microstripline preamplifiers for 1296 MHz

Complete design and construction information for low noise, solid-state preamplifiers for the amateur 23cm band

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In a previous article I described a low-cost transceive converter for the amateur 1296-MHz band.[1] Calculated parameters and preliminary measurements indicated a receive system noise figure of 7.5 dB. As is often the case, these original claims have proved overly optimistic. Subsequent measurements with an argon-discharge noise source and an automatic noise-figure meter showed the true ssb noise figure to be more on the order of 9.5 to 10 dB. Neither improvements in the homebrew balanced mixer nor the use of high-grade commercial mixers significantly altered this figure.

It became apparent that optimum performance would necessitate the use of one or more low-noise preamplifier stages preceding the diode balanced mixer. Several popular amplifier circuits[2,3,4] were tried and the results were entirely satisfactory. However, I wanted to depart from the conventional technology (pi-network input and output with slab inductors) used in these designs. Since I had obtained considerable success with microstriplines in a family of transmit linear amplifiers, I decided to build a family of receive preamps using the same basic design techniques. The resulting circuits, presented here, represent a reasonable tradeoff between cost, performance, simplicity and reproducibility.

system considerations

It was assumed that these amplifiers would precede a receive converter with a ssb noise figure on the order of 10 dB. The Simple Sideband System[1] meets this criteria, as do most properly adjusted trough-line converters,[5] and at least one popular commercial unit.*

A workable rule of thumb is the principle that, if the gain preceding a receive converter is at least 10 dB greater than the converter's noise figure, then the noise figure of the total system will, for all intents and purposes, equal the noise figure of the preamplifier. Thus, 20 dB of preamplifier gain preceding a 10-dB NF converter will have the effect of masking the converter's noise.

A number of readily available microwave transistors are capable of 2- to 3-dB noise figures in the 1296-MHz

*Spectrum International MM_C 1296, $85.95 from Spectrum International, Box 1084, Concord, Massachusetts 01742.

amateur band. These devices offer conservatively rated power gains on the order of 10 dB per stage so if two such stages of preamplification precede the 10-dB NF converter, an overall noise figure of 2 to 3 dB will result.

When receive preamplifiers are connected in cascade, the first stage should be adjusted for optimum noise figure; subsequent stages are adjusted for optimum power gain. In practice the only difference lies in the input matching circuit. When a power-match is desired, the input circuit presents a complex conjugate match to the transistor's input impedance. In the case of a noise-match, a predetermined mismatch is introduced into the input circuit to minimize the stage's noise figure. In both cases the transistor's collector should look into a complex conjugate match.

microstripline considerations

The exact details of a microstripline design vary with the material used for the substrate. Such dielectrics as Teflon, Rexolite, Duroid, and glass offer superior performance at microwave frequencies. The material most readily available to the experimenter, however, is fiberglass-epoxy printed-circuit board. After having designed and built more than a dozen preamplifiers of the type described here, on a variety of substrates, I can state conclusively that the degradation in performance resulting from building on lowly glass-epoxy board is beyond the measurement capabilities of the average experimenter. Thus the amplifiers presented here were designed to be etched onto 1/16-inch (1.5mm) thick G-10 PC board, double-clad with 1-ounce copper.

It is possible to design an amplifier using microstriplines that requires no external tuning — all the resistive and reactive matching elements are provided by the microstriplines. Such designs have been published previously, and if properly built, will provide excellent performance without the need for tuning adjustments. However, amplifiers without tuning adjustments are practically an affront to the amateur spirit. Though considered frivolous by some, my amplifiers include trimmer capacitors. In addition to giving the dyed-in-the-wool experimenter something to tweak, these adjustable components provide some degree of compensation for slight variations between transistors, as well as variations from one PC board to the next.

preamplifier transistors

The second-stage amplifier is built around a Hewlett-Packard 35826E, a low-cost version of the well known HP-21 family. An acceptable substitute is the VO21 manufactured by the Nippon Electric Company, a second-source device which performs identically in the circuit. At the time of this writing, both transistors sell for $17.50 each in single quantities.*

The device used in the optimum-noise-matched first stage depends on the needs and budget of the individual builder. The design presented here uses a Hewlett-Packard 35866E option 100, one of the least expensive members of the low-noise HP-22 family. Priced at $45, it provides an overall system noise figure of 2 dB, challenging the best parametric amplifiers of yesteryear. Unless you anticipate EME or long-haul troposcatter communications, you will probably find it more cost-effective to use the lower priced HP-21 or VO21 in

*For the address of your nearest Hewlett-Packard distributor, look in the *Yellow Pages* or write to HPA Division, 540 Page Mill Road, Palo Alto, California 94304. Nippon Electric Company semiconductors are distributed in this country by California Eastern Labs, Inc., 1 Edwards Court, Burlingame, California 94010.

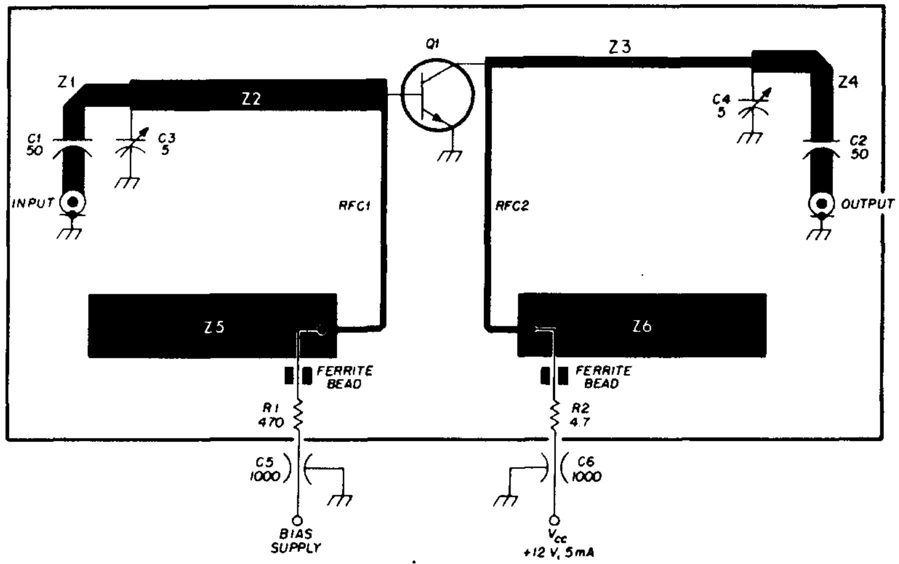
the first stage as well. These devices, when tuned for optimum noise figure in the circuit shown in **fig. 1**, will yield an overall system noise figure on the order of 3 dB.

The cost factor can be easily analyzed in terms of dollar expenditure per dB improvement in signal-to-noise ratio.

For example, two stages of HP-21 or VO21 at $17.50 per transistor will improve a 10-dB NF converter by 7 dB at a device cost of $5.00 per dB improvement. With the higher performance HP-22 in the front end, an additional dB of sensitivity can be achieved, but at an added cost of $27.50. The overall cost-effectiveness of the two-stage preamplifier now becomes $7.81 per dB – not an unreasonable figure when maximum performance is required.

Since it is common practice for microwave semiconductor manufacturers to produce a family of transistors by mounting the same chip in a variety of packages, it is often possible to substitute a different transistor of the same family without significantly influencing amplifier performance.

The part numbering system used at



C1,C2 50 pF chip capacitor (ATC 100 or equivalent

C3,C4 1 to 5 pF precision piston trimmer (Johannson JMC 4642)

C5,C6 1000 pF feedthrough capacitor

Q1 Hewlett-Packard 35866E, option 100 preferred (HP 35826E or NEC VO21 acceptable)

R1 470 ohm, ¼ watt carbon composition

R2 4.7 ohm, ¼ watt, carbon composition

RFC1 RFC2 100 ohm, quarter-wavelength microstripline, 0.02" (0.5mm) wide, 1.25 inch (32mm) long

Z1,Z4 50 ohm microstripline, 0.1" (2.5mm) wide, any convenient length

Z2 41.7 ohm, quarter-wavelength microstripline, 0.14" (3.5mm) wide, 1.18" (30mm) long

Z3 75 ohm, quarter-wavelength microstripline, 0.04" (1mm) wide, 1.22" (31mm) long

Z5,Z6 Rf short. 25 ohm, quarter-wavelength open-circuited microstripline, 0.30" (7.5mm) wide, (32mm) long

fig. 1. Circuit for the noise-matched 1296-MHz preamplifier stage. Printed-circuit layout is shown in fig. 2.

Hewlett-Packard provides some insight into the interchangeability of their devices. A system of five digits and a one-letter suffix is used, the first three digits being 358. The fourth digit position indicates the device family, with the number 6 designating HP-22 type devices (low noise microwave transistors), the number 2 referring to HP-21 type transistors (for general-purpose microwave applications), and the numbers 3 and 5 indicating the linear power transistors of the HP-11 and HP-12 families.

The fifth digit of the Hewlett-Packard part number indicates the type of package in which the semiconductor chip is mounted. Among the strip-packs, a number 2 indicates a 70-mil (1.8mm) diameter; 1, a 200-mil (5mm) diameter; and 6, a 100-mil (2.5mm) square package. The number 4 refers to a metal TO-72 package, and 7 is a coaxial package. Grounded-stud packages are designated by a 4, and 3 indicates a grounded-bar configuration. Unpackaged chips are coded zero. The letter suffix indicates whether the device is mounted in common-emitter (E) or common-base (B) configuration. Unpackaged chips carry the suffix A. The HP 35821E, for example, is a general-purpose (HP-21) microwave transistor mounted in a 200-mil (5mm) round strip package in the common-emitter configuration.

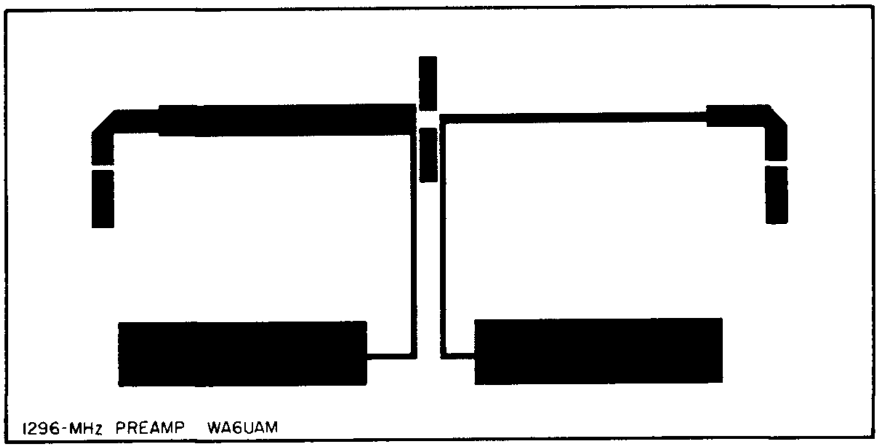


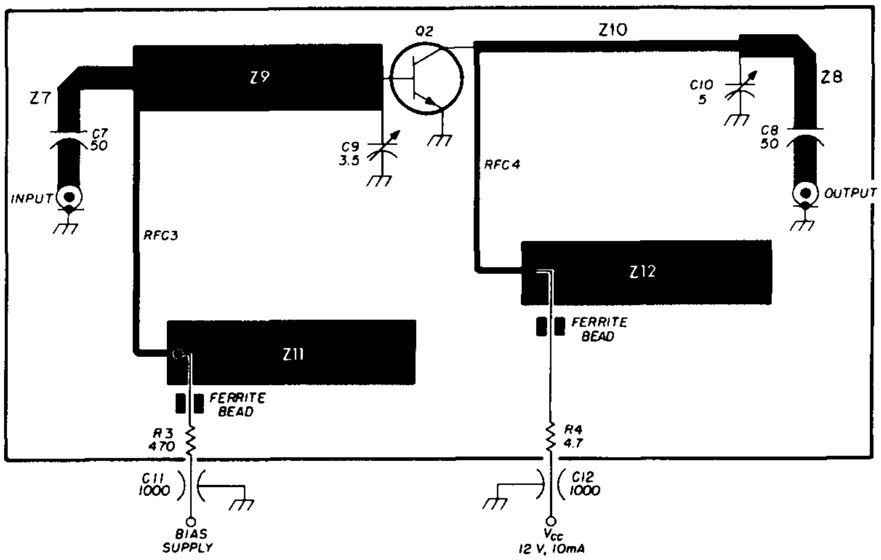
fig. 2. Full-size printed-circuit layout for the noise-matched 1296-MHz preamplifier shown in fig. 1.

Obviously, differently packaged versions of the same semiconductor chip will differ from one to another in terms of their complex input and output impedances. However, at low frequencies (and for microwave devices 1.3 GHz can be considered low), the contribution of package parasitics to the overall input (S_{11}) and output (S_{22}) S-parameter values can be considered small in relation to the characteristics of the chip itself.* These amplifiers, though designed around the characteristics of the 100-mil (2.5mm) square package, function well with like transistors in the 70-mil (1.8mm) and 200-mil

*The S-parameters or scattering parameters are one of several methods used for describing the input-output characteristics of transistors. The basic difference between S-parameters and other systems are that the S-parameters are measured with transmission-line inputs and outputs; this greatly simplifies measurements at uhf and microwave.

(5mm) round packages with only minor degradation in input and output vswr.

An analogous situation exists with Nippon Electric Company transistors, although NEC lacks the elaborate numbering system which identifies the HP with this newer transistor in the existing boards, performance was wholly satisfactory. As a matter of fact, the VO21 in the 100-mil (2.5mm) square package exhibited 0.2 dB lower noise figure than the original device.



C7,C8	50 pF chip capacitor (ATC 100 or equivalent)
C9	0.35 to 3.5 pF precision piston trimmer (Johannson JMC 5801)
C10	1 to 5 pF precision piston trimmer (Johannson JMC 4642)
C11, C12	1000 pF feedthrough capacitor
Q2	Hewlett-Packard 35826E or NEC VO21
R3	470 ohm, ¼ watt carbon composition
R4	4.7 ohm, ¼ watt carbon composition
RFC3 RFC4	100 ohm, quarter-wavelength microstripline, 0.02" (0.5mm) wide, 1.25" (32mm) long
Z7,Z8	50 ohm microstripline, 0.10" (2.5mm) wide, any convenient length
Z9	25 ohm, quarter-wavelength microstripline, 0.30" (7.5mm) wide, 1.14" (29mm) long
Z10	75 ohm, quarter-wavelength microstripline, 0.04" (1mm) wide, 1.23" (31mm) long
Z11, Z12	Rf short. 25 ohm, quarter-wavelength open-circuited microstripline, 0.30" (7.5mm) wide, 1.14" (29mm) long

fig. 3. Circuit for gain-matched 1296-MHz amplifier stage. Printed-circuit layout for this amplifier is shown in fig. 4.

devices. The NEC VO21 was initially manufactured in a 150-mil (3.8mm) round package, and it was the scattering parameters for that device which were used in the design of these amplifier boards. Recently NEC introduced a VO21 chip mounted in a 100-mil (2.5mm) square package. Although perfect matching could not be achieved

When ordering the NEC VO21 transistor, specify the 320 package if a 150-mil (3.8mm) round package is desired. For the 100-mil (2.5mm) square, request package ML-3. There is no price difference.

construction

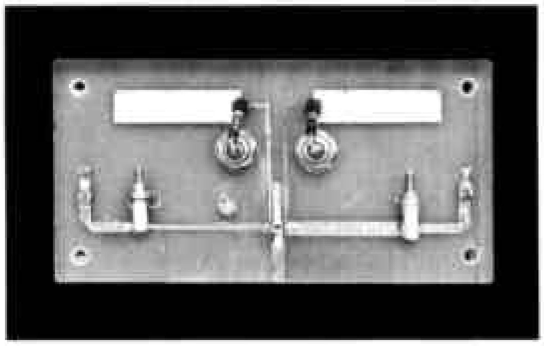
Figs. 1 and **3** are functional schematics of the amplifier stages. Microstriplines opposite a groundplane (the unetched side of the double-clad printed-circuit board) comprise all matching transformers, rf chokes and rf bypasses. **Figs. 2** and **4** are full-size printed-circuit layouts for the noise-matched and power-matched stages, respectively. Note that dimensions are applicable only to 1/16-inch (1.5mm) G-10 double-sided glass-epoxy printed-circuit board.

Mounting of the transistors is detailed in **fig. 5**. A hole is drilled in the PC board to allow direct strapping of the emitter traces to the groundplane. It is desirable to cover this hole on the groundplane side after the transistor is installed, to improve shielding and furnish physical protection for the transistor.

A word of caution is in order regarding mounting the transistors to the boards. There is a difference in lead layout between the HP and NEC transistors which can cause considerable confusion. The emitter leads of both devices are the wider pair of leads. One of the two remaining leads is tapered 45° at the end. The slashed lead is the *base* of the NEC devices, but the *collector* of the HP transistors. This small difference cost me two expensive transistor failures while developing the prototype amplifiers for this article.

The grounded (adjusting screw) end of the trimmer capacitors must be connected directly through the board to the groundplane to minimize tuning-tool interactions when making adjustments. Although concentric-ring piston trimmers (Johannson, JFD, etcetera) are ideal for this purpose, performance of the lower cost ceramic piston trimmers used in uhf TV tuners is entirely satisfactory.

The dc blocks at the input (C1 and C7) and output (C2 and C8) should ideally be ceramic chip capacitors. If these are not available, modified miniature disc ceramic capacitors are usable. An Xacto knife is used to scrape the insulation off of the sides of the capacitor, the leads are removed and, using the lowest possible soldering heat, the PC traces are bridge-soldered directly to the capacitor plates.



Microstripline side of the noise-matched 1296-MHz preamplifier. The low-noise H-P 35866E transistor is at the top center of the board. The gain-matched stage is similar except for the width of the input quarter-wave transformer and placement of the input trimmer capacitor.

The prototype preamplifiers shown in the photographs use miniature SMA coaxial connectors, but the less expensive JCM connectors made by E. F. Johnson are satisfactory substitutes. TNC connectors may be used, too, but they are much larger. BNC connectors should be avoided as they will slightly degrade the noise-figure performance of the amplifier stages. **Fig. 6** shows a method of modifying flange-type bulkhead connectors for use as microstripline launchers.

All power-supply leads for the collector supply and bias current are isolated from the rf circuitry by rf chokes, ferrite beads and bypass and feedthrough capacitors. Therefore, the necessary bias circuitry can be installed on the groundplane side of the circuit board.

bias circuits

With the transistors specified, optimum noise figure occurs at a collector current of approximately 5 mA. A higher collector current will improve stage gain. In the conventional configuration of a low-noise first stage and higher-gain second stage, it is desirable to bias the two stages for collector currents of 5 and 10 mA, respectively.

Virtually any bias circuit which will maintain the desired collector current is acceptable. Many of the simpler resistive bias circuits should be avoided due to their low stability factor (that is, high dependence of collector current on transistor dc current gain) and the resulting danger of thermal runaway. To quote a useful Hewlett-Packard applications note, "Often the least considered factor in microwave transistor circuit design is the bias network. Considerable effort is spent in measuring S-parameters, calculating gain, and optimizing bandwidth and noise figure, while the same resistor topology is used to bias the transistor. Since the cost per dB of microwave gain or noise figure is so high, the circuit designer cannot afford to sacrifice rf performance by inattention to dc bias considerations."[6]

An active bias circuit (variable constant-current source) is desirable in

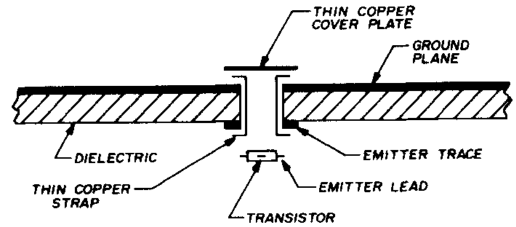


fig. 5. Method for mounting the transistors to the printed-circuit boards. Note thin copper strap used to connect the emitter circuit trace to the ground plane. Solder emitter leads as close as possible to the transistor package and use minimum soldering heat to avoid transistor damage.

that it affords a degree of protection for the transistor while permitting ready collector current adjustment for optimizing stage gain and/or noise figure. One such circuit, shown in **fig. 7**, furnishes a variable collector current of 2 to 12 mA, more or less independent of the dc current gain of the transistor being biased.

For initial amplifier tuneup, adjust the trimpot in the base of the bias transistor to produce 5.5 mA of total current in the first amplifier stage, and 10.5 mA in the second. (This accounts for a quiescent bias circuit current of 0.5 mA). Upon completion of all rf tuning, the trimpots may be adjusted to optimize the overall system noise figure.

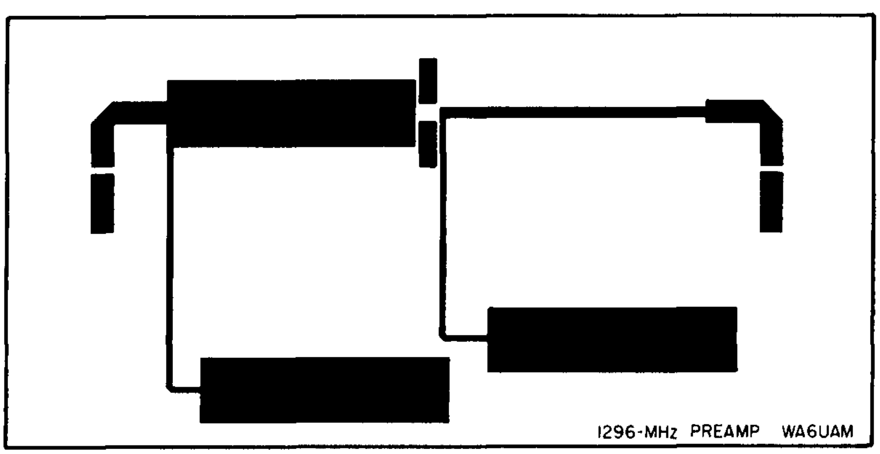


fig. 4. Full-size printed-circuit layout for the 1296-MHz gain-matched amplifier shown in fig. 3.

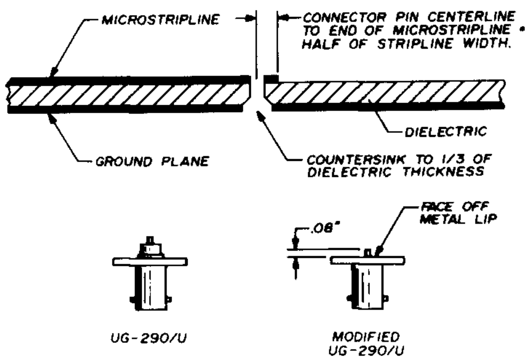


fig. 6. Modifying coaxial bulkhead connectors for use as microstripline launchers.

alignment

The two preamplifier stages are built as separate subassemblies and connected together with a short length of coaxial cable. The reasons for this modular approach are twofold. Most obvious is the fact that many operators may wish to add a single stage of preamplification initially, expanding their system later as needs and budget dictate. In this case it is recommended that the power-gain matched stage (which uses the less expensive transistor) be built first.

Even if two stages of preamplification are built, tuneup and matching can be most readily accomplished if the stages are in separate modules. The power-matched stage is connected to the receiving converter first, its input terminated in 50 ohms, and the trimmer capacitors adjusted for maximum received signal from a beacon source.[7,8,9] Once a power match is achieved in the second stage (both input and output), the first stage is connected and, using a weak-signal source, its *output circuit only* is adjusted for maximum received signal. A power match now exists between the two amplifiers as the output of the first stage and the input to the second stage are each matched to 50 ohms, and a 50-ohm coaxial cable connects the two.

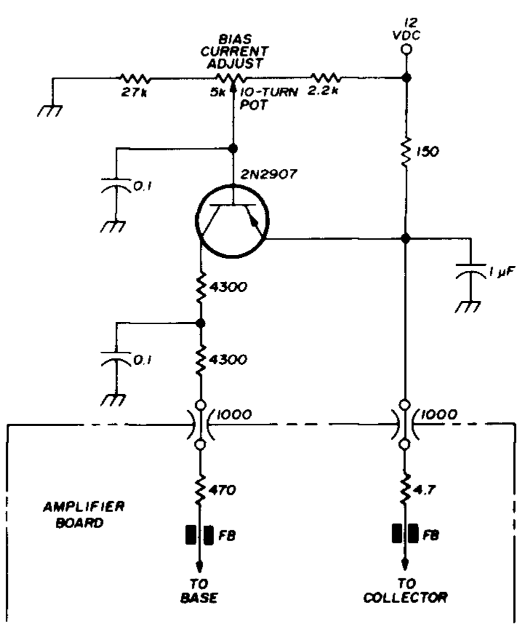


fig. 7. Active bias circuit provides adjustable collector currents from 2 to 12 mA. Components below dotted line are part of the amplifier circuit (see figs. 1 and 3).

The tuning process is completed by obtaining a proper noise match into the first stage. This is most readily accomplished by using an argon-discharge type noise source.[10,11] Unfortunately, few experimenters have access to such equipment, except perhaps at regional uhf conferences.* A semiconductor diode noise source is an acceptable alternative.[12,13] A number of articles have described the process of tuning an amplifier for minimum noise figure.[14-18]

When adjusting the input circuit of the first stage, an important consideration is the interactive nature of the input matching and bias current adjustments. Since the transistor's optimum source reflection coefficient, Γ_o (see page 24), varies with collector current,

*One of the more popular attractions at the annual West Coast VHF Conference is a receiver noise-figure competition, during which participants may optimize the performance of their receiving equipment with an automatic noise measuring system. A gas-discharge noise source is usually available for noise-figure measurements on both 1296 and 2304 MHz.

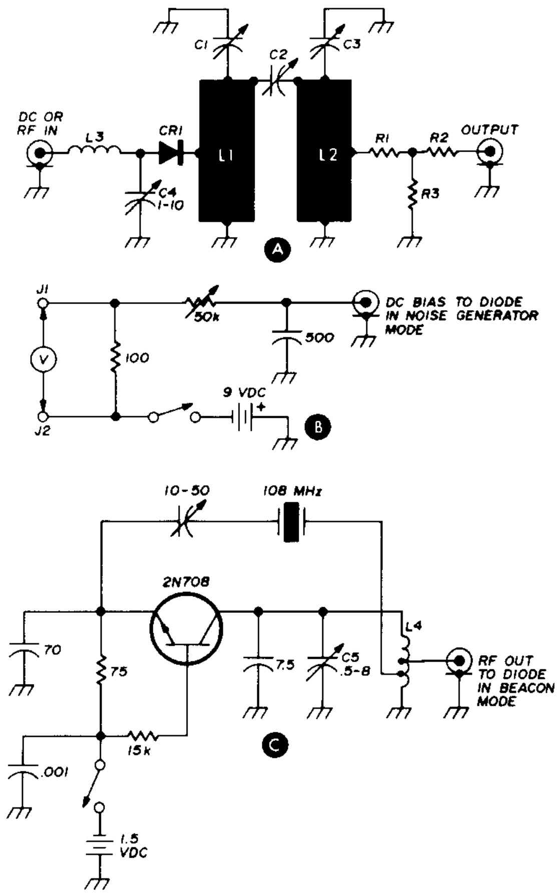
the ultimate in performance is achieved by alternate adjustments to first the input trimmer capacitor, then the bias pot. These adjustments are repeated until no further improvement in noise figure can be achieved.

The tuning elements of these amplifiers are inherently broadband. Without adequate selectivity at the input, bipolar transistors can exhibit disastrous overload characteristics. The cross-modulation and intermodulation effects of operating these preamplifiers in a high rf-density environment (i.e., virtually any populated area of the world today) can completely nullify any system noise-figure improvement. As a precaution it is good practice to provide input selectivity external to the preamplifier in the form of a high-Q filter in series with the antenna input. Single-pole coaxial or trough-line resonators, as well as some multi-pole microstripline filters, can provide adequate selectivity against out-of-band signals with a minimum of insertion loss (which would add to the amplifiers's noise figure in establishing receiver sensitivity).

When an input filter is used it is important to first optimize the preamplifier's noise figure as discussed above, then adjust the filter for minimum insertion loss at the operating frequency in a 50-ohm system. After mating the two, no further adjustments should be made to either the filter or the preamplifier unless a precision automatic noise-figure meter is available.

multipurpose uhf tuning instrument

It is interesting to note the similarities between the diode noise sources and rf beacons commonly used for adjusting uhf receiving preamplifiers. Both instruments normally have an output circuit consisting of a microwave diode feeding a tuned circuit, with an output port matched to 50 ohms. The two pieces of test equipment differ in that the diode is fed with direct current



C1,C2	0.5 to 2 pF piston trimmer capacitor
C3	(low-cost ceramic type acceptable)
CR1	microwave diode (1N25 or equivalent)
L1,L2	grounded microstripline, 0.3" (7.5mm) wide, 0.9" (23mm) long, tapped 0.2" (5mm) up from grounded end
L3	6 turns no. 20, 0.1" (2.5mm) diameter, 0.5" (13mm) long
L4	5 turns no. 20, 0.25" (6.5mm) diameter, 0.5" (13mm) long, tapped at 1 and 2½ turns from grounded end
R1,R2	10 ohm, ¼ watt carbon composition
R3	110 ohm, ¼ watt carbon composition

fig. 8. Combination weak-signal source (C) and diode noise generator (B). Microstriplines L1 and L2 are etched on 1/16" (1.5mm) double-clad, fiberglass-epoxy circuit board (A).

when used as a noise generator, versus rf when used as a weak signal source. The diode functions as a white-noise generator in the former application as opposed to harmonic generation in the latter.

There is no reason why the two

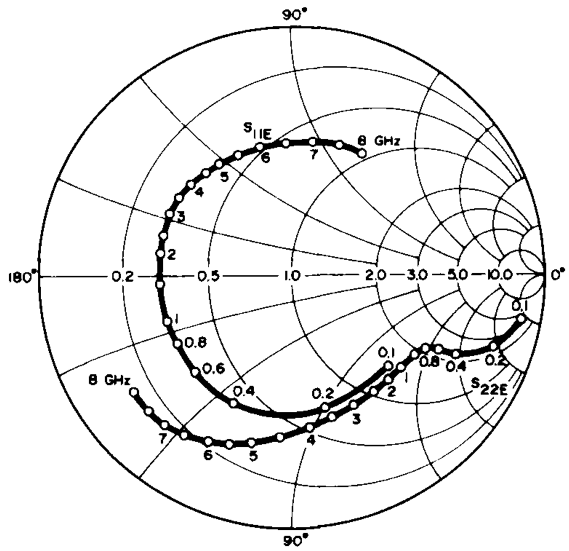


fig. 9. Input reflection coefficient, S_{11}, and output reflection coefficient, S_{22}, vs frequency for the Hewlett-Packard 35866E transistor (V_{CE} = 10 volts, I_C = 10 mA).

functions cannot be combined into a single instrument. **Fig. 8** represents an attempt to do so. The circuitry associated with the diode and its two-pole resonator is similar to the scheme I used in the local-oscillator section of the 1296-MHz transceive converter with filtering consisting of microstripline inductors etched on G-10 glass epoxy PC board. The three resistors at the output port form a 3-dB T-attenuator which assures a reasonable approximation of a 50-ohm output impedance. (In the weak-signal-source mode, additional pads may be placed between the output of the generator and the input to the receiver.)

The diode's input circuit is an adaptation of the familiar L-match used in the multiplier circuits of trough-line converters. The crystal-controlled 108-MHz oscillator which feeds the diode in the signal-generator mode is borrowed from a popular signal source design[7] while an existing noise-generator circuit[17] provided a workable diode biasing scheme. In short, the design presented in **fig. 8** is an amalgamation of various pieces of uhf test equipment into a single package.

Tuneup of the multipurpose instrument consists of merely connecting it in the rf beacon mode and tuning capacitors C1, C2, C3, C4 and C5 for maximum 1296-MHz signal into your receiver.

circuit design

For those readers who care to follow the calculations involved, the remainder of this article documents the procedure used to design the microstripline matching circuits of these 1296-MHz amplifiers. It should be pointed out that there are at least as many different methods for designing microstripline amplifiers as there are microwave engineers, and no one technique is necessarily any better or more workable than the others. The method shown here represents nothing more sacred than my own personal preference. It should be pointed out, however, that many of the more elegant amplifier designs used in the microwave industry are so complex as to be solvable only with the aid of a large digital computer. The designs shown here, though somewhat less precise, can be calculated by the average experimenter using only a slide rule.

Let us first consider a method for obtaining a complex conjugate impedance match to the amplifier transistors. Toward the end of this article I will discuss the special case of precisely mismatching the input to the first stage to obtain optimum noise figure.

Nearly all microwave semiconductor manufacturers publish Smith charts depicting the complex input and output impedances (S_{11} and S_{22}, respectively) as a function of frequency. **Figs. 9, 10** and **11** show such data for the HP 35866E, HP 35826E and NEC VO21, respectively. In addition to these charts, most manufacturers furnish tabulated data listing the input and output impedances at various frequencies and differing bias conditions. It is important to note that these impedances vary significantly with changes in the dc operation of the transistor.

When tabular data is furnished, complex impedances are generally shown in polar form, i.e., magnitude and angle. This polar form may be converted to the more familiar rectangular notation ($A \pm jB$) on a Smith chart as indicated in **fig. 12**. Note that the magnitude is a decimal indication of the distance along a radius of the Smith chart, from zero (center) to 1 (circumference). The angle listed represents the direction of that radius. (For further material on the use of Smith charts see reference 19.)

Table 1 lists complex impedances, in both polar and rectangular form, for the transistors used in these amplifiers under the applicable dc bias conditions, at 1296 MHz. It is useful to convert the complex series impedances to their shunt equivalent circuit values. The applicable formulas are

$$R_p = R_s + \frac{X_s^2}{R_s}$$

$$X_p = \frac{R_s \times R_p}{X_s}$$

where R_s and X_s represent the resistive and reactive components, respectively, of the complex series impedance, and R_p and X_p represent the components of the parallel equivalent circuit. The parallel equivalents are included in **table 1**.

table 1. Typical S-parameters at 1.3 GHz (given in polar form, rectangular form and as parallel shunt equivalent, respectively).

	HP 35826E $I_c = 5$ mA	HP 35826E $I_c = 10$ mA	HP 35866E $I_c = 5$ mA	NEC V021 $I_c = 10$ mA
Input reflection coefficient, S_{11}	—	$0.61 \angle 178°$ $12.5 + j0.5$ $12.77 \parallel +j325.6$	—	$0.626 \angle 171°$ $11.75 + j4.0$ $13.11 \parallel -j38.5$
Output reflection coefficient, S_{22}	$0.57 \angle -41°$ $75 - j80.0$ $160 \parallel -j150$	$-0.51 \angle -40°$ $76 - j67.5$ $136 \parallel -j153$	$0.61 \angle -37°$ $80 - j90.0$ $181 \parallel -j161$	$0.266 \angle -65°$ $55 - j27.5$ $69 \parallel -j138$

output matching

It was desired to develop a single output circuit which would approximate a match to 50 ohms for any of the three referenced transistors, under either of the two indicated bias conditions. Assume for a moment that the reactive component of each transistor's parallel-equivalent complex output impedance is cancelled by a reactance of like magnitude and opposite sign shunting the collector (this will be accomplished shortly). Under these conditions the impedance to be matched to 50 ohms is a real value of magnitude R_p. For the conditions being considered, the value of R_p varies between 68.75 and 181.0 ohms. A compromise output circuit should match to the geometric mean of these outside values

$$R_p\ (\text{mean}) = \sqrt{R_p\ (\text{max}) \times R_p\ (\text{min})} = 111.6 \text{ ohms}$$

A 111.6-ohm nonreactive source may be matched to a 50-ohm nonreactive load through a quarter-wave matching transformer of characteristic impedance

$$Z_o = \sqrt{Z_{in} \times Z_{out}} = \sqrt{111.6 \times 50} = 74.7 \text{ ohms}$$

Still assuming no reactive component, the actual amplifier output impedance resulting from the use of each transistor

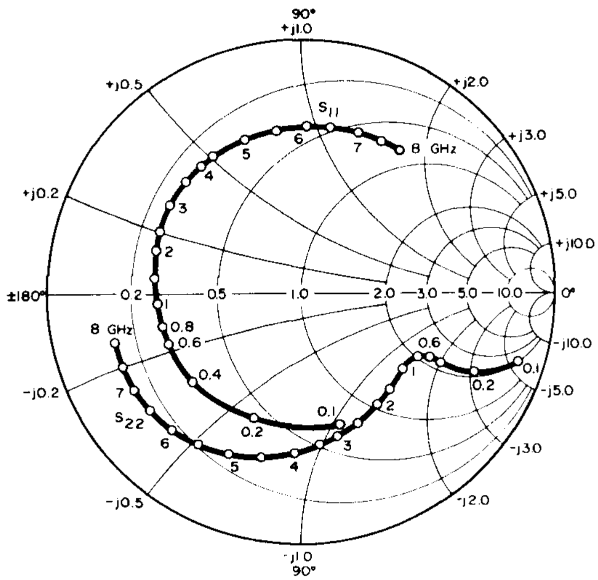


fig. 10. Typical input reflection coefficient, S_{11}, and output reflection coefficient, S_{22}, vs frequency for the H-P 35826E transistor (V_{CE} = 15 volts, I_C = 15 mA).

in this circuit would be that transistor's equivalent parallel output resistance transformed through a 75-ohm quarter-wave section. These values, along with

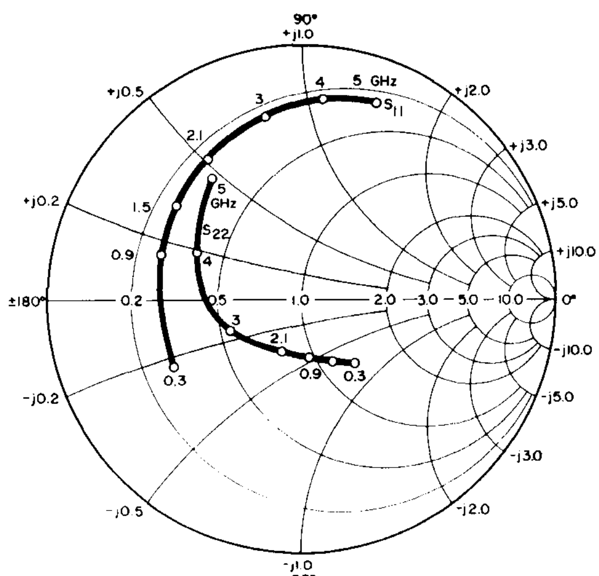


fig. 11. Typical input reflection coefficient, S_{11}, and output reflection coefficient, S_{22}, vs frequency for the NEC VO21 transistor (V_{CE} = 10 volts, I_C = 10 mA).

the resulting output vswr, are listed in **table 2**. Note that, even though restricted to a single output circuit for various combinations of device and collector current, this results in an acceptably low vswr well below 2:1.

Now, what about the reactive component of the transistor's output impedance so blithely ignored up to this point? **Table 1** reveals the various values of parallel equivalent reactance, X_p, to be a shunt capacitive reactance (-j) in all cases, varying between 137.5 and 161.0 ohms. Obviously these capacitive reactances could be cancelled out by a variable inductor of like reactance range connected in shunt with the transistor's collector. However, from a practical standpoint, a variable capacitor is a more desirable tuning element than a variable inductor.

table 2. Actual output impedance and vswr of the Hewlett-Packard and NEC transistors in the circuits of figs. 1 and 3.

device	I_c	$Z_{out}=75^2/R_p$	vswr
HP 35826E	5 mA	35.16 ohms	1.42:1
HP 35826E	10 mA	41.36 ohms	1.21:1
HP 35866E	5 mA	31.08 ohms	1.61:1
NEC V021	10 mA	81.82 ohms	1.64:1

The desired inductance may be realized by connecting a shunt capacitance to the collector through a quarter-wave transformer. As luck would have it, a quarter-wave transformer already exists at the collector circuit — the 75-ohm section used to match the transistor's parallel equivalent resistance to 50 ohms!

What value of capacitive reactance must be connected to the load end of the 75-ohm quarter-wave transformer? It must match the inductive reactance, X_{out}, resulting from transforming the transistor's shunt capacitive component, X_p, through a 75-ohm quarter-wave section. The relationship is

$$X_{out} = -\frac{Z_o^2}{X_{in}}$$

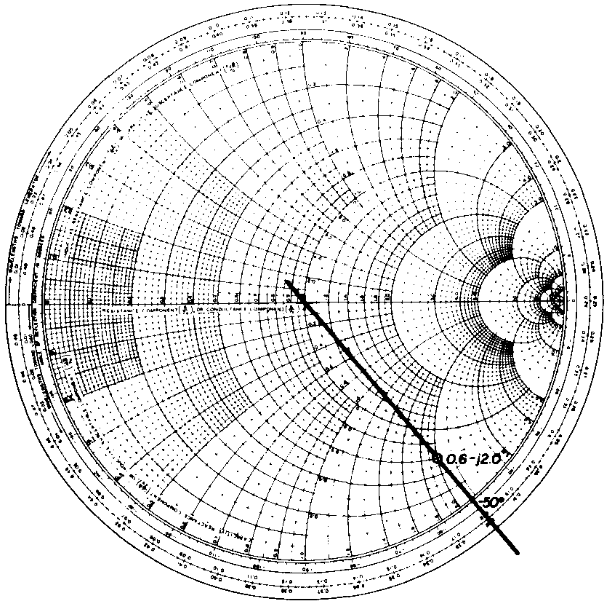


fig. 12. Using the Smith chart to convert impedance in the polar form to rectangular form (A ±jB). The magnitude and angle 0.8 $\angle -50°$ lies on a radius passing through the −50° point on the outer circumference, 80% of the linear distance from the center to the edge of the chart, at 0.6 − j2.0 on the normalized Smith chart shown here (30 −j100 ohms in a 50-ohm system).

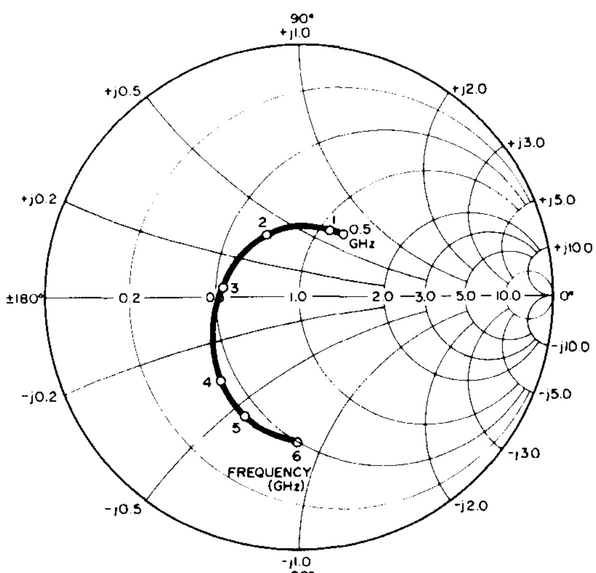


fig. 13. Typical optimum source reflection coefficient, Γ_o vs frequency for the Hewlett-Packard 35866E transistor (V_{CE} = 10 volts, I_C = 5 mA).

figure will result. **Fig. 13** shows that, for an HP 35866E option 100 device operated with I_C = 5 mA, Γ_o at 1.3 GHz is equal to 40 + j25 ohms. This translates to a shunt equivalent circuit of 34.8 ohms in parallel with + j55.6 ohms. Disregarding the inductive reactance shunting the base for a moment, the required real component can be readily realized by transforming the 50-ohm input impedance through a quarter-wave section with a characteristic impedance

$$Zo = \sqrt{50 \times 34.8} = 41.7 \text{ ohms}$$

As before, the desired inductive reactance shunting the transistor is achieved by adding a shunt capacitive reactance a quarter wavelength away from the transistor. The required capacitive reactance is

$$X_C = \frac{Z_o^2}{X\Gamma_o} = 31.3 \text{ ohms}$$

which at 1.3 GHz would represent a capacitance of 3.9 pF. A 1 to 5 pF trimmer is used for tuning the input to the first stage for optimum noise figure as shown in **fig. 1**.

modified interstage circuit

A number of the active 1296 operators who reviewed preliminary copies of this manuscript expressed an interest in a modified interstage design which would allow both stages of preamplification to be combined on a single amplifier board. Although I personally prefer separate modules, I concede that such a design would be of some value. A two-stage 1296-MHz preamplifier design by W6KQG[20] transformed the output impedance of one HP-21 into the complex conjugate of the second HP-21's input impedance through a single quarter-wave microstripline. A similar approach for these amplifiers, with the added provision of reactive tuning, is shown in **fig. 14**.

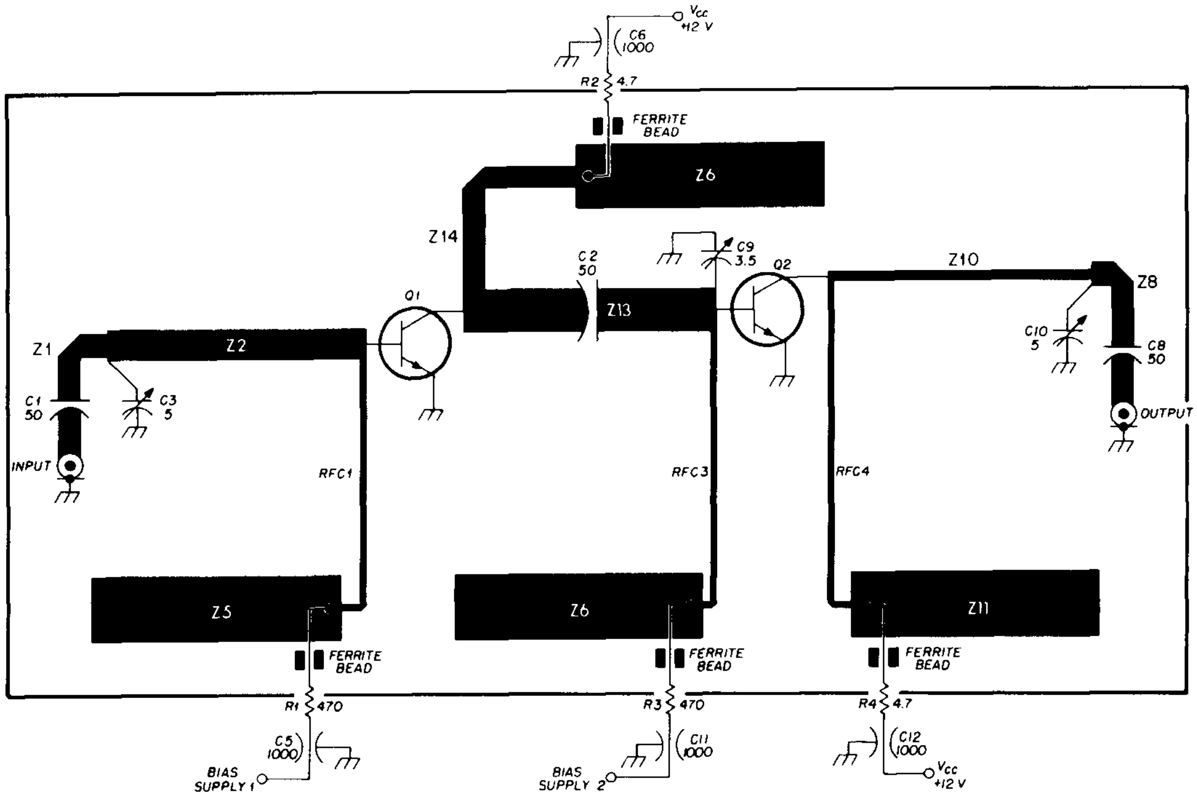


fig. 14. Modified interstage circuit for building the two 1296-MHz amplifier stages in figs. 1 and 3 on one circuit board. Circuit-board layout is shown in fig. 15.

It should be pointed out that, with two stages cascaded in this manner, it is impossible to measure the output vswr of the first stage or the input vswr of the second. Therefore, proper matching can be approximated only by tuning the amplifier for maximum gain. This is complicated by the fact that interstage matching is influenced by the output reflection coefficient, S_{22}, of transistor Q1 (which varies with I_{C1}), the input reflection coefficient, S_{11}, of Q2 (which varies with I_{C2}), and the complex impedance of the interstage network (which is controlled by capacitor C9). In practice C9 should be adjusted for maximum amplifier gain with I_{C1} set at 5 mA and I_{C2} at 10 mA. Further adjustments should then be made while monitoring total system noise figure with an automatic noise meter.

A full-size printed-circuit layout for the unified two-stage amplifier board is provided in **fig. 15**. Preliminary tests show the total gain of this amplifier to be within 0.5 dB of the two-module arrangement. With HP-22 transistors at Q1, both amplifier configurations yielded noise figures of 2.3 dB, measured on a Hewlett-Packard 340B Automatic Noise Meter with an AIL 7010 argon-discharge noise source.

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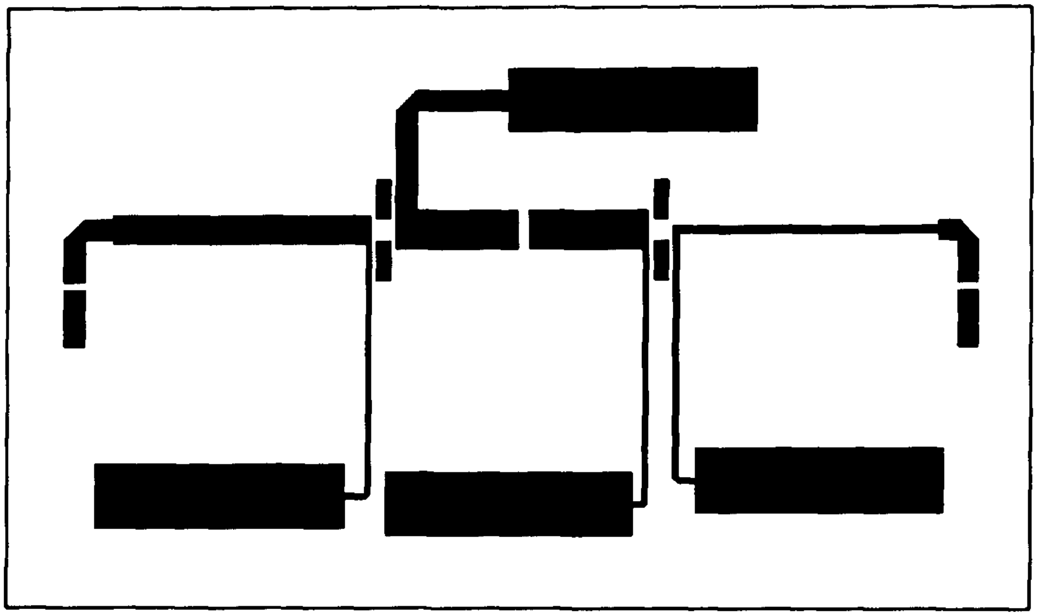

Z13 37 ohm, quarter wavelength micro-stripline, 0.18" (4.5mm) wide, 1.16" (29.5mm) long, gap in center for installing C2

Z14 150 ohm shunt inductive reactance consisting of 50 ohm, 0.2-wavelength microstripline, 0.10" (2.5mm) wide, 0.96" (24.5mm) long

fig. 15. Full-size printed-circuit layout for the two-stage 1296-MHz amplifier. For identification of microstriplines Z13 and Z14, see fig. 14.

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