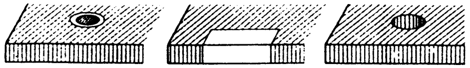
parasitics and grounds

During board layout, care should be taken to minimize all parasitics. Remember that extra lead length equals extra inductance added to the design. This is particularly important if the circuit is to be operated above 1 GHz. Transmission lines should, whenever possible, run flush to the package. This requires that a hole be made in the board so that the MAR-amplifier leads are in the same plane as the transmission line. MAR-amplifiers should be mounted on the etched side of the board to minimize the inductance of feedthrough connections. Abrupt changes in transmission line width also create parasitic effects, called step discontinuities. Although the complete model for such a discontinuity can become quite complicated, the overall effect of the step from an MAR-iplifier lead to a 50 ohm transmission line is typically .05 to .2 nH of extra series inductance. Tapering the transmission lines from 50 ohms down to the amplifier lead width helps minimize this effect. Bends in transmission lines also create parasitic effects and should be avoided when possible; when they must be used, the corners should be chamfered to prevent the bends from acting as extra shunt capacitance. (Reference: K.C. Kupta et all, "Microstrip lines and slot lines," Artech, 1979, p 140-142). The effects of parasitics on gain loss and VSWR, is shown in Table 2.

Ground planes should be kept as large and as solid as possible. Return paths for high frequency circulating currents must be kept as short as possible, especially at the emitter leads (MAR ground lead connections). If plated through holes are used as ground returns, they should be placed directly under the ground leads of the MAR and be located as near as possible to the body of the package .050 inches. Any additional path length acts as ries inductance, which translates into unwanted emitter resistance at operating frequencies. Gain, power compression, and high frequency rolloff will all be degraded if proper grounding techniques are not used. A gain decrease of more than 1dB can be expected at 1GHz for approximately 2nH of lead inductance. Fig. 3 shows good return paths between topside ground connections and the bottom ground plane. The effects of parasitic emitter inductance due to poor RF grounding is shown in Fig. 4, with emitter inductance of zero to 4nH.

table 2 Effects of Parasitics on VSWR and Gain

Freq. MHz	MAR-2, No Parasitics		0.9 nH Parasitics Only		MAR-2, + Parasitics	
	VSWR	Loss, dB	VSWR	Loss, dB	VSWR	Loss, dB
500	1.09:1	0.0	1.01:1	0.0	1.18:1	03
1000	1.23:1	.04	1.12:1	.01	1.39:1	11
1500	1.29:1	.07	1.22:1	04	1.46:1	15
2000	1.29:1	.07	1.30:1	07	1.45:1	15
2500	1.26:1	.05	1.38:1	11	1.45:1	15
3000	1.26:1	.05	1.45:1	15	1.53:1	19



a. Wire "feed through" soldered in place in hole drilled through board.

b. "Wrap around" of copper foil (may be used with slit in board).

c. Plated through hole (side walls of hole metalized).

figure 3

Methods of realizing minimal length return paths to ground.

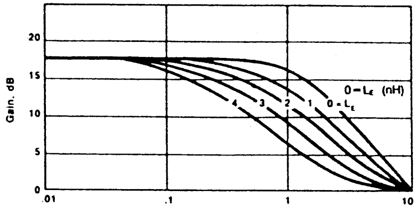


figure 4

Gain vs. frequency as a function of emitter inductance (L_E) for the MAR-1.

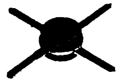
for 75-ohm systems

When an ideal 50-ohm unit is used in a 75-ohm system, return loss will drop to 14dB (VSWR 1.5:1) and mismatch loss will be 0.18dB at each port. In practice, the return loss change may be higher due to finite isolation. Table 3 shows the gain and return loss of a MAR-3 amplifier in a 75-ohm test system. Gain is about the same as in the 50 ohm system. Input and output return loss (75 ohm) is better than 9dB over most of the range. At high frequencies there is some improvement in input return loss probably due to the tuning effect of parasitics.

table 3 75-ohm Gain and Return Loss

Frequency (MHz)	Pin (dBm)	Gain (dB)	RL#IN (dB)	RL#OUT (dB)
30	32.99	11.72	9.78	11.61
1	33.15	11.62	9.72	11.62
12	33.54	11.64	9.86	11.69
102	33.88	11.57	9.75	11.67
254	34.18	11.45	10.18	11.63
495	34.59	11.29	11.06	11.23
1009	35.25	10.85	16.01	10.38
1507	35.50	9.94	21.23	9.99
2200	36.17	8.32	13.50	8.72

Barend's MAR boekje



Uitgave: Barend Hendriksen HF Techniek
Box 314, 7200AH Zutphen, tel.05756-1866 fax 05756-5012

De inmiddels niet meer weg te denken MAR MMIC breedbandversterkers die wij enkele jaren geleden introduceerden vormen een familie, overeenkomend met b.v. de Avantek neefjes. De meeste zijn simpel toepasbaar (de Amerikanen zeggen "drop-in") en geven prima resultaten van DC tot enkele GHz. In het schema is de HF smoorspoel alleen benodigd indien de voorweerstand een lage waarde heeft- de impedantie van R en RFC samen moet minstens 500 Ω zijn, bij de laagste te verwerken frequentie. Hoe hoger R, des te beter de temperatuurstabiliteit. Uit de tabel kunt u een type uitzoeken, gelet op ruis, intercept point, vlakheid freq./gain curve enz. enz. Bij hogere frequenties moeten de koppelcondensatoren een lage zelfinductie hebben; kleine keramische typen (of SMD) voldoen goed.

De toepassingen zijn legio, bedenk wel dat bij gebruik als voorversterker in ontvangers zelfs de beste breedbandversterker overstuurd zal worden; als regel moet u wat voorselektie toepassen.

Het is mogelijk de MAR's parallel of in balans te gebruiken. De MAR 8 is de enige die niet altijd stabiel is; u zult deze enigszins op 50 Ω moeten aanpassen. Doet u dit, dan heeft u een prima amp met veel versterking.

In ons BOUW BOEKJE-1 vindt u enkele aardige schema's met MAR's.

Tot slot nog dit: de MAV typen zijn identiek aan Avantek (de MAR's alleen inwendig) en dus geeft bij deze typen de stip de uitgang aan !

Model No.	Color Dot	FREQ. MHz	GAIN, dB Typical (at MHz) 100	500	1000	2000	(Note 4) MIN	MAXIMUM POWER, dBm Output (1dB) Compression	Input (no damage)	DYNAMIC RANGE NF dB Typ.	Intercept pt. dBm 3rd Order Typ.	VSWR In	VSWR Out	MAXIMUM RATING (25°C) I(mA)	P(mW)	DC POWER at Pin 3 Current (mA)	Volt Typ.
MAR-1	Brown	DC-1000	18.5	17.5	15.5	—	13.0	0	+10	5.0	15	1.5:1	1.5:1	40	100	17	5
MAR-2	Red	DC-2000	13	12.8	12.5	11	8.5	+ 3	+15	6.5	18	1.3:1	1.6:1	60	325	25	5
MAR-3	Orange	DC-2000	13	12.8	12.5	10.5	8.0	▲+ 8	+15	6.0	23	1.6:1	1.6:1	70	400	35	5
MAR-4	Yellow	DC-1000	8.2	8.2	8.0	—	7.0	+11	+15	7.0	27	1.9:1	2.1	85	500	50	5
MAR-6	White	DC-2000	20	19	16	11	9	0	+15	2.8	15	2:1	1.8:1	50	200	16	3.5
MAR-7	Violet	DC-2000	13.5	13.1	12.5	10.5	8.5	+ 4	+15	5.0	20	2:1	1.5:1	60	275	22	4
MAR-8	Blue	DC-1000	33	28	23	—	19	+10	+15	3.5	27	□	□	65	500	36	7.5
MAV-4		DC-1000	8.5	8.5	8.0	—	7.0	11	20	7.0	28	1.8:1	1.8:1	85	500	50	5.25
MAV-11		DC-1000	13	12.2	11.5	—	9.0	16	15	3.8	30	2.5:1	2.2:1	80	550	60	5.6

proper biasing calculations

In order to deliver full performance, MAR-amplifiers must be biased correctly. The internal resistive networks determine individual transistor operating points; all the user needs to do is present the proper voltage at the DC input terminal. For the purpose of bias stability over temperature, the internal transistors should have their bias supplied through a collector resistor (labeled R_C in Fig. 5). This resistor compensates for increases in device β (beta) with temperature by dropping the transistor's collector voltages whenever they try to draw more collector current. Coupled with this effect is the fact that the collector resistor will itself be changing in value over temperature.

Resistors with positive temperature coefficients such as the common carbon composite (+.0001% per degree C°) do an excellent job of compensating for the temperature drift of the negative coefficient on-chip resistors.

For bias stabilization over a temperature range of −10° to +100°C, a drop of at least 1.5 volts across the collector resistor is necessary. The larger this voltage drop, the more stable the bias will be.

For a fixed bias (constant quiescent current vs. temperature), gain will decrease as temperature increases. A voltage drop of about 2V across the collector resistor allows the bias swing over temperature to compensate for this gain change, yielding best gain flatness over temperature. The effect of bias stabilization resistor R_C on performance over a temperature range is shown in Table 4. Notice that the amplifier may self-destruct at high temperatures if no bias resistor is used.

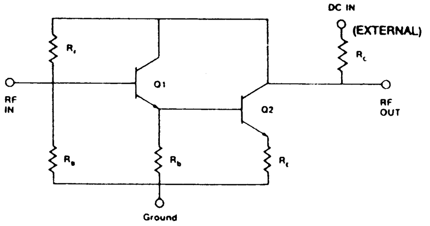


figure 5
General MAR AMP schematic.

table 4 Effects of R_C on performance over temperature.

MAR-1 Operating Voltage 5.07 V

Voltage Drop, volts	Resistor Value, ohms	Temperature degrees C	Bias Current, mA	Power Gain @ 100 MHz, dB
0	0	10	9.5	0.5
		25	18.4	18.8
		100	**	**
1.5	82	10	14.2	17.0
		25	17.3	18.3
		100	24.1	19.0
2.0	100	10	16.3	18.5
		25	18.9	18.9
		100	24.6	19.0
7.0	412	10	16.1	18.3
		25	18.8	18.1
		100	18.3	17.5

** Device destroyed due to excessive current draw

the value of the bias stabilization resistor R_C
is given by:

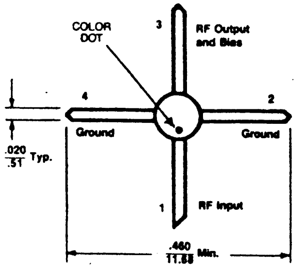
$$R_C = \frac{V_{CC} - V_d}{I_d} \text{ ohms}$$

where

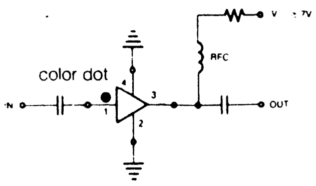
V_{CC} = the power supply voltage applied to R_C (in volts)
V_d = the voltage at the DC input terminal of the MMIC (in volts)
I_d = the quiescent bias current drawn by the MMIC (in amps)

The dissipation of this resistor is given by:

$$P_{diss} = I_d^2 \times R_C \text{ watts}$$



Typical Circuit Arrangement



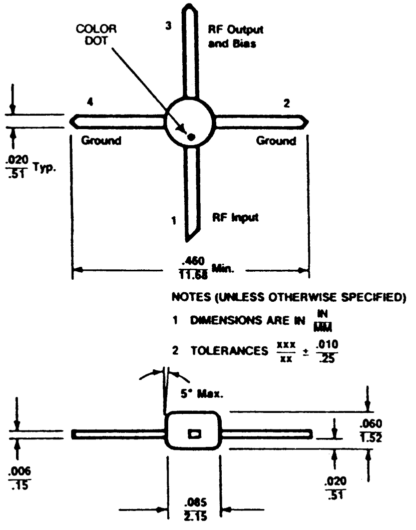


figure 1b

Since device β (beta) and, therefore, collector current, given a fixed bias tends to increase with temperature, R_C also serves as a temperature compensating element. A dc blocking capacitor must be placed at the amplifier input to isolate other devices and the input source.

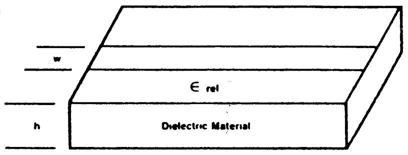
Since the internal resistive networks prematch both input and output to 50 ohms, the MAR-amplifiers are particularly easy to design with. To design an amplifier section, all that's needed is a 50 ohm microstrip line, blocking capacitors, and very simple bias circuitry; but there are some basic construction rules that should be followed.

board layout suggestions

In a typical microstrip structure, Fig. 2(a), line impedances are determined by strip width (w), board dielectric material (E_r), and dielectric thickness (h). Since the impedances of the MAR-units are prematched to operate in a 50 ohm system, microstrip lines should be as close to 50 ohms as possible to realize full specified performance. For various board materials, line width dimensions for a 50-ohm line are given in Fig. 2(b). Operation in systems with characteristic impedances other than 50 ohms is possible with somewhat reduced performance. MAR amplifiers offer very good return loss in a 50-ohm system.

The board material for the microstrip structure should be selected to suit the intended frequency of operation. PTFE woven-glass performs well to frequencies in excess of 2 GHz, is a fairly rugged material that can tolerate substantial rework, and is not particularly sensitive to heat or humidity.

Duroid is the favored material of microwave designers because of its high dielectric consistency and low dielectric dissipation. RT/duroid is a somewhat fragile material which crushes fairly easily; glues do not adhere well to its substrate so thin metallization patterns are subject to lifting if abused with repeated rework. Some versions can also be quite hydroscopic, and can show substantial dielectric shifts with variations in humidity. Because of these factors, care should be taken when working with the material.



Bottom side ground surface
(solid metalization)
microstrip structure

figure 2a

Line Widths for 50 ohm line for various board materials.

Board material	ϵ	Thick	w/h for 50Ω	w for 50Ω
RT/Duroid 5870[1]	2.3	.015″	2.90	.044″
PTFE-Woven Glass Fiber (Typ.)	2.55	.010″	2.55	.025″
		.031″	2.55	.079″
		.062″	2.55	.158″
Epoxy-Glass (G10)	4.8	.062″	1.75	.108″
Alumina/E10[2]	10.0	.025″	0.95	.024″
		.050″	0.95	.048″

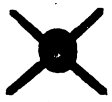
[1] Trademark of Rogers Corp. for its PTFE nonwoven glass PC material. (RT is reinforced teflon and PTFE is polytetraflourethylene)

[2] E-10 and Epsilam-10 are trademarks of 3M for its ceramic filled PTFE substrate.

figure 2b

- easy to use, 50 ohm input/output
- smooth response over the band
- easy for printed-circuit designs, one input and one output
- can operate as low as 5Vdc
- extremely broad bandwidth, usable up to 4GHz
- smooth response over entire band, no external resonances
- low impedance, less susceptible to EMI

MAR amplifiers



Door parallelschakeling kan het vermogen opgevoerd worden, de impedantie bij 4 stuks is dan 12,5 Ω, deze waarde kunt u met een geschikte trafo weer op 50 Ω brengen.
In de schetsen ziet u een balansschakeling, de voordelen boven parallelschakeling zijn: Sterke vermindering van even harmonischen, bijna viermaal het vermogen van 1 MAR plus goede bandbreedte en versterking. De andere schets toont een balansversterker met neutrodynisatie. In de getekende vorm, unilateralisatie geheten, worden zowel de reële als de imaginaire component gecompenseerd waardoor een versterker met goede input/output isolatie ontstaat.
Deze techniek, waardoor belastingsvariaties niet terugwerken op de ingang, is geknipt voor de MAR omdat diens interne terugkoppelelementen een zeer lage Q hebben.
Voor bovenstaande toepassingen is de MAR4/MAV4 en ook de MAV11 bij uitstek geschikt.

Tevens vindt u op de Engelstalige pagina's tips voor een verantwoord UHF/SHF ontwerp. Hoe hoger de frequentie, des te belangrijker grote massavlakken en korte verbindingen, maar ook b.v. taps toelopende soldeervlakken.

Laat u niet afschrikken door alle technische rimram, in de praktijk soldeert u het Marretje erin en Bingo! het versterkt...

N.B.
Daar er voor de MAR/MAV 2 en 3 betere alternatieve typen bestaan, voeren wij deze niet in ons programma.

($T_A = 25°C, I_d = 17$ mA)

	Freq. MHz	S_{11} (Input Return Loss)			S_{21} (Power Gain)			S_{12} (Isolation Out-In)			S_{12} (Output Return Loss)		
		dB	Mag	Ang	dB	Mag	Ang	dB	Mag	Ang	dB	Mag	Ang
MAV-4	100		.06	141	18.4	8.31	170	−22.3	.077	5		.07	9
	300		.10	94	17.8	7.75	151	−22.0	.079	15		.07	22
	500		.13	70	16.9	7.01	134	−21.0	.089	19		.07	37
	800		.16	41	15.4	5.87	114	19.5	.106	27		.08	53
	1000		.17	28	14.3	5.21	102	18.9	.114	29		.08	61
MAV-11	100		.04	−81	12.7	4.29	171	−16.4	.152	2		.05	−137
	300		.06	−105	12.4	4.18	156	−16.2	.155	4		.10	136
	500		.09	−124	12.1	4.01	141	−15.8	.162	6		.15	144
	800		.15	−147	11.3	3.68	120	−15.2	.174	7		.22	161
	1000		.18	−161	10.7	3.43	106	−14.7	.184	6		.26	173

Table 5 shows the recommended bias resistor values for MAR amplifiers.

table 5 bias resistor values for MAR amplifiers

Amplifier	Bias Current	Bias Voltage	Approximate Bias Resistor (Ohms)				Resistor Dissipation (Watts)
	I_B (mA)	· V_O	· 5V	· 9V	· 12V	· 15V	· V_{CC} 12V
MAR-1	17	~5	—	235	412	588	12
MAR-2	25	~5	—	160	280	400	18
MAR-3	35	~5	—	114	200	286	25
MAR-4	50	~6	—	60	120	180	30
MAR-6	16	~3 5	98	344	531	719	14
MAR-7	22	~4	45	227	364	500	18
MAR-8	36	~8	—	—	111	194	14
MAV11	60	~5.6			100		

DC blocking capacitors are used in both the RF input and output lines to isolate the resistive bias circuits from the source and load resistances. These capacitors will also put limits on the frequency response of the finished amplifier. Low frequency response will be determined by the capacitor's value; it must be high enough to be a reasonable RF "short" at the lowest frequency of operation. High frequency response will be limited to the frequency at which the capacitor's associated parasitic inductance becomes resonant with the blocking capacitor. Operation above the frequency may lead to highly unpredictable circuit behavior. Blocking capacitors with high Qs (Q defined as ratio of capacitive reactance to parasitic resistance) should always be used to minimize insertion losses. Fig. 6 illustrates the variations in VSWR as a function of frequency parasitic inductance, and value of blocking capacitor. For your design convenience, Mini-Circuits offers a full line of capacitors, shown on page 3.

An RF choke (Dale IM-2 or equivalent) should be used in series with the bias stabilization resistor. Although the choke is not generally needed to keep the RF out of the DC, it is needed to keep the stabilization resistor from appearing in parallel with the load circuit, and thus degrading the output match. A good rule-of-thumb is that the impedance of the choke at the lowest frequency of operation plus the value of the stabilization resistor should be at least 500 ohms. A 10 uH inductor works well as a choke at frequencies as low as 10 MHz; it can be either a molded inductor (for low-cost applications) or a chip inductor (in cases where space is at a premium). At higher frequencies, several turns of wire on a high permeability ferrite bead should be used. If the choke is omitted, expect a gain loss of between 0.5 and 1 dB and a decrease in $P_{1 dB}$ of as much as 2 dB from the guaranteed performance due to load impedance mismatch.

A large value bypass capacitor (1 uF or so) should be used in conjunction with the choke to present a low impedance path to ground for any signal that does manage to get past the choke. This capacitor should be attached between the supply side of the RF choke and ground.

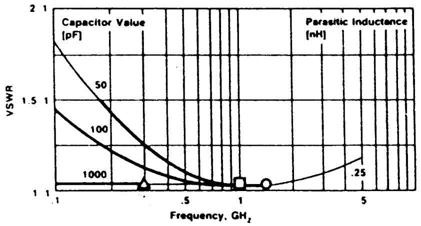


Figure 6a.

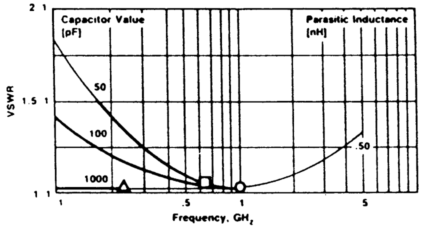


Figure 6b.

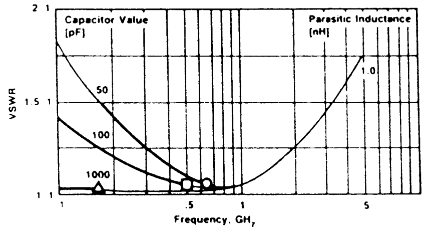


Figure 6c.

□	resonant frequency, .25 nH parasitic inductance
O	resonant frequency, 0.5 nH parasitic inductance
△	resonant frequency, 1.0 nH parasitic inductance

figure 6

Effects of DC blocking capacitors on VSWR as a function of frequency, capacitance and parasitic inductance.

paralleling MARs for higher power output

Since they are unconditionally stable (except for the MAR-8 discussed), MAR-amplifiers may be easily paralleled for increased output power (Fig. 11). Fortunately, the input and output impedances of paralleled amplifiers fall within the range that conveniently terminates standard 4:1, 9:1 and 16:1 broadband transformer configurations.

The bandwidth of the resulting multi-stage circuit will be limited by the bandwidths of the impedance matching elements. In Fig. 12, for example, the bandwidth would be limited by the 4:1 impedance transformers.

In applications not demanding excessive bandwidth, there are many appropriate impedance matching and combining techniques, such as quarter-wave transmission lines and Wilkinson n-way divider/combiners. Mini-Circuits' power splitter/combiners can be used for 50-ohm wide bandwidth applications

MAR-amplifiers may also be connected in push-pull (Fig. 12). The advantages of push-pull over straight paralleling are that stability and gain are retained, even-order harmonics tend to be cancelled and the push-pull circuit shown provides four times the power output of a single device.

In the circuit of Fig. 12, the input and output transformers are baluns (for BALanced to UNbalanced); a balun provides two signals at the balanced output which are 180 degrees out of phase, but equal in magnitude with respect to ground.

In the push-pull connection, even-order harmonic cancellation occurs because the output currents of even-order harmonics appear across the load resistor in phase with each other and, assuming that each is a perfect replica of the other, their sum is therefore zero at all times.

Note that the gain of the amplifier in the push-pull configuration is still the same as the gain of a single single-ended amplifier channel. Thus, to get four times the output power, the resulting push-pull amplifier must be driven with four times the input power.

A push-pull pair of MAR-amplifiers also lends itself to neutralization, and to the even more worthwhile design concept of true unilateralization. Unilateralization is a circuit technique in which the imaginary as well as the real term of the feedback elements are canceled. This creates an amplifier with a large degree of isolation between the input and the output.

At first glance, unilateralization might appear to be the same as neutralization as a means of stabilizing an amplifier. In neutralization, though, only the imaginary terms

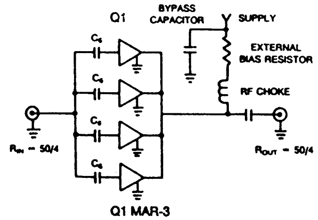


figure 11

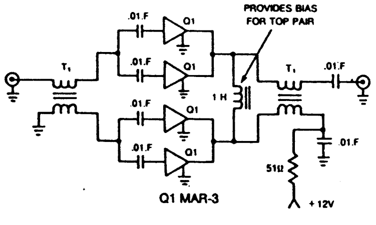


figure 12

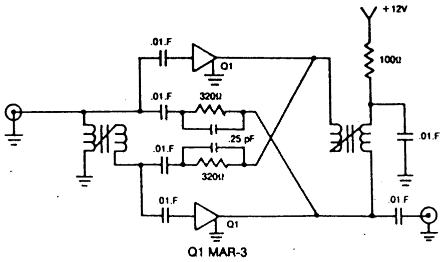


figure 13

cascading MAR-8s

The MAR-8 is a high-gain amplifier optimized for low noise figure with a typical power output of 10dBm up to 2GHz. But the amplifier is only conditionally stable, which means under certain conditions of temperature, current, load, and/or source impedance, the amplifier can oscillate. However, if the load and source impedance both have a VSWR of less than 3:1, at all frequencies below 2GHz the amplifier will be stable. Furthermore, as the source and load impedance show better VSWR (closer to 50-ohm impedance), stability will be further improved. With an interstage matching network, unconditionally stable two stage amplifier can be designed, which preserves noise and power output characteristics of the amplifier. Following are two examples which are optimized for low and high frequencies.

low frequency amplifier

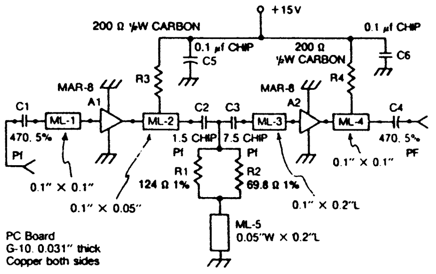


figure 9

Fig. 9 shows the schematic of an amplifier optimized for the frequency range of 100-500MHz, with a gain of 34 ± 1.5dB, and a noise figure of 3dB typical.

GAIN AND COMPRESSION **RETURN LOSS**

Frequency (MHz)	Gain (dB)	Comp (dB)	P_{out} (dBm)	RL_{IN} (dB)	RL_{OUT} (dB)
47 31	27 15	12	4 39	10 41	8 59
99 38	33 05	50	9 59	12 82	14 41
121 69	34 04	71	10 47	14 01	16 51
225 81	35 39	1 14	11 16	19 12	25 21
322 50	35 18	1 07	10 79	21 51	29 13
419 19	34 69	81	10 39	20 60	25 83
501 00	34 10	59	9 94	18 01	23 35
523 31	33 88	55	9 76	18 63	22 80
620 00	33 21	27	9 27	16 95	21 07
820 81	31 83	03	7 74	14 33	19 24
917 50	31 20	07	7 07	13 44	18 74
999 31	30 67	04	6 31	12 83	18 43
1014 19	30 59	04	6 15	12 72	18 34
1103 44	29 99	10	5 43	12 05	17 77
1118 31	30 01	13	5 50	12 05	17 88
1215 00	29 39	13	5 09	11 49	17 25
1311 69	28 84	10	4 39	10 66	17 77
1415 81	28 12	12	3 31	10 17	17 28
1512 50	27 75	07	3 09	9 49	17 48
1609 19	[illegible]	09	2 72	8 87	[illegible]
1713 31	26 50	[illegible]	1 70	8 09	21 06
1810 00	24 62	27	22	5 60	8 46
1899 25	23 40	46	1 06	7 05	8 81
[illegible]	23 44	38	1 16	7 06	8 27
2003 75	24 07	16	1 12	8 28	8 58

high frequency amplifier

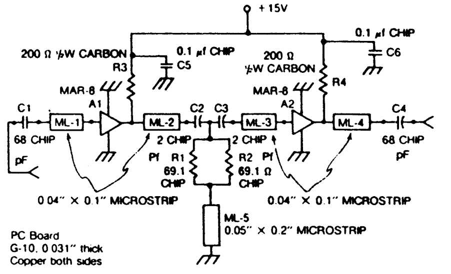


figure 10

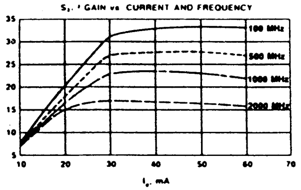
Fig. 10 shows the schematic of a two-stage amplifier optimized for a frequency range of 500-2000MHz, with a gain of 27 ± 2.5dB and, typically 3dB noise figure.

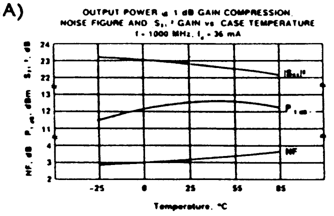
GAIN AND COMPRESSION **RETURN LOSS**

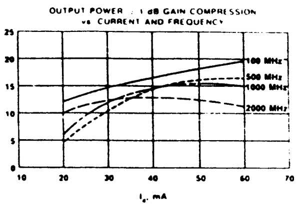
Frequency (MHz)	Gain (dB)	Comp (dB)	P_{out} (dBm)	RL_{IN} (dB)	RL_{OUT} (dB)
[illegible]	27 32	1 42	12 90	11 85	5 83
[illegible]	28 41	1 81	13 46	16 23	6 30
[illegible]	29 05	2 03	13 60	24 21	6 92
[illegible]	29 17	2 01	13 58	22 00	7 39
[illegible]	29 17	2 00	13 64	19 15	7 77
[illegible]	29 12	1 75	13 40	17 18	8 26
[illegible]	29 03	1 63	13 37	15 86	8 77
[illegible]	28 98	1 54	13 45	15 02	9 43
[illegible]	28 97	1 44	13 32	14 65	9 91
[illegible]	28 97	1 42	13 28	14 44	10 16
[illegible]	28 84	1 05	13 00	14 16	11 57
[illegible]	28 63	1 05	13 31	14 61	12 36
[illegible]	28 49	1 02	13 06	15 02	13 15
[illegible]	28 26	89	12 68	15 39	14 27
[illegible]	27 93	75	12 22	16 10	15 23
[illegible]	27 65	76	12 04	16 65	16 28
[illegible]	27 32	83	12 14	17 43	17 44
[illegible]	26 30	34	10 77	18 46	24 19
[illegible]	25 82	25	10 21	15 94	20 82
[illegible]	25 39	28	10 23	16 37	18 45
[illegible]	24 87	26	9 27	15 69	16 90
[illegible]	23 34	03	7 36	12 58	15 76
[illegible]	24 05	19	8 56	10 69	11 57
[illegible]	23 26	17	7 92	9 83	9 66
[illegible]	21 99	14	6 55	8 41	8 35

S-parameter data and performance curves

MAR-8 (T_A = 25°C, I_d = 36 mA)







Freq. MHz	S_{11} (Input Return Loss) dB	S_{11} Mag	S_{11} Ang	S_{21} (Power Gain) dB	S_{21} Ang	S_{12} (Isolation Out-In) dB	S_{12} Mag	S_{12} Ang	S_{22} (Output Return Loss) dB	S_{22} Mag	S_{22} Ang	K
100	−15.92	0.61	−21	33.0	162	−40.00	0.01	38	−4.73	0.58	−24	0.79
500	−8.18	0.39	−77	27.8	109	−27.96	0.04	52	−9.37	0.34	−96	0.75
1000	−11.37	0.27	−113	23.0	80	−24.44	0.06	51	−13.56	0.21	−147	0.89
1500	−11.70	0.26	−139	19.4	62	−21.94	0.08	46	−14.89	0.18	174	0.96
2000	−10.46	0.30	−155	16.9	47	−20.00	0.10	41	−15.39	0.17	153	0.97
2500	−9.63	0.33	−180	14.8	32	−18.42	0.12	32	−14.42	0.19	127	1.01
3000	−8.87	0.36	167	12.9	20	−17.72	0.13	27	−17.08	0.14	111	1.07
3500	−7.54	0.42	153	11.4	6	−17.08	0.14	21	−17.72	0.13	107	1.06
4000	−6.94	0.45	141	9.8	−5	−16.48	0.15	14	−19.17	0.11	106	1.10

single and three-stage layouts

A typical MAR-layout is shown in Fig. 7 using 1/32″ PTFE woven-glass board—a reasonable compromise between cost durability, and electrical performance. Note that the transmission lines have no bends and are tapered near the package to minimize step discontinuities. Twelve plated through holes, including two under the emitter leads, provide solid ground planes and minimal emitter parasitics for best high frequency performance. The gaps in the transmission line are appropriate for 50 mil ceramic chip capacitors, which have relatively low associated parasitic inductances—typically about 0.5 nH. Mini-Circuits offers a wide variety of values, see Table 1A. The DC pad arrangement requires that a bias stabilization resistor be used, but makes the use of an RF choke optional. If the choke is not used, the stabilization resistor would be connected between the output 50-ohm line and the V_{CC} supply line, and the bypass capacitor would be attached between the V_{CC} line and ground. Spacing is appropriate for 1/4 watt carbon resistors, molded inductors, and 1 uF electrolytic capacitors. The layout has been designed so that Fig. 8 can be repeated for multiple cascaded stages. Overall circuit dimensions are 1″ × 1.5″ for a single stage, with each additional stage adding one inch to the overall length. A three-stage cascaded design using chip resistors and inductors (R and L in diagram) is shown in Fig. 8.

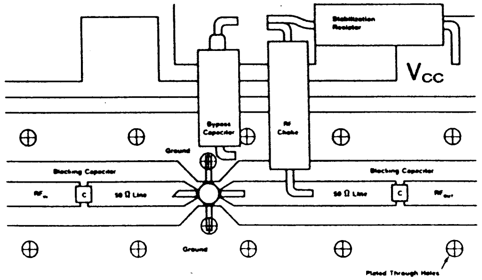


figure 7

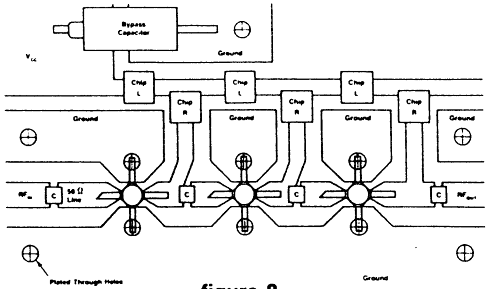


figure 8

of the feedback reactances are canceled because of the necessary inverse feedback is provided through an inductor (or capacitor), which does not track the reactance of the capacitive (or inductive) feedback over frequency. Consequently the conventionally-neutralized amplifier is stable only over a small frequency range. What is really needed is not an inductor or a capacitor, but a circuit element which is always equal in magnitude but opposite in sign to the positive feedback reactance of the device including all parasitic elements of the device itself, its package and the circuits in which it is installed.

Such a negative element can best be simulated with a duplicate active device. In the case of the push-pull configuration, it may be obtained by cross-coupling between the input of one of the two amplifier devices and the output of the other, and vice-versa. In Fig. 13, this condition is provided, with the input transformers serving the dual purpose of 4:1 impedance transformation and balun.

The reason that the MAR-amplifier is so easily unilateralized is that its internal feedback network is of very low Q compared to that of conventional amplifiers; in conventional amplifier devices, the feedback elements tend to be more reactive than resistive. Unilateralization of a push-pull pair of MAR-amplifiers would appear to negate some of the advantages of the basic amplifier itself. In general, this is true. Unilateralization is only useful in providing slightly higher gain or substantially more isolation.

After an amplifier has been unilateralized, the load impedance will no longer affect the input impedance and vice-versa, but unilateralization often increases the effective input and output impedance of the amplifier, the mechanism which actually increases the gain. Careful attention must be paid to the effects of unilateralization on the input and output match.

experimental results of paralleling

The paralleling techniques were tested in single-ended and push-pull experimental amplifiers. Two amplifiers were built with little effort toward optimizing board layout, and with standard components such as carbon composition resistors and chip capacitors. To optimize performance, microwave printed circuit layout techniques and microwave components would be preferable.

Table 6 shows the performance results while Figs. 14 and 15 show the single-ended and push-pull harmonic performance.

table 6

Configuration	Freq. (MHz)	Gain (dB)	P_{-1dB} (dBm)	2nd Harmonic @ P_{-1dB} (dB below carrier)
Single-ended	100	12	+10	−15
Push-pull (Unilateralized)	100	15	+13.5	−26
Push-pull	100	12	+17	−34

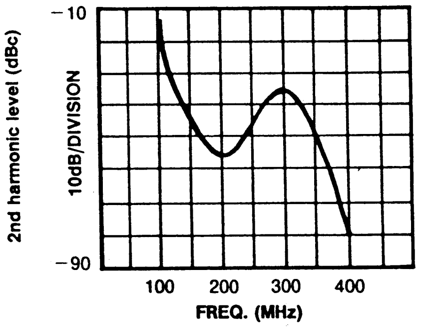


figure 14
Single-ended

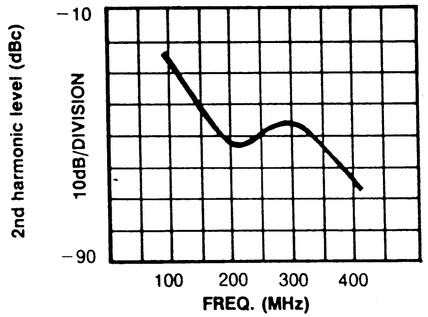


figure 15
Push-pull

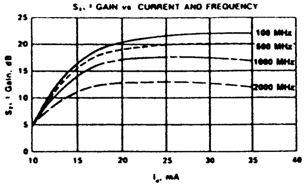
applications

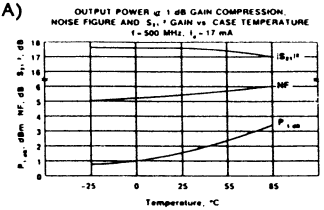
- low-power transmitter
- boost signal for improved detector efficiency
- multi-stage amplifier chain
- buffer amplifier for oscillators
- isolator

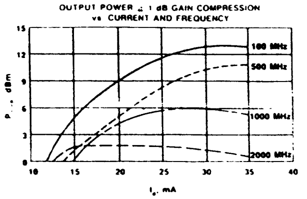
MAR amplifiers dc to 2GHz S-parameter data and performance curves

Mini-Circuits

MAR-1 (T_A = 25°C, I_d = 17 mA)

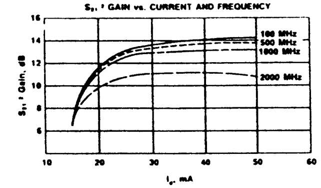

$|S_{21}|^2$ GAIN vs. CURRENT AND FREQUENCY

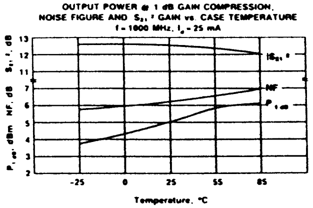

OUTPUT POWER @ 1 dB GAIN COMPRESSION, NOISE FIGURE AND $|S_{21}|^2$ GAIN vs. CASE TEMPERATURE f = 500 MHz, I_d = 17 mA

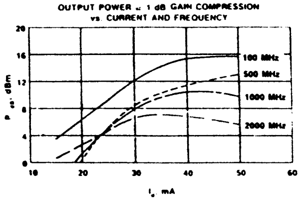

OUTPUT POWER @ 1 dB GAIN COMPRESSION vs. CURRENT AND FREQUENCY

Freq. MHz	S_{11} (Input Return Loss)			S_{21} (Power Gain)		S_{12} (Isolation Out-In)			S_{22} (Output Return Loss)		
	dB	Mag	Ang	dB	Ang	dB	Mag	Ang	dB	Mag	Ang
100	−23.10	0.07	164	18.5	171	−21.94	0.08	4	−23.10	0.07	−14
500	−24.44	0.06	106	17.5	141	−21.94	0.08	15	−23.10	0.07	68
1000	−24.44	0.06	72	15.5	111	−20.00	0.10	24	20.92	0.09	−124
1500	−27.96	0.04	59	13.7	87	−17.72	0.13	26	20.00	0.10	161
2000	−24.44	0.06	149	12.3	67	−15.92	0.16	21	−15.92	0.16	−177
2500	−20.00	0.10	142	10.6	49	−14.89	0.18	18	−15.92	0.16	155
3000	18.42	0.12	139	9.3	34	−13.98	0.20	12	16.48	0.15	144
3500	−13.98	0.20	129	7.9	19	13.15	0.22	5	−15.92	0.16	135
4000	−11.06	0.28	120	6.6	3	−12.04	0.25	−4	−15.39	0.17	127

MAR-2 (T_A = 25°C, I_d = 25 mA)

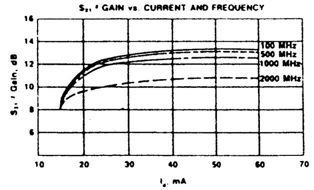

$|S_{21}|^2$ GAIN vs. CURRENT AND FREQUENCY

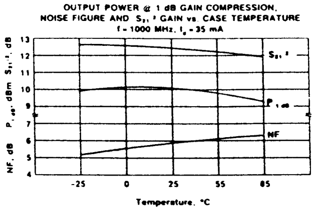

OUTPUT POWER @ 1 dB GAIN COMPRESSION, NOISE FIGURE AND $|S_{21}|^2$ GAIN vs. CASE TEMPERATURE f = 1000 MHz, I_d = 25 mA

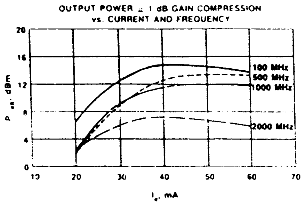

OUTPUT POWER @ 1 dB GAIN COMPRESSION vs. CURRENT AND FREQUENCY

Freq. MHz	S_{11} (Input Return Loss)			S_{21} (Power Gain)		S_{12} (Isolation Out-In)			S_{22} (Output Return Loss)		
	dB	Mag	Ang	dB	Ang	dB	Mag	Ang	dB	Mag	Ang
100	−18.42	0.12	173	13.0	174	−18.42	0.12	1	−17.72	0.13	−8
500	−19.17	0.11	154	12.8	156	−18.42	0.12	5	−18.42	0.12	38
1000	−20.00	0.10	130	12.5	131	−17.72	0.13	7	−18.42	0.12	−75
1500	−21.94	0.08	120	11.8	109	−17.08	0.14	10	−18.42	0.12	−112
2000	−24.44	0.06	126	11.0	90	−16.48	0.15	9	−17.72	0.13	−121
2500	−20.92	0.09	147	10.4	67	−15.39	0.17	6	−17.72	0.13	−165
3000	−17.08	0.14	147	9.4	46	−14.42	0.19	−1	−17.72	0.13	171
3500	−13.56	0.21	136	8.2	30	−13.98	0.20	−5	−17.72	0.13	155
4000	−11.37	0.27	123	7.3	14	−13.56	0.21	10	18.42	0.12	142

MAR-3 (T_A = 25°C, I_d = 35 mA)

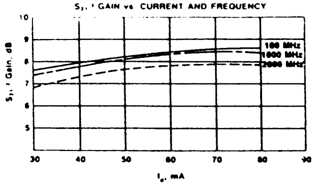

$|S_{21}|^2$ GAIN vs. CURRENT AND FREQUENCY

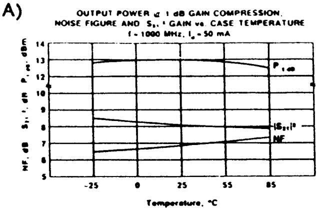

OUTPUT POWER @ 1 dB GAIN COMPRESSION, NOISE FIGURE AND $|S_{21}|^2$ GAIN vs. CASE TEMPERATURE f = 1000 MHz, I_d = 35 mA

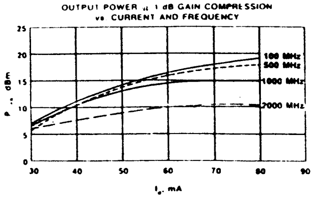

OUTPUT POWER @ 1 dB GAIN COMPRESSION vs. CURRENT AND FREQUENCY

Freq. MHz	S_{11} (Input Return Loss)			S_{21} (Power Gain)		S_{12} (Isolation Out-In)			S_{22} (Output Return Loss)		
	dB	Mag	Ang	dB	Ang	dB	Mag	Ang	dB	Mag	Ang
100	−23.10	0.07	172	13.0	174	−18.42	0.12	1	16.48	0.15	11
500	−24.44	0.06	156	12.8	152	18.42	0.12	5	15.92	0.16	45
1000	−26.02	0.05	146	12.5	128	−17.72	0.13	10	14.98	0.18	88
1500	27.96	0.04	172	11.8	103	−17.08	0.14	12	13.56	0.21	120
2000	−24.44	0.06	173	10.5	83	14.98	0.18	11	12.04	0.25	142
2500	15.39	0.17	175	10.3	59	−14.42	0.19	5	11.70	0.26	173
3000	12.40	0.24	157	9.1	38	13.98	0.20	0	12.04	0.25	168
3500	9.90	0.32	140	7.8	21	13.56	0.21	6	12.04	0.25	152
4000	8.18	0.39	124	6.5	3						

MAR-4 (T_A = 25°C, I_d = 50 mA)

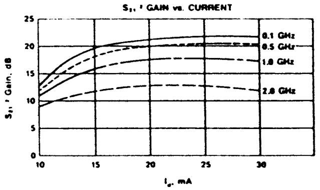

$|S_{21}|^2$ GAIN vs. CURRENT AND FREQUENCY

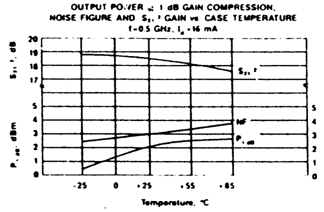

OUTPUT POWER @ 1 dB GAIN COMPRESSION, NOISE FIGURE AND $|S_{21}|^2$ GAIN vs. CASE TEMPERATURE f = 1000 MHz, I_d = 50 mA

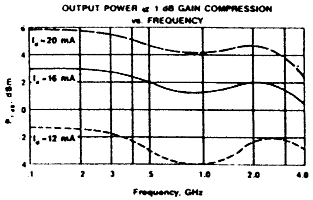

OUTPUT POWER @ 1 dB GAIN COMPRESSION vs. CURRENT AND FREQUENCY

Freq. MHz	S_{11} (Input Return Loss)			S_{21} (Power Gain)		S_{12} (Isolation Out-In)			S_{22} (Output Return Loss)		
	dB	Mag	Ang	dB	Ang	dB	Mag	Ang	dB	Mag	Ang
100	14.42	0.19	177	8.2	174	15.92	0.16	0	20.00	0.10	−14
500	14.89	0.18	169	8.2	156	15.92	0.16	1	17.72	0.13	−54
1000	15.39	0.17	159	8.1	135	−15.92	0.16	3	14.89	0.18	−94
1500	15.39	0.17	157	8.0	112	15.39	0.17	4	12.40	0.24	−121
2000	14.42	0.19	151	7.7	90	14.42	0.19	3	11.06	0.28	−145
2500	12.40	0.24	159	7.6	69	13.98	0.20	1	9.37	0.34	165
3000	10.17	0.31	151	6.9	46	13.15	0.22	6	9.12	0.35	176
3500	8.18	0.39	139	6.0	27	12.77	0.23	11	8.64	0.37	160
4000	6.74	0.46	126	4.9	8	12.04	0.25	23	8.40	0.38	147

MAR-6 (T_A = 25°C, I_d = 16 mA)

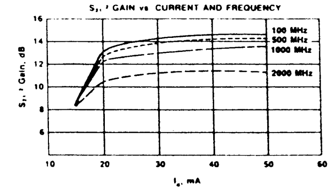

$|S_{21}|^2$ GAIN vs. CURRENT

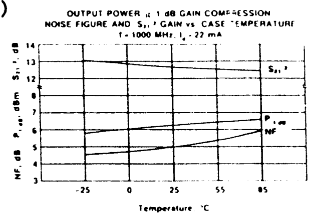

OUTPUT POWER @ 1 dB GAIN COMPRESSION, NOISE FIGURE AND $|S_{21}|^2$ GAIN vs. CASE TEMPERATURE f = 0.5 GHz, I_d = 16 mA

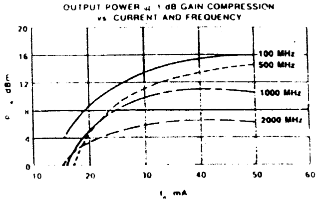

OUTPUT POWER @ 1 dB GAIN COMPRESSION vs. FREQUENCY

Freq. GHz	S_{11} (Input Return Loss)			S_{21} (Power Gain)			S_{12} (Isolation Out-In)			S_{22} (Output Return Loss)		
	dB	Mag	Ang	dB	Mag	Ang	dB	Mag	Ang	dB	Mag	Ang
0.1	27.96	.04	171	20.1	10.09	171	22.5	.075	5	27.96	.04	−30
0.5	26.02	.05	105	18.7	8.57	138	21.3	.086	21	20.00	.10	104
1.0	17.72	.13	118	16.4	6.57	107	18.8	.115	28	17.08	.14	150
1.5	13.56	.21	140	14.1	5.06	84	17.1	.140	28	16.48	.15	180
2.0	10.75	.29	163	12.0	3.98	65	15.8	.163	26	15.92	.16	157
2.5	9.37	.34	176	10.3	3.26	55	15.2	.174	28	15.92	.16	150
3.0	7.74	.41	169	8.7	2.71	42	14.8	.181	25	16.48	.15	143
3.5	6.74	.46	157	7.2	2.31	30	14.2	.194	22	17.72	.13	144
4.0	6.20	.49	146	6.1	2.01	18	13.8	.203	20	20.00	.10	156

MAR-7 (T_A = 25°C, I_d = 22 mA)


$|S_{21}|^2$ GAIN vs. CURRENT AND FREQUENCY


OUTPUT POWER @ 1 dB GAIN COMPRESSION, NOISE FIGURE AND $|S_{21}|^2$ GAIN vs. CASE TEMPERATURE f = 1000 MHz, I_d = 22 mA


OUTPUT POWER @ 1 dB GAIN COMPRESSION vs. CURRENT AND FREQUENCY

Freq. MHz	S_{11} (Input Return Loss)			S_{21} (Power Gain)		S_{12} (Isolation Out-In)			S_{22} (Output Return Loss)		
	dB	Mag	Ang	dB	Ang	dB	Mag	Ang	dB	Mag	Ang
100	26.02	0.05	169	13.5	173	19.17	0.11	1	17.08	0.14	7
500	30.46	0.03	133	13.1	150	18.42	0.12	6	17.72	0.13	41
1000	40.00	0.01	58	12.5	122	17.72	0.13	10	19.17	0.11	94
1500	24.44	0.06	113	11.8	95	16.48	0.15	10	17.72	0.13	148
2000	16.48	0.15	144	10.5	70	15.39	0.17	7	15.39	0.17	174
2500	11.37	0.27	165	9.6	48	14.89	0.18	1	14.42	0.19	154
3000	8.64	0.37	169	8.1	27	13.98	0.20	6	14.89	0.18	144
3500	6.94	0.45	150	6.5	10	14.42	0.19	11	15.39	0.17	144
4000	5.98	0.51	134	5.0	4	14.42	0.19	15	15.39	0.17	152

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