WIDEBAND RF POWER AMPLIFIER

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Wideband RF Power Amplifier

This Amplifier Operates Over A Wide Range Of Supply Voltages.

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A single amplifier covering frequencies from HF to VHF at a power output level of 300 watts would have been considered impossible or impractical a few years ago. This would still be true if not for the advances in power FET technology.

This article covers the design aspects of a 300 watt unit with a frequency range of 10 to 150 MHz.

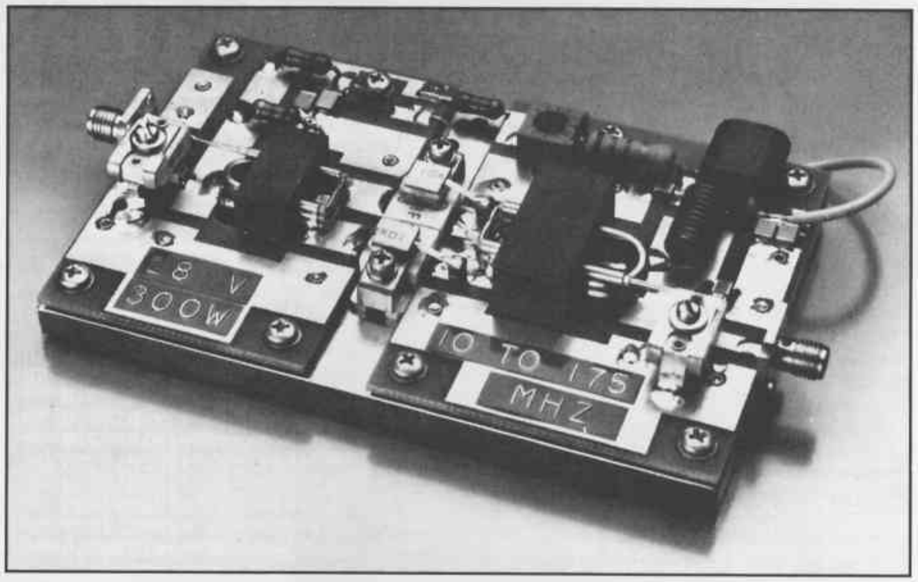


Figure 1. Overall view of the 300 watt, 10-150 MHz amplifier. Separate circuit boards are used for the input (left) and the output.

The MRF141G, used in this design, is housed in a special push-pull header commonly known as "Gemini" (twins), meaning that there are two identical transistors mounted next to each other on a common carrier or a flange. There are transistors (mainly FETs) available in the Gemini type packages rated from 20 watts to 300 watts. The lower power units can be used to frequencies of 1 GHz and higher, while the 100-150 watt units are designed to operate up to 500-600 MHz.

The advantages of a push-pull package such as the Gemini become apparent at higher frequencies, where the normal push-pull configuration with discrete devices would be impractical. In the push-pull circuit configuration the critical factor is the mutual inductance between the two push-pull halves, and not the device to ground inductance, as is the case in single ended designs. The Gemini or any other push-pull transistor housing permits the minimization of the mutual inductance to a level that approaches the ultimate in physical terms.

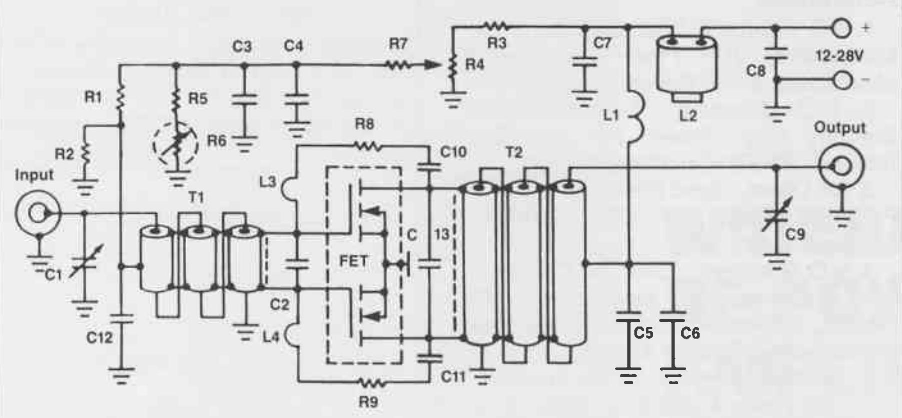
There are a couple of penalties we must pay for all this. One is a slightly higher cost when compared to two discrete units due to matching procedures involved and lower production yields resulting from double the possible reject rate. Another one is the reduced thermal characteristics. Twice as much dissipated power is concentrated virtually in the same area as in the case of a discrete design, leading to special cooling requirements.

About Power FETs

There have been designs of high power HF amplifiers using the T0-3 packages, and lower power versions with the T0-220 plastic units. With a given die geometry, a FET has approximately four times higher unity gain frequency than a bipolar transistor. This explains the fact that even the larger low frequency power FETs may have 10 dB or more power gain at 30 MHz, where a similar bipolar counterpart would be totally unusable. The difference is mostly in the figure of merit of the die itself, which is the ratio of feedback capacitance to the input capacitance or impedance. (This should not be confused with the more common base area/emitter periphery figure of merit die design formula.) With bipolar transistors the feedback capacitance (collector to base) is not usually specified, but it is 15-20 times higher than the drain to gate capacitance of a comparable FET, while the base/gate input impedances become about equal at increased frequencies. This feedback capacitance normally produces feedback within the device itself, whose exact phase angle depends on the capacitance values and other parameters.

In FETs designed specifically for RF, the die geometry is usually finer (larger ratio of the gate periphery to the channel area) than in the switching power FETs. This reduces the device capacitances automatically. Further reduction is achieved by splitting the die into a multiple of cells (groups of source sites and gate fingers) where the gates and sources are connected in groups of two or four by individual bonding wires to the common package terminals. For example, in the MRF141G one of the two die consists of 36 cells each having around 70 individual small FETs, making the total about 2,500.

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



R1 — 1K/½W
R2 — 10K/½W
R3 — 330 Ohms/2W
R4 — 1K Trimpot
R5 — 6.8K/¼W
R6 — Thermistor, 10K at 25°C/2.5K at 75°C
R7 — 2K/½W
R8, R9 — EMC Technology Type 5310 or KDI Pyrofilm Type PPR 515-20-3 Power Resistors, 25 Ohms
C1, C9 — ARCO 402, 1.5-20pF Compression Mica or equivalent
C2 — 200pF Ceramic Chip
C3 — 0.1uF Ceramic Chip
C4, C5, C12 — 1000pF Ceramic Chip
C6, C7 — 0.01uF Ceramic Chip
C8 — 0.47uF Ceramic Chip or lower values in parallel for the value indicated
C13 — 300pF, 2x150pF or 100pF and 200pF Ceramic Chips in parallel

L1 — 10 Turns AWG #16 Enameled Wire, 5mm I.D.
L2 — Ferrite Beads over AWG #16 Wire, 1.5 uH total
L3, L4 — Lead Lengths of R8 and R9, 20mm total
FET — MRF141G
T1 — 9:1 RF Transformer. (15-25 Ohm, 1.2-1.6mm O.D. semi rigid coax.) Magnetic Core Required: Fair-Rite Products Corp. #9461012002/9361021002 E and I (2 ea.) Types, Respectively, or Equivalent.
T2 — 1:9 RF Transformer. (15-25 Ohm, 1.8-2.3mm O.D. semi rigid coax.) For Magnetic Core See Text.

Note: All chip capacitors of 0.01uF and less in value are ATC type 100 or equivalent.

Figure 2. Schematic of the amplifier.

similar transconductances.

Another fact to mention is the gate resistance. Most modern power FETs use a gate structure of polycrystal silicon, which can have a bulk resistance comparable to carbon. It is also used as a conductor between the metal pattern and each individual gate. In the RF power FETs, each gate is fed through a separate contact having a resistance of approximately 0.1 ohms. In the 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 still results in at least 30-40 times higher gate resistance with a die of comparable size.

In high frequency applications the high gate resistance permits a part of the drain-source RF voltage or transients to be fed back to the gate through C_{RSS} in amplitudes that can rupture the gate-source oxide layer. The rupture will first occur in the far end of the die, away from the gate terminal. Since the gate resistance is internal to the FET die, external limiting or clamping circuits at the gate are of no help. The gate of a MOSFET is the most sensitive part of the device, which can be permanently damaged even by static charges during the handling. Although the larger FETs (100-150 W), due to their higher gate capacitance, are not as vulnerable as the smaller ones, proper precautions should be exercised.

Design and Construction

As discussed earlier, the common mode inductance in a push- pull circuit is not critical, and the ground path is only used for DC feed to the amplifier. The input and output impedance levels are established from gate to gate and drain to drain respectively. This allows the circuit board, which is made of the standard 1.6 mm G10 material, to be split into two sections. One carries the input matching network and part of the bias circuit, while the second section holds the output matching network, the bias set and the drain voltage feed and filtering circuitry. (See Figures 1 and 2). In addition to allowing wider design flexibility, this arrangement also simplifies the repair and maintenance of the unit, if required.

The two circuit boards including the space between them for the FET measures 115 × 75 mm. They are mounted on a copper plate with the same dimensions having a thickness of 6 mm. The input and output connectors (SMA) are mounted to the edges of the copper plate. They can also be placed at a remote location with coax connections to the amplifier utilizing any connectors that have good RF characteristics such as BNC.

Due to the large amount of heat concentrated in a small area in the form of dissipated power, it is important that the copper plate be employed as a heat spreader unless the heat sink itself is made of copper. The heat spreader can then be bolted to a piece of aluminum extrusion with thermal resistance of 1° C/W or less for normal intermittent operation without forced air cooling. The heat spreader and the extrusion surfaces should be flat without any burrs, and silicone thermal compound must be applied to the interface. The same practices should be followed in mounting the FET into the heat spreader. If the FET gate and drain leads are bent sharply up along the package sides, they will be aligned along

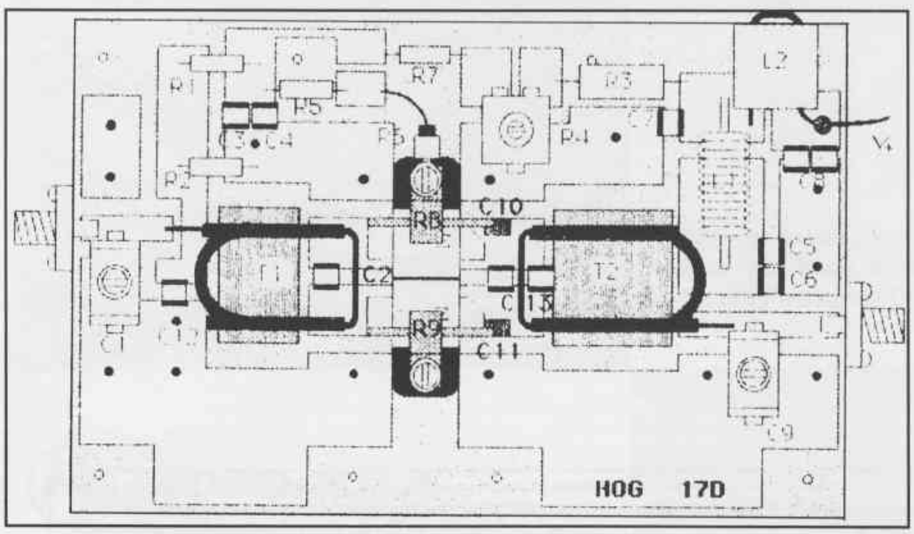
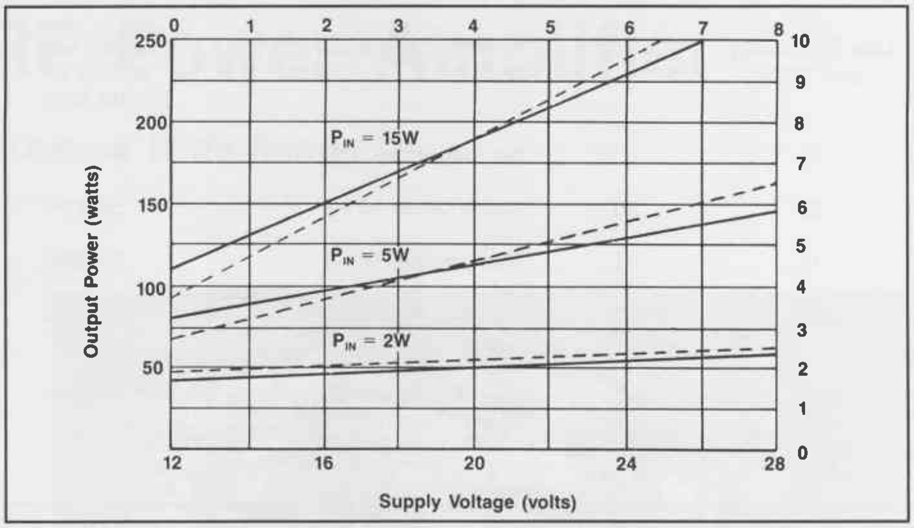


Figure 3. The component layout diagram. The only critical component locations are those of C2 and C13. They must be soldered in place (½ of C13) before the mounting of the input-output transformers.

the edges of the circuit boards. This makes the board spacing from the heat spreader less critical, which then can be anywhere from 1 to 3 mm. The FET lead lengths to the board connection points are variable by the same amount, but they have a minimal effect on the impedance matching and performance at these frequencies.

Details of the electrical design concepts of a similar amplifier are given in reference 1. The input-output transformers require a special low impedance semi-rigid coax cable making construction difficult in single quantities. The output transformer only requires a magnetic core if operation below 75 MHz is desired. In contrast, the input transformer always requires one regardless of the frequency of operation. In a push-pull FET amplifier design the gates of the two halves must be isolated by sufficient inductance or resistance (7,8). In order to prevent instabilities which will occur at the resonant frequency of the device capacitances, the internal wire bond inductances and the external inductances, sufficient isolation is required between the two gates which the magnetic core will provide. Without this, the two FETs of the push-pull circuit would see a parallel connection at some resonant frequency, which would result in serious instability problems.

The importance of the negative feedback (L3, L4-R8, R9-C10, C11) must be emphasized. Without it the power gain would exceed 30 dB at low frequencies, resulting in increased conditions for instabilities. The feedback is designed to lower the low frequency power gain close to the 150 MHz level it is at. L3 and L4, which consist of the lead lengths of R8 and R9 represent a reactance of 20 ohms each at 150 MHz. It also controls the frequency-amplitude slope. This in series with the 25 ohm resistor values lowers the power gain by one dB at 150 MHz but increases to as much as 15 dB at 10 MHz. C10 and C11 are only used for DC blocking and their values are not critical as long as their reactances are less than 10-15 percent of R8+R9. C10 and C11 are ceramic chip capacitor that are mounted vertically on the circuit board (Figure 1). Although unusual, it allows the feedback resistor leads to be soldered directly to the capacitor top terminals. This provides a much lower inductance path than the conventional mounting technique and saves board space. Since R8 and R9 must be able to dissipate up to 15 Watts each depending on the frequency of operation, they must be of a type that can be easily heat sunk. The type resistors designated



Fgure 4. Amplifier power output versus the supply voltage at various input levels. Solid lines represent 150 MHz and dashed lines 10 MHz.

have mounting lugs which are terminally connected to the copper heat spreader through 5 mm high spacers.

These are mounted on top of the ends of the FET flange, allowing the use of common screws for fastening the resistors and the FET. The spacers must be of material with low terminal resistance like aluminum, brass or copper, and must have a larger surface area than thin wall tubing. A couple of stacked brass nuts, one size larger than the mounting screws is a good solution. Although not very professional it works rather well. If the unit is used for other than intermittent modes of operation such as voice communication, a thermistor (R6) can be used for bias stabilization. Without it the drain idle current will approximately triple if the FET case temperature is doubled, and would result in decreased efficiency. The thermistor can be attached to a solder lug, which is fastened with one of the resistor-FET mounting screws.

The input and output impedance matching is achieved with unique wide-

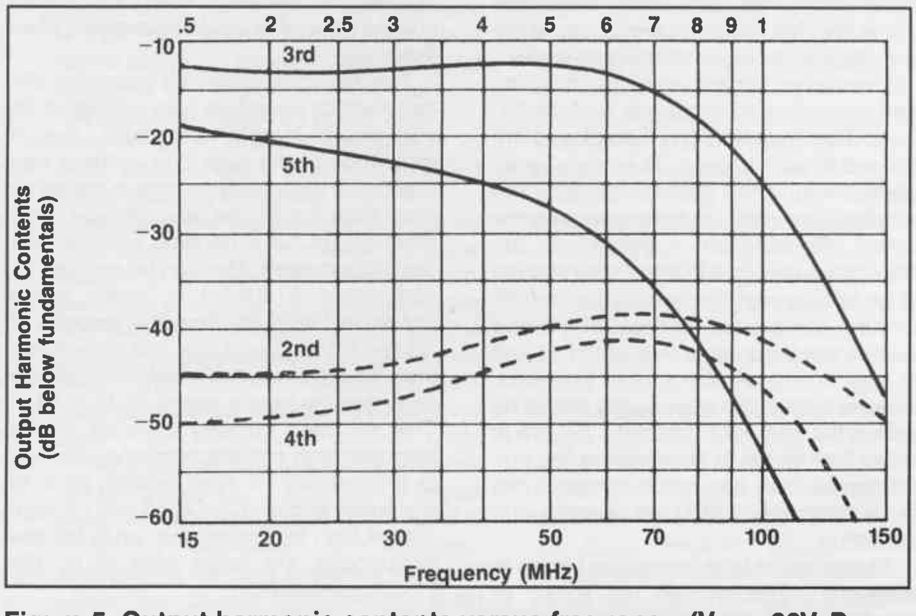


Figure 5. Output harmonic contents versus frequency. (V_{DS} = 28V, P_{OUT} = 300W.) The benefit of the push-pull configuration can be seen in the suppressed even order products. The data does not change considerably with varying the supply voltage or power output.

Low Frequency end MHz	u_1	Manufacturer and type #	Drain Eff. at 300 W, 75-100 MHz
75	1	No magnetic core	62-66%
25	20	Micrometals 101-2	59-63%
15	35	Micrometals 101-8	54-59%
7.5	125	Fair-Rite Prod. Corp. 9461014002/9361020002	46-52%
2	850	Fair-Rite Prod. Corp. 9443014002/9343020002	36-43%

Table 1. Effect of the output transformer magnetic core material on amplifier bandwidth and efficiency.

band transformers described in References 1 and 2. Some of their advantages are: DC isolation between the primary and the secondary, automatic balanced to unbalanced function and compact size in comparison to the power handling capability. Their principle is the same as in ordinary low frequency transformers, except that tight coupling between the windings is achieved through the use of low impedance transmission line, in this case semi-rigid coax cable. The low impedance side always has one turn and consists of parallel connected segments of the coax outer conductor. The inner conductor forms the high impedance winding, where the segments are connected in series.

This arrangement only permits impedance ratios with integers such as 1:4, 9, 16. The magnetic cores employed are the old E and I types. They can be inserted around the transformer after the windings are made up and mounted to the board. Rectangular openings in the boards are required to allow the I section to be laid against the heat spreader with thermal compound interface. The E and I cores are then cemented together and to the edges of the board openings. Special heat conductive epoxy would be preferable, but not mandatory. If there is no air flow on top of the amplifier, the output transformer can reach temperatures in excess of 100°C in continuous operation.

As a rule, the high frequency losses in magnetic material such as ferrite or powdered iron, are more or less directly related to its permeability, and appear as heat generated within the core. Since this part of the RF energy is not delivered to the output terminal, and the drain current is equal in each case, the result is lowered overall efficiency.

From the above we can conclude that the magnetic core material should be selected according to the lowest desired frequency of operation. For example, from 2 to 150 MHz, initial permeability (u_i) of over 600 and cross sectional area of about 1 cm² would be required. Ferrites in this category have Curie temperatures of 130-140°C, above which temperature they become paramagnetic and causes serious malfunctions in the operation of at lower frequencies. In such case special cooling structures would be required (See Table 1).

The amplifier described was originally designed for operation from a constant 28 volt power supply, for which reason regulation of the gate bias voltage was omitted. If the supply voltage is varied by more than 2 volts, the bias will have to be reset by R4 for a nominal 400-500 mA drain idle current. This can be avoided by connecting a 6.8-8.2 V zener diode (1N5921A-1N5923A) from the junction of R3 and R4 to ground. The idle current can then be set once, and would not change considerably from a supply of 12 to 28 V. The V_{DS} feed circuitry consists of the standard high and low frequency filtering to prevent any RF from feeding back to the power supply. C5, C6, L1 and C7 handle the high frequency end, while the low frequencies are taken care of by the L2-C8 combination.

Performance

With the 1:9 impedance ratio output transformer employed, the optimum power output at 12 and 28 V supplies would be only 50 and 265 watts respectively.

$$P_o = \frac{2V^2_{DS} - V_{DS_{ON}}}{50/9}$$

At these power levels the IM distortion is better than −30 dB at all frequencies, the worst case being at 50-100 MHz. From Figure 4 it can be seen that higher output levels are possible with increased drive power, but the amplifier will be close to saturation and can be only used for nonlinear applications such as FM or CW. For the best IMD, the idle current should be 500-800 mA total, but disregarding the linearity, it can be as low as 100-200 mA. Lower idle current will result in loss of power gain by 0.5-1.0 dB, while increasing the efficiency.

The stability of any RF power amplifier (especially solid state) under mismatched load conditions is always a concern. The power MOSFETs have been proven superior in this respect to the BJTs, although the stability is also circuit dependent to a great extend. The stability of the amplifier described here has been tested against load mismatches using a simulator of 30:1 at all phase angles and a 3 dB power attenuator to the amplifier output, which results in approximately 3:1 VSWR. Unconditional stability was shown at a combination of any power output level and supply voltage at 10, 50 and 150 MHz. Stability into a 3:1 mismatched load is almost considered a standard specification in the industry, meaning that the harmonic filter-antenna combination (if applicable) should have its input VSWR equal or lower. Normally 2:1 is easy to achieve over a fraction of an octave bandwidth, unless the filters are improperly designed. Figure 5 shows that at 150 MHz and beyond the output harmonics are well suppressed to start with, but a filter is still required to meet the FCC regulations. More elaborate filtering is necessary at lower frequencies, where the 3rd harmonic is only 12-13 dB below the fundamental. For most industrial applications, however, harmonic filtering may not be necessary. Although data is not shown, the amplifier can be used up to 175 MHz with a power gain of 10-11 dB. C1 should be adjusted for lowest input VSWR and C9 for the peak power output at the highest desired frequency of operation.

As the MRF 141G basically operates from a 28 V supply, lowering the voltage down to 20 or below would make the unit almost indestructible against load mismatches in case of an open coax or broken antenna. Figure 4 shows that the power output is still almost 200 watts at

20 V and 150 watts at 16 V. The ruggedness criterion does not apply against possible transients to the input from the signal source and assumes that the FET is properly mounted to the heat sink. A normal guideline is that a transistor should have its break down voltage (BV_{dss}) 2-25 times the operating voltage. The break down voltage is set by choosing the starting material (silicon) with proper resistivity or doping. If the break down voltage is too low, the output voltage swing may exceed it and cause an avalanche. If it is too high, the transistor will saturate at a low power level, but it will be harder to blow up since the device is less likely to exceed its dissipation limits. For the same reason, devices made for 50 V operation are often used at 30-40 V and at reduced power levels in applications like laser drivers and magnetic resonance imaging, where they must momentarily withstand a large output load mismatch. The circuit boards and other components for this design are available from Communication Concepts, Inc., 121 Brown Street, Dayton, OH 45402. rf

References

1. H.O. Granberg, "Building Push-Pull, Multioctave, VHF Power Amplifiers," *Microwaves & RF,* November 1987.
2. H.O. Granberg, "New MOSFETs Simplify High Power RF Amplifier Design," *RF Design,* October 1986.
3. Bill Olson, "Solid State Construction Practices, Parts 1-3," *QEX,* A.R.R.L. January-March, 1987.
4. H.O. Granberg, "Good RF Construction Practices and Techniques," *RF Design,* September-October, 1980 and Motorola, Inc. Reprint AR164.
5. Motorola Semiconductor Products, Inc., "RF Power MOSFETs," Publication AR165S.
6. Edwin S. Oxner, "Controlling Oscillation in Parallel Power MOSFETs," Siliconix Technical Article TA83-2, September 1983.
7. David Giandomenico, et. al., "Analysis and Prevention of Power MOSFET Anomalous Oscillation," *Proceedings of Powercon II,* Power Concepts, Inc. 1984.
8. Motorola Semiconductor Products, Inc. "RF Device Data," Publication DL110, Rev. 2.
9. EMC Technology, Inc. "Microwave Components," Publication 386-15, 1986.

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