similar. Only the network values differ in most cases. At VHF and above, the differences become more noticeable, and multi-octave amplifier designs with power outputs in the 200- to 300-W range are not uncommon. Another difference is the FET's higher ratio of feedback capacitance (drain-to-gate v collector-to-emitter) to input impedance/capacitance. For this reason, FETs exhibit considerably higher unity gain frequencies with comparable geometries (see Fig 9). For example, a 150-W FET may have a power gain of 10 to 12 dB at 150 MHz, while its BJT counterpart is usable only to 30 to 50 MHz. Lower feedback capacitance (C_{RSS}), reduces the effects of the drain load and power output variations on the gate, which appear as changes in the reflected input power. The gate-source capacitance (C_{ISS}) is multiplied by C_{RSS} because of the Miller effect. C_{ISS} goes through some rather complex excursions during the on-off cycle, but in high-frequency operation the effective value is an average, usually higher than the number given in a data sheet.

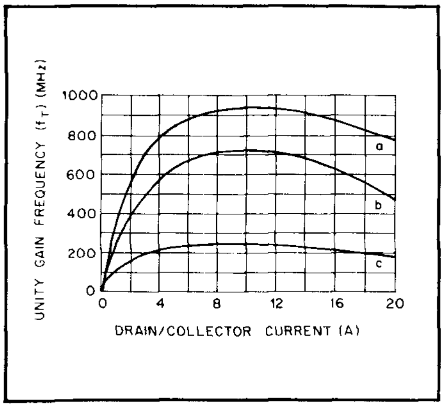

Fig 9—Unity gain frequency versus I_D or I_C. Curve A represents a 150-W RF power FET. Curve C is a BJT with the same basic geometry. Curve B is a standard switching power MOSFET with roughly equivalent gate periphery, for comparison.

There are two major parameters that can make FETs less efficient than BJTs. One parameter is the high drain-source saturation voltage mentioned earlier, and the other is the greater output capacitance (C_{OSS}/C_{OB}) of FETs. FETs generally require 30 to 50 percent more die area for a comparable power rating, depending on die geometry and other factors.

One of the claimed advantages of power FETs is that they can be connected in parallel and maintain equal current sharing as long as the gate threshold voltages are matched. In many cases, the literature does not provide sufficient details, forcing many experimenters to learn the hard way. Because of the FET's high power gains and unity gain frequencies, directly paralleled devices will form a multivibrator-type oscillator, similar to the bipolar emitter-coupled version described in many textbooks. Inductance and capacitance determine the oscillator frequency; FETs usually have enough internal capacitance and inductance to oscillate without the added elements. Depending on the type, electrical size, package style and external connections, the oscillations may occur at frequencies of a few hundred to more than a thousand megahertz. Depending on the conditions mentioned earlier, the oscillations may be weak (or unnoticeable), or strong enough to destroy the FET, usually by inducing voltage swings that exceed the rupture voltage of the gate-oxide layer.

There are several ways to cure to the problems mentioned above. The most common and practical solution is to lower the Q of the resonant circuit by introducing series gate resistance or lossy inductance (ferrite beads, etc). Because this forms RC/LC filter networks effectively in series with the input, loss of high-frequency gain, and reduced system bandwidth results. The only way to parallel power MOSFETs while maintaining stability and without affecting their high-frequency characteristics is to reduce the source-to-source inductance to an absolute minimum. This is difficult to achieve in practice. The latter method becomes important when individual dies are paralleled within one package. In this case, the source bonding wires can be made extremely short, and multiple wires can be used to further lower the source inductance if necessary. The additional gate isolation is usually not necessary, but increasing the length of the gate wire bonds has a beneficial effect. As a rule, directly paralleled FETs require a minimum source inductance, while the gate-to-gate inductance should be as low as can be tolerated to maintain stability.

In push-pull circuits, a similar situation may be present depending on the configuration of the input-matching network. Whether the network consists of lumped elements, striplines, wideband transformers or a combination of these, the gates may see it as a parallel connection at a frequency at which element reactances are low and the Q is sufficiently high. For example, a 100- to 175-MHz amplifier can be designed with an input transformer having only one turn in the secondary. This one turn may present a low enough reactance between the gates to result in oscillations at the resonant frequency of the elements. Thus, the secondary winding, even if lower-frequency response is not required, should always be "de-Qd" by slipping a bead of suitable magnetic material over the winding. Different approaches may be necessary when using lumped-element and stripline designs.

Next Month

Our discussion of RF power FETs will continue in February *QEX*. Topics include circuit configurations, linearity, noise performance, switching-mode applications and a look into the future of RF power MOSFETs.

Bits

Pomona Electronics Catalog

The 134-page, 1989 edition of the Pomona Electronics catalog is now available. This catalog describes and illustrates 900 test products, 90 of which are new for 1989.

Pomona offers a wide selection of banana plugs/jacks/adapters, coaxial and triaxial audio connectors, test clips, probes, patch cords, cable assemblies and SMD test products. Newly added to the catalog are static-control products, oscilloscope probe kits, multimeter test-lead-adapter kits.

Copies of the catalog are available free on request. Contact ITT Pomona Electronics, 1500 East Ninth St, Pomona, CA 91766, tel 714-623-3463. —*Paul K. Pagel, N1FB*

Path Selection—Part 2

By Dennis L. Haarsager, N7DH
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Last month, N7DH covered basics of earth curvature, Fresnel zones and radio-wave refraction in the atmosphere. In this second of two parts, actual path planning is covered.

Topographic Maps

Path terrain may be analyzed by using topographic maps published by the US government and available at a nominal cost. In the US, topographic maps are generally available in 7.5-minute-per-side, 15-minute-per-side, and 1° × 2° sizes. The scales on these maps are 1:24,000 (1 inch ≈ 2000 feet), 1:62,500 (1 inch ≈ 1 mile), and 1:250,000 (1 inch ≈ 4 miles), respectively. The first two types are good for terrain analysis and contain elevation contours every 10, 20, 40, or 80 feet, depending upon local topography. The 1:250,000 scale maps are fine for broad planning work, but because their elevation contours are usually drawn every 200 feet, they are not recommended for specific path studies. Maps with metric contours are now being introduced in the US, and are generally available in other parts of the world. Map indexes can be obtained from the following:

Regions east of the Mississippi:
Branch of Distribution
US Geological Survey
1200 South Eads St
Arlington, VA 22202

Regions west of the Mississippi:
Branch of Distribution
US Geological Survey
Box 25286, Federal Center
Denver, CO 80225

The map indexes tell you where you can obtain the maps in your area. Some map dealers not listed in the indexes; these can be found in the classified section of your telephone book under Maps. Outdoor outfitters, surveying suppliers, and libraries are good sources of topographic maps. Libraries in the US, often located at universities designated as Federal Depository Libraries, usually have complete collections of topographic maps. You may even be able to check library maps out for short periods of time. A booklet entitled *Topographic Maps*, outlining map symbols and other hints on map reading, is available from:

National Cartographic Information Center
US Geological Survey
507 National Center
Reston, VA 22092

Fig 3 shows a small section of a 7.5-minute-series topographic map. Notice that every fifth contour is heavier than the other contours. When the normal contour is 20 feet, the heavier contours are every 100 feet; when the normal contour is 80 feet, the heavier contours are every 400 feet, and so on. The heavier contours are labeled with the elevation, and thereby provide a reference for reading the map.

Other terrain features to look for are blue lines representing rivers, and progressively smaller concentric contours that usually indicate hilltops. The presence of rivers or streams, of course, tells you that the terrain is sloping downward to the water. The progressively smaller contours usually tell you the terrain is sloping upward toward the smallest one. The only exception to this—and it is a relatively rare occurrence—is that sometimes progressively smaller contours indicate a depres-

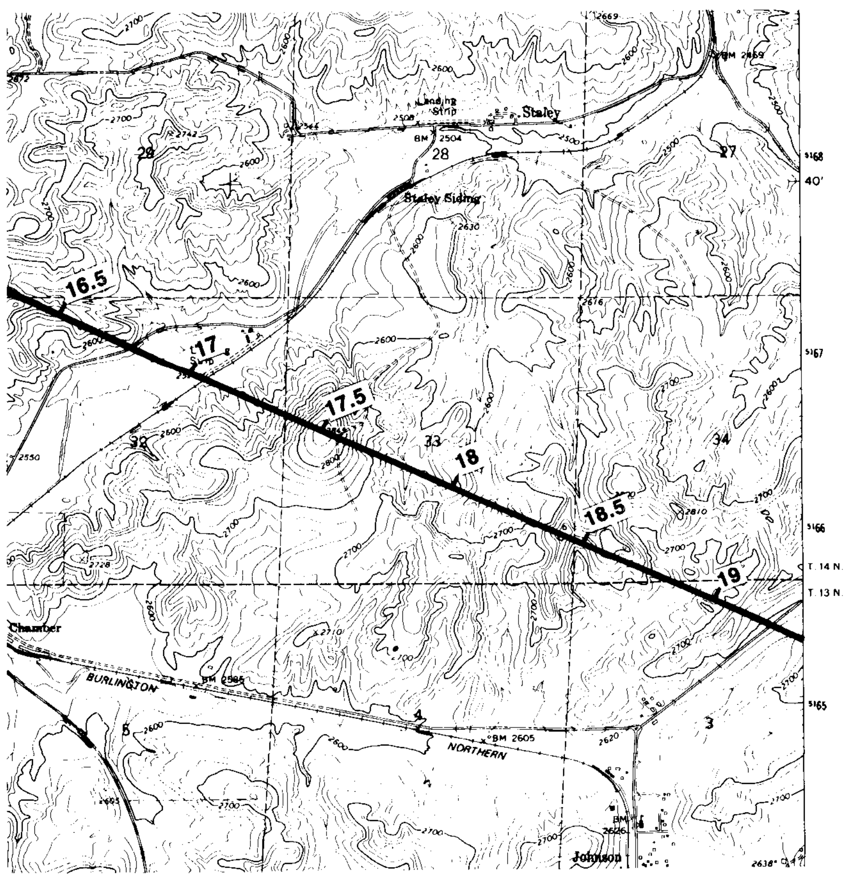

Fig 3—A section of a topographical map showing part of a radio path. The elevations of each high point along the path must be considered when evaluating a path. See text and Table 2.

sion instead of a hilltop. These are easy to spot, though, because the inner portion of the depression is marked with small ticks perpendicular to the contour lines and in the direction of the depression.

Not infrequently, communications paths go over the tops of hills, and instead of being able to interpolate the elevation between contours, you are left without a higher contour for a reference. The rule of thumb in these cases is to assume that the hilltop is half the smallest normal map contour above the highest contour shown.

Trees and buildings should be treated as if they are solid granite. Heavily forested areas are usually shaded green on topographic maps, but tree heights are not included. You will have to determine tree heights independently, along with the height of any other obstructions, such as buildings, along the path.

Table 2 shows the distances and elevations from a section of a path which coincides with the terrain contours along the line drawn on the map in Fig 3. Close examination of Fig 3 will show how the data were obtained. Drawing the profile on rectilinear graph paper is a simple matter of choosing a scale, transferring the data to the graph, and drawing in the lines in dot-to-dot fashion.

Table 2

Elevations of Key Points Along the Path Shown in Fig 3

Distance	*Elevation (feet above sea level)*
16.50	2685
16.87	2525
17.50	2845
17.82	2630
18.02	2700
18.10	2635
18.42	2800
18.50	2740
18.59	2800
18.93	2660
19.00	2680

Distances and Azimuth Headings

Distances and azimuth directions between sites are relatively easy to calculate from geographical coordinates. The accuracy of these calculations depends upon the accuracy of the coordinates from which they are calculated. In calculating site coordinates, it is important to use the best maps available, as mentioned previously.

The easiest way to determine geographical coordinates is a simple method using proportions. Mark the desired site on your map and, with a straightedge, draw a box around the site from the nearest reference ticks provided on the map. These are at 2-minute, 30-second intervals on 7.5-minute maps, and at 5-minute intervals on 15-minute maps.

With a metric scale (or an English scale calibrated in decimal inches), carefully measure and note the distance from the site to the eastern and southern lines you have drawn, and also the distances between the horizontal and vertical reference lines. A point to remember: For the distance between the vertical reference lines, be sure to place your scale at the site location, because on maps for the northern hemisphere, the distances are larger south of the site and smaller north of the site. Inaccurate reading can result from placement of the scale at any point other than the site of interest.

Use the following equation to determine the latitude or longitude:

$$L_s = L_r + W(S/T) \qquad \text{(Eq 11)}$$

where

L_s = site latitude or longitude in decimal degrees
L_r = eastern reference longitude or southern reference latitude in decimal degrees
W = width of the reference interval in decimal degrees
S = distance of the site from the reference line
T = width of the reference interval (in the same units as S)

To convert degrees, minutes, and seconds to decimal degrees, use the function on your calculator designated for that purpose, or the following algorithm:

$$D_{D.d} = D + M/60 + S/3600 \qquad \text{(Eq 12)}$$

where

$D_{D.d}$ = decimal degrees
D = degrees portion of coordinates
M = minutes portion of coordinates
S = seconds portion of coordinates

To convert decimal degrees back to degrees, minutes and seconds, use the following algorithm:

$$D = \text{int}(D_{D.d}) \qquad \text{(Eq 13A)}$$

$$M = \text{int}(60\ \text{frc}[D_{D.d}]) \qquad \text{(Eq 13B)}$$

$$S = 60\ \text{frc}(60\ \text{frc}[D_{D.d}]) \qquad \text{(Eq 13C)}$$

where

$D_{D.d}$ D, M, and S are as defined in Eq 12
int(x) = integer part of x
frc(x) = fractional part of x

Use Eqs 12 and 13 to demonstrate that 45.1234 degrees = 45° 7′ 24.24″.

There are a number of methods for determining distance and azimuth from geographical coordinates. The method given in *The ARRL Antenna Book*[3] uses standard spherical trigonometry based on the assumption that the earth is a perfect sphere. In fact, the earth is oblate—that is, slightly flattened at the poles and bulging at the equator. A number of geodetic algorithms have been developed to compensate for the earth's shape, but they are too complicated to detail here. The most accurate algorithms are Cunningham's azimuth formula and Rudoe's formula for distance. For the short distances involved in typical UHF and microwave paths, the following distance algorithm agrees closely enough with Rudoe's formula for amateur *and* professional work.

$$H = (B_1 - B_2) \times (68.962 + 0.04525A_m - 0.01274\,A_m^2 + [4.117 \times 10^{-5}]A_m^3) \qquad \text{(Eq 14)}$$

$$V = (A_1 - A_2) \times (68.712 - [1.184 \times 10^{-3}]\,A_m + [2.928 \times 10^{-4}]\,A_m^2 - [2.162 \times 10^{-6}]\,A_m^3) \qquad \text{(Eq 15)}$$

$$D_{km} = 1.609 \times \sqrt{(H^2 + V^2)} \qquad \text{(Eq 16A)}$$

$$D_{mi} = \sqrt{(H^2 + V^2)} \qquad \text{(Eq 16B)}$$

where

H = horizontal distance in miles
V = vertical distance in miles
A_1, A_2 = latitudes in decimal degrees
A_m = mean latitude in decimal degrees
B_1, B_2 = longitudes in decimal degrees
D = path distance in miles or kilometers

For azimuth, the spherical trig formula from *The Antenna Book* provides quite acceptable accuracy:

$$\cos(C) = (\sin(A_2) - \sin(A_1)\cos(D_m))/[\cos(A_1)\sin(D_a)] \quad \text{(Eq 17)}$$

where

C = bearing east of North if $\sin(B_1 - B_2)$ is positive (360-C if it is negative)
A_1, B_1 = your latitude and longitude
A_2, B_2 = other latitude and longitude
D_a = angular distance between sites:

$$\cos(D_a) = \sin(A_1)\sin(A_2) + \cos(A_1)\cos(A_2)\cos(B_1 - B_2) \quad \text{(Eq 18)}$$

If you are using these algorithms with a computer, be sure to define the variables used as arguments for trigonometric functions with as large a degree of precision as is available to you. Small angles are particularly troublesome for some BASICs to handle with good accuracy.

Example

Krell Hill's coordinates are 47°34′34″ N, 117°17′58″ W. Kamiak Butte's coordinates are 47°51′43″ N, 117°10′26″ W. What is the distance and azimuth heading from Krell Hill to Kamiak Butte?

We first convert the coordinates to decimal degrees. Krell Hill's become 47.576°N, 117.299°W. Kamiak Butte's become 46.862°N, 117.173°W. The mean latitude is (47.576 + 46.862)/2 = 47.219°N.

From Eq 14, we find that the horizontal distance is 5.905 miles. From Eq 15, we find that the vertical distance is 49.336 miles. Then, from Eq 16B, we find that the path distance is 49.69 miles. This agrees within two decimal places with Rudoe's formula.

From Eq 18, we find that the angular distance, D_a, is 0.7192°. Using this in Eq 17, we find that the azimuth heading from Krell Hill to Kamiak Butte is 173.14° east of north. Cunningham's equation gives 173.12°.

A good, large protractor will allow you to use the azimuth for determining, with accuracy suitable for amateur purposes, the points at which your signal will "cross" the edges of topographic maps.

Path Budgets

Once you have determined that the path has sufficient clearance to eliminate or minimize obstruction losses, it is useful to construct a decibel budget for the communications circuit. This is simply estimating and aggregating all the gains and losses (debits and credits, if you will) involved in the circuit. The largest loss in a circuit is path loss. This can be determined from

$$A = 92.45 + 20\log(D) + 20\log(f) \quad \text{(Eq 19A)}$$

where

A = path attenuation in decibels
D = path distance in kilometers
f = frequency in GHz, and $\log(x) = \log_{10}(x)$

$$A = 96.58 + 20\log(D) + 20\log(f) \quad \text{(Eq 19B)}$$

where

D = path distance in miles

Attenuation resulting from atmospheric gases can generally be ignored at frequencies below 20 GHz, but above that frequency, these losses become significant. At 24 GHz, use 0.11 dB/km; at 48 GHz, use 0.22 dB/km, and—for those of you who can generate RF at 74 GHz—use 0.26 dB/km.

Additional losses occur in transmission lines and connectors, if you are using them. Waveguide losses at 3.4 GHz are about 1.8 dB/100 meters of waveguide; at 5.8 GHz, 4 dB/100 m; at 10.25 GHz, 11 dB/100 m, and at 24.1 GHz, 37 dB/100 m. Many amateurs operating at microwave frequencies mount transmitters right at the antenna.

The positive components of your path budget include the transmitter output power and antenna gains. Transmitter power and receiver threshold (the smallest signal that produces usable receiver output) are, at microwave frequencies, generally expressed in decibels relative to one milliwatt (dBmW).

The receiver FM improvement threshold (about 10 dB above the total noise input and generally considered the smallest usable FM signal) may be calculated from

$$T_{FM} = -103.98 + 10\log(B) + F \quad \text{(Eq 20)}$$

where

T_{FM} = FM improvement threshold in dBmW
B = receiver IF bandwidth (at the 3-dB points) in MHz
F = receiver noise figure in dB

This makes possible an example using the path in Fig 2 in part 1 of this article. Assume that the operating frequency is 10.25 GHz, the transmitter power is 100 mW, and the transmitting and receiving antennas are horns with gains of 17 dBi each. The receiver has an IF bandwidth of 200 kHz and a noise figure of 11 dB. The path budget for this circuit is as follows:

Transmitter power (10 log 100 mW)	20.0 dBmW
Path loss (Eq 19B)	−150.3 dB
Absorption loss	0.0 dB
Obstruction loss (Fig 2)	0.0 dB
Transmit-antenna gain	17.0 dBi
Receive-antenna gain	17.0 dBi
Transmission-line loss	0.0 dB
Connector loss	0.0 dB
Received signal	−96.3 dBmW
Receiver threshold (Eq 20)	−100.0 dBmW

Algebraically subtracting the threshold from the received signal, we find that we have 3.7 dB of usable signal (or fade margin) over the FM improvement threshold. That's not much, but makes it likely that skilled operators can make contact.

The useful aspect of working in decibels is that equipment changes can be easily evaluated. For example, if a receiver with a phase-locked frequency control circuit is substituted and the receiver bandwidth is reduced to 3 kHz, all other things being equal, the receiver threshold becomes −118.2 dBmW, resulting in an increase in fade margin to 21.9 dB. If, instead of substituting a different receiver, the transmitter antenna is changed to a commonly used 25-inch snow-sled dish antenna with about 28 dBi gain, the receiver threshold remains the same, but the received signal increases by 11 dB, making the fade margin 14.7 dB.

Summary

Although path analysis looks like a complicated process at first blush, it can be done in a systematic and relatively easy way. Applying these path-evaluation techniques can permit you to do intelligent planning and evaluate trade-offs in equipment performance.

Notes

[3]G. L. Hall, ed, *The ARRL Antenna Book*, 14th edition (Newington: ARRL, 1984), pp 16-4 to 16-5.

References

B. R. Bean, C. A. Cahoon, C. A. Samson and G. D. Thayer, *A World Atlas of Atmospheric Radio Refractivity* (Boulder: Environmental Science Services Administration, 1966), ESSA Monograph 1.
G. Bomford, *Geodesy* (Oxford: Clarendon Press, 1971).
I. T. Corbell, *Microwave Engineering for the Broadcaster* (Washington: NAB, 1975).
E. T. Stephansen, "Clear-Air Propagation on Line-of-Sight Radio Paths: A Review," *Radio Science*, Sep-Oct 1981, p 610.
R. F. White, *Engineering Considerations for Microwave Communications* (San Carlos, CA: GTE Lenkurt, Inc, 1970).

Correspondence

Here's another bibliography I've compiled. This one is for those interested in crystal filter design. Note: The asterisk (*) following an item indicates information reprinted in *Single Sideband for the Radio Amateur,* 5th ed, (Newington: ARRL, 1970). *—Domenic M. Mallozzi, N1DM, 26 Carey Ave, Apt 8, Watertown, MA 02172.*

Crystal Filter Design Bibliography

H.V. Bell, "An Introduction to Crystal Filters," *RSGB Bulletin* (London): Jan-Feb 1962.

C. P. Clarke, "Variable-Bandwidth Filters—Irish Style," Technical Correspondence, *QST*, Mar 1978, p 35.

D. DeMaw, "Simple Ladder IF Filter," *QST*, Aug 1978, p 18.

M. Dishal, "Modern Network Theory of Single Sideband Crystal Filters," *Proc. IEEE*, Sept 1965, pp 1205-1216.

W. Doyle, "Systematic Design of Crystal Ladder Filters," *ham radio*, Feb 1982, pp 40-50.

J. A. Hardcastle, "Some Experiments with High-Frequency Ladder Crystal Filters," *QST*, Dec 1978, pp 22-24. (This is a reprint of the December installment of the following article.)

J. A. Hardcastle, "Computer-Aided Ladder Crystal Filter Design," *Radio Communications*, May 1983, pp 414-420.

P. Hawker, "Technical Topics," *Radio Communications*, Sept 1976, pp 672-673.

W. Hayward, "Ladder Crystal Filter Design," *QST*, Nov 1980, pp 20-23.

W. Hayward, "A Unified Approach to the Design of Crystal Ladder Filters," *QST*, May 1982, pp 21-27.

D. J. Healey, "High-Frequency Crystal Filters for SSB," *QST*, Oct 1960, pp 35-41.*

W. Herzog, "Narrow-Band Quartz Crystal Filters with Losses," *Proc. Institute of Radio and Electronic Engineers of Australia*, Jan 1968, pp 18-24.

H. Ingwersen, "Crystal Filters for FM Receivers," Technical Correspondence, *QST*, Apr 1971, p 48.

D. I. Kosowsky, "High-Frequency Crystal Filter Design Techniques and Applications," *Proc. IRE*, Feb 1958, pp 419-429.

J. Perolo, "Practical Considerations in Crystal Filter Design," *ham radio*, Nov 1976, pp 34-38.

J. K. Stevenson, "Simple Design Formulas for Third and Fourth Order Crystal Lattice Filters with Finite Attenuation Poles," *International Journal of Electronics* (London), Jun 1975, pp 697-710.

B. M. Singhi & M. R. Datta, "Modified Crystal Selective Networks," *Journal of the Institution of Electronics & Telecommunications Engineers*, (New Delhi, India), Sep 1975, pp 495-498.

B. H. Vester, "Surplus-Crystal High-Frequency Filters," *QST*, Jan 1959, pp 24-27.*

C. E. Weaver and J. N. Brown, "Crystal Lattice Filters for Transmitting and Receiving," *QST*, Jun 1951, pp 48-51.*

A. I. Zverev, *Handbook of Filter Synthesis*, (New York: John Wiley and Sons, 1967), pp 160-163 and 414-498.

"How to Homebrew Your Own Crystal Filters," *73 Magazine*, Sep 1979, pp 56-58.

Bits

SPRAGUE PPS FILM CAPACITORS

Sprague has introduced Polyphenylene-Sulfide (PPS) film-dielectric capacitors in two varieties: Type 880P metallized-film capacitors and Type 882P film and foil capacitors. PPS capacitors offer low loss and good high-temperature stability. The specifications of the 880P and 882P varieties are shown below.

	Type 880P	*Type 882P*
Temp range (°C)	−55 to +150	−55 to +125
Size range (in.)	0.017 × 0.4 to 0.470 × 1.75	0.21 × 0.438 to 0.38 × 1.00
Capacitance range (μF)	1 @ 400 V dc to 10 @ 30 V dc	0.001 @ 200 V dc to 0.22 @ 200 V dc

Standard capacitance tolerances for both types are +10%, +5%, +2% and +1%. Information on pricing and availability can be obtained from Sprague, PO Box 9102, Mansfield, MA 02048-9102. Technical information (Data Sheets 2880 and 2882) are available from Sprague's Technical Literature Service at the address given above. *—Rus Healy, NJ2L*

New Teledyne CMOS Dual MOSFET Drivers

Teledyne Semiconductor, Inc has announced the availability of a family of low-cost, dual 1.2-A MOSFET drivers that are priced below discrete MOSFET drive circuits. Just one of the low-cost drivers from the TSC1426-1428 family (along with one decoupling capacitor) can replace eight or more discrete components, thereby reducing the total number of parts on the circuit board. This results in a savings in design time, an increase in system reliability and cost-cutting.

The TSC1426-1428 family has dual 1.2-A outputs, and is available in inverting and noninverting configurations. The devices are pin compatible with the industry standard TSC426-28 MOSFET drivers. The TSC1426 is pin compatible with National Semiconductor's DS0026.

The Teledyne devices can be driven from TTL, CMOS or other sources of 2.4 to 18 V without the need for external speed-up capacitors or a resistive divider required by other devices. The TSC1426-1428 family is fabricated on an epitaxial layer that makes them latch-up proof.

This MOSFET driver family is available in 8-pin plastic DIP and surface-mount (SO) packages. Pricing begins at $1 each (in quantities of 100) for the TSC1426. Production quantities are available from stock. The low-cost devices have a 0- to 70°C temperature range.

For more information about the TSC1426-1428 family or any of Teledyne Semiconductor's other products, contact Rich Clarke, Teledyne Semiconductor 1300 Terra Bella Ave, Mountain View, CA 94039-7267, tel 800-888-9966. *—Paul K. Pagel, N1FB*

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Reflector Antennas: Dish Feeds and Recommendations

Last month's column contained a very general discussion of reflector antennas, with emphasis on parabolic-dish antennas. We learned that, for maximum gain, a parabolic-dish surface with a particular shape (f/D ratio) must be used in conjunction with a feed antenna that properly matches (illuminates) the dish. This month I will discuss dish feeds, considering the advantages and disadvantages of certain types. I'll stay away from a lot of math, with the intention of imparting system-level understanding. For a more rigorous treatment of the subject of dish antennas and feeds, and some basic cookbook designs, see the references at the end of this article.[1]

As discussed last month, there are a few important factors to be considered when matching a feed antenna to a parabolic dish for maximum gain. First of all, the pattern of the feed antenna must be such that the dish is fully illuminated with energy, with the minimum amount of spillover. This requires that the edge illumination be about 10 to 13 dB below that at the center of the dish.

Edge illumination is determined by two factors. The first is, of course, the pattern of the feed antenna. The second is a factor known as space loss, which simply causes power density at the edge of a dish to be less than at the center, as a result of the edge of the dish being farther from the feed than the center of the dish. This effect is more pronounced as the f/D is decreased. For instance, the space loss at the edge of a focal-plane dish is 6 dB, but that of a dish with f/D of 0.6 is 1.5 dB. Because we rarely try to feed a focal-plane dish, the space loss concern is a secondary one. Space loss is between 1 and 2 dB for most dishes. We must look for a feed antenna with a pattern that is 8 to 10 dB down at the angle off the dish axis subtended by the rim of the dish.

The second factor to be considered is selecting a feed antenna with equal patterns in the E and H planes. It is important to keep the edge taper at the required level in both vertical and horizontal planes. Because most dishes we would be inclined to use are symmetrical (round), this requires the E- and H-plane antenna patterns to be symmetrical. This is not usually the case with simple feed antennas, so there are some obvious trade-offs here! The E- and H-plane patterns of an antenna aren't always the same, and the difference *is* important. The third factor to be considered when trying to eke every last dB out of a particular dish geometry is blockage of the antenna aperature by the feed-antenna structure (more trade-offs!).

Most simple feed antennas are quite small in cross-sectional area, and therefore don't block the dish very much. More sophisticated feeds can illuminate the dish more evenly, but are often much larger and require more substantial support structures. (Remember, the feed support blocks the dish aperture, too.) For very large dishes, loss resulting from feed blockage is insignificant, but it can be a significant factor when using a smaller reflector.

All the above factors affect the gain of a dish antenna and are all a part of the efficiency in the dish-gain equation presented last month. The gain of a parabolic dish antenna at a particular frequency is directly proportional to the product of reflector area and efficiency, η. The number we most often give for η is 50% and, the above discussion shows why η isn't closer to 100%.

Practical Dish Feeds

So, you found this dish at a flea market, and you want to use it at some frequency or other. Where do you start? First, determine whether it is best suited for use as an antenna or some sort of planter or birdbath for the backyard! See Table 1 in last month's column, and determine the approximate gain of the antenna at the frequency in question. If the gain is less than that of a simple Yagi, come up with a higher-frequency application or get out the potting soil and petunias right now. If the gain figure is encouraging, measure the dish depth and plug this number and the diameter into Eq 2 from last month's column, and determine the focal length. Now you have f and D and can determine f/D. If this is much below 0.35, the dish will be rather hard to feed efficiently, though it will have a beautiful pattern (read: nice, low side and back lobes). The result of under-illuminating such a dish will be lower-than-optimum gain (resulting from low efficiency). If your dish is a TVRO type, the f/D ratio is probably between 0.4 and 0.45. Although such a dish isn't the easiest to feed, it can be fed effectively if you are willing to do some work. If the f/D ratio comes out in the 0.5 to 0.7 range, the dish can be fed easily, but the feed antenna will be slightly harder to support and the pattern will not be as clean as with a deeper dish. So, what are the dish-feed choices?

There are two or three basic feed-antenna configurations commonly used by amateurs. The first is an open-ended waveguide radiator, or horn antenna. These antennas can be rectangular or circular. The old familiar "coffee-can feed" (Fig 1A) is an example of a waveguide feed. The second type of feed antenna in common use is a dipole or a combination of dipoles with a reflector. The reflector can be a rod or wire, a screen or a solid-plane reflector. The third common feed-antenna category is an extension of the second—a Yagi or log-periodic feed antenna.

The simplest feed antennas are the circular waveguide horn (Fig 1A) and the dipole with reflector (Fig 1B), commonly known as a "splasher."[2-5] These antennas are small and easy to build, but suffer from having very dissimilar E- and H-plane patterns. If you just want to get on a particular band, though, either of these feeds will do the job. Both have H-plane patterns which are much broader than their respective E-plane patterns. This type of feed is usually chosen to properly illuminate a dish in the H plane and under-illuminate it in the E plane. Dish gain is reduced, but the pattern is relatively clean.

One simple feed antenna that has a better match between E- and H-plane patterns is a full-wave loop (round or square) placed over a reflector. This structure has a bit more aperture blockage than the previous two antennas, but the more symmetrical pattern more than makes up for the additional blockage.

Some more-complicated feeds have symmetrical E- and H-plane patterns and much cleaner patterns. In some cases, they have patterns with a slight null in the

[1]Notes appear on page 14.

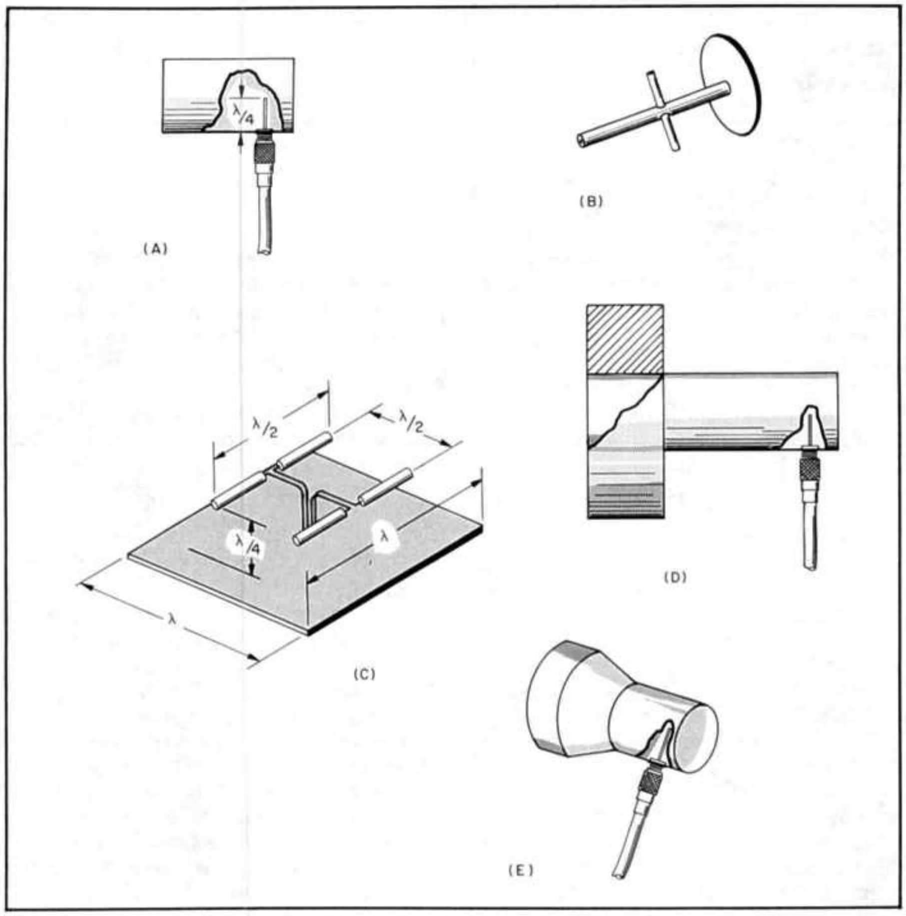

Fig 1—Several types of dish feeds are commonly used by amateurs. At A, a circular-waveguide (coffee can) feed; at B, a dipole feed with "splasher"; at C, an EIA feed; at D, a circular-waveguide feed with a single scalar-ring choke; at E, a dual-mode horn.

center. (Most of the simple feed antennas have what is known as a *cosine pattern*—that is, power density is greatest along the boresight and falls off to either side. The greatest power density is where it does the least good, right where it will get blocked by the feed structure).

Modern, efficient feed designs have patterns tailored to fit inside the dish and around the feed. (This is a good trick!) What are these "better" feed designs? The EIA feed antenna[6] consists of a pair of half-wave dipoles spaced ½ λ apart, both over a 1-λ-square plane reflector (Fig 1C). This structure is much more symmetrical than a simple dipole over a reflector, has almost identical E- and H-plane patterns, and is ideal for illuminating a dish in the 0.45 to 0.5 f/D range. Rectangular-horn feeds with unequal sides can be constructed to have identical E- and H-plane patterns, but are large and complicated to build, so they're not used much by amateurs.

Circular-waveguide horns with choke flanges are probably the best antennas for feeding dishes in the 0.45 f/D range at frequencies above 900 MHz (Fig 1D). (This is the type of antenna used in TVRO systems at 4 GHz, at efficiencies approaching 70%.) Barry, VE4MA, has done extensive work with circular-waveguide feeds with scalar-ring chokes at 1296 and 2304 MHz. His results with modest-size TVRO dishes on EME on these bands indicate illumination efficiencies of more than 55%![7]

Another circular feed-horn type that has gained wide acceptance on 1296-MHz EME is the W2IMU dual-mode feed horn (Fig 1E).[8] This feed horn, while complex, surpresses some of the higher-order waveguide modes, and in so doing exhibits nearly identical E- and H-plane patterns. This horn has a pattern that is optimum for dishes with f/D ratios in the 0.5 to 0.6 range.

Circular Polarization

The requirement for circular polarization (CP) for satellite and EME work makes feed design even more tricky. For this, we need a symmetrical feed such as a circular waveguide or crossed dipoles. Any of the circular coffee-can-type feeds can be built for circular polarization by inserting two probes into the can perpendicular to each other (one vertical and one horizontal), and feeding them with equal-amplitude signals 90° out of phase. Efficiency depends on the feed having the correct pattern and similar E- and H-plane patterns. Dish antennas make simple CP antennas, because it is usually simpler to build a circularly polarized feed than, say, two separate cross-polarized Yagis. The rules as outlined in the August 1988 column apply. Remember, however, that for RHCP, the feed antenna must be set up for LHCP, because there is a sense reversal when the wave is reflected from the dish surface.

Broadband Feeds

Because a parabolic reflector is not tuned, it can be used at many frequencies. Obviously, dipole-type feeds have inherently narrow bandwidths—even waveguide horns excited by quarter-wave monopoles are narrowband structures. Log-periodic arrays are often used in commercial applications to cover a multi-octave frequency range with a single feed. This approach is something of a compromise, because the phase center (discussed later) of a log-periodic feed changes with frequency, but it is one way to build a high-gain, broadband directional antenna. Peter, OE9PMJ, described a 1.2- to 2.4-GHz horn feed in *VHF/UHF and Above Information Exchange*.[9] This is a rectangular waveguide horn with a broadband coupling loop.

To cover more than one ham band with a single dish, the antenna doesn't need to cover all the spectrum in between, so amateurs often mount more than one feed horn in the same dish (this is possible unless using a single feed line is important). The highest-frequency feed is placed at the focal point, and the others beside this feed. There is a "squint" to the pattern when the feed antenna is not right at the focal point, but gain is usually not sacrificed. Of course, there's more blockage of the aperture, but that is a reasonable compromise if you want to operate more than one band with the same dish (sure beats *two* dishes!).

Phase Center

When installing a feed antenna at the focal point of a dish, what part of the feed goes at the focal point? The phase center is the point from which—as far as the dish is concerned—the feed-antenna energy appears to come. The phase centers of two common feed antennas are shown in Fig 2.

A little experimentation to find the best position for the feed antenna is usually worthwhile, because the phase center is often different in the E and H planes, and sometimes varies with frequency. The best approach is to get the feed antenna close to the dish focus, then tune for "maximum smoke." (After all, you may not even be able to determine the exact location of the

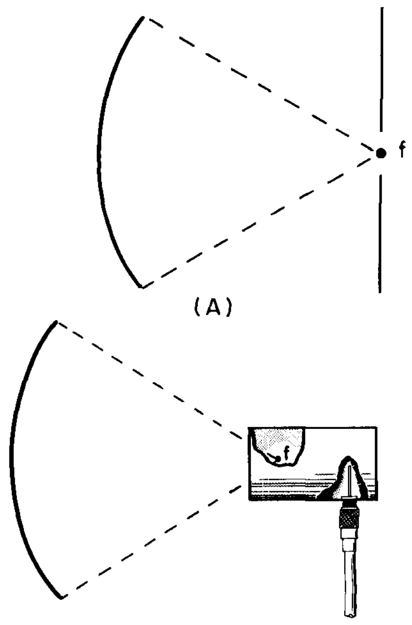

Fig 2—Phase center (f) is a point in a feed antenna from which RF energy appears to emanate. At A, the phase center for a single- or dual-dipole feed; at B, the phase center for a horn feed.

focus of the dish!) Most dish users mount the feed antenna such that the feed can be moved to optimize dish gain.

Next month, I'll discuss dish construction and surfaces, as well as compromises regarding surfacing material, wind loading, and so on.

Notes

[1]J. Reisert, "VHF/UHF World: Reflector Antennas, Part 1," *ham radio*, Feb 1986, p 51; Part 2, *ham radio*, Mar 1986, p 68.
[2]N. J. Foot, "Cylindrical Feed Horn for Parabolic Reflectors," *ham radio*, May 1976, p 16.
[3]N. J. Foot, "Second Generation Cylindrical Feed Horns," *ham radio*, May 1982, p 31.
[4]N. J. Foot, "Cylindrical Feed Horns Revisited," *ham radio*, Feb 1986, p 20.
[5]G. R. Jessop, ed, *RSGB VHF/UHF Manual*, fourth edition (Hertfordshire: RSGB, 1987), pp 9.63-9.77.
[6]J. Reisert, "VHF/UHF World: Measuring and Determining Antenna Performance," *ham radio*, May 1984, p 110.
[7]B. Malowanchuk, "Use of Small TVRO Dishes for EME," *Proceedings of the 21st Conference of the Central States VHF Society* (Newington: ARRL, 1987), p 68.
[8]R. H. Turrin, "Dual-Mode Small Aperture Antennas," *IEEE Transactions on Antennas and Propagation*, Mar 1967, p 307.
[9]P. Riml, "Wide Band Horn, 1.2-2.4 GHz," *VHF/UHF and Above Information Exchange*, Vol 3, no. 6, Jun 1986.

Bits

IMPROVED RECEIVER PREAMPLIFIERS

Electron Processing, Inc, has announced the first models of a new line of receiver preamplifiers. The RFP-40 answers the need for a high-quality, yet inexpensive, preamplifier that covers the MF to UHF bands. An RFP-40 provides 15 dB of gain from 1 MHz to 1300 MHz with a 2.8-dB noise figure. Available with a choice of BNC, SO-239 or F connectors, and powered by either 120-V ac-line-operated wall transformers or 12-V dc supplies, these preamplifiers are easily tailored to almost any installation. For an additional $5 per unit, you can have your choice of N, SMA, TNC, Motorola or mini UHF connectors and the ability to operate from 24-V dc power sources.

The RFP-40 is housed in a rugged, 1¼ x 1¼ x 3.5-inch cast aluminum enclosure and equipped with improved lightning/static protection. A one-year limited warranty on the RFP reflects EPI's assurance of quality.

Prices for RFPs equipped with either BNC, SO-239 or F connectors start at $69.95; quantity discounts are available. To order, or obtain additional information, contact the Sales Department, Electron Processing, Inc, at 516-764-9798, or write EPI at PO Box 708, Medford, NY 11763. —*Paul K. Pagel, N1FB*

DUAL SLOPE A/D CONVERTER WITH EXPANDABLE ANALOG MULTIPLEXER

Teledyne Semiconductor has announced a 12-bit, microprocessor-compatible, dual-slope A/D converter that includes an on-board analog multiplexer (MUX). The on-chip MUX may be configured for either four differential or eight single-ended channel operation--a key feature for process-control applications.

Designated the TSC804, the A/D converter can accommodate additional multiplexers in parallel with its own, enabling it to address even more channels. Both the ADC and the MUX are microprocessor controlled, providing convenient digital interfacing to fit a variety of system configurations.

The TSC804 features fast overload recovery, selectable conversion rate and the inherent noise immunity of an integration converter. A crystal-controlled clock and internal voltage reference make the TSC804 simple and reliable to use.

Two flexible modes of digital interfacing are provided. The Direct Mode is a fully complemented microprocessor interface that supports either an 8- or 16-bit data bus and allows the microprocessor to control how data is transferred. The Handshake Mode is an alternative means of interfacing the TSC804 to digital systems. It allows the TSC804 to become active in controlling the flow of data, and provides a direct interface between the TSC804 and standard UARTs, with no external logic required. The TSC804 provides all the control and flag signals necessary to sequence the data into the UART and initiate the serial transmission.

Similar in operation to Teledyne's popular TSC7109A, the TSC804's advanced CMOS design offers minimal power consumption, high noise immunity and high reliability.

The TSC804 is available in a 68-pin plastic-leaded chip carrier (PLCC) and a 60-pin plastic small-outline (SO) surface-mount package. Production quantities are available from stock. Prices begin at $10.95 each in quantities of 100. For more information about the TSC804 or any of Teledyne Semiconductor's products, contact the company at 800-888-9966, or write Ted Dabney, Teledyne Semiconductor, 1300 Terra Bella Ave, Mountain View, CA 94039-7267. —*Paul K. Pagel, N1FB*

AO-13 HANDBOOK

The *AMSAT-OSCAR-13 Handbook*, written by Richard Limebear, G3RWL, and published by AMSAT-UK, is now available. *The AO-13 Handbook* covers many aspects of the newest amateur spacecraft in detail. Chapter titles include: Brief History; History of AO-13 and the Ariane Rocket; Lauch/Positioning Typical Session; Bandplan; Operating Events; General Information; Operations (Scheduling, Communications, Time Delay, Doppler, Station Requirements, TX System, RX System, Typical Stations); Orbital Predictions; Tracking; Satellite Hardware (Transponders, Attitude Control, Beacons, Mode B, L, JL and S); Telemetry and data formats (RTTY, BPSK, decoding equations); Information Sources; Reference Books; The Future; AMSAT Organizations; Acknowledgments.

The booklet is over 50 pages long, and AMSAT-UK has made arrangements with Project OSCAR to release the book in North America. Those outside of North American should contact AMSAT-UK directly. *The AO-13 Handbook* can be obtained for $12 (this includes shipping from the UK) from Project OSCAR, PO Box 1136, Los Altos, CA 94023-1136. For more details, please send an SASE to Project OSCAR. [Tnx Ross Forbes, WB6GFJ, President, Project OSCAR]. —*Rus Healy, NJ2L*