



A microstripline design of a 50 mHz power amplifier
by Louis LeeGrande Barrett

A thesis submitted in partial fulfillment of the requirements for the degree of MASTER OF SCIENCE
in Electrical Engineering
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Abstract:

Microstripline is a name given to the use of double-sided, copper-clad, circuit board for construction of very-high-frequency (VHF), ultra-high-frequency (UHF) and microwave transmission lines, inductors and capacitors. Microstripline differs from stripline construction in that stripline uses two ground-planes with a smaller main conductor sandwiched between them. Microstripline employs a single ground-plane. Microstripline is constructed by etching one side of the double-clad circuit board to a width corresponding to a desired characteristic impedance while the opposite, wider side acts as a ground-plane. If the width is narrow, the series inductance characteristic of the transmission line will dominate. Conversely, a wide width would cause the shunt capacitance of the line to dominate. By cascading microstripline sections of the proper widths, networks may be synthesized. This type of construction lends itself well for the application of VHF radio frequency amplifiers, where tab type transistors are used. Advantages of microstripline construction as contrasted with the use of all discrete components are improvements in amplifier performance, plus ease in design and fabrication of the amplifier. An amplifier for the 50 mHz region was designed, built, and tested using microstripline techniques and is described here. Methods of measuring circuit board parameters, criteria for amplifier network design, and results are given. The amplifier performed admirably. Results are discussed and analyzed and suggestions made for further study.

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Date

25 May 1977

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by

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ABSTRACT

Microstripline is a name given to the use of double-sided, copper-clad, circuit board for construction of very-high-frequency (VHF), ultra-high-frequency (UHF) and microwave transmission lines, inductors and capacitors. Microstripline differs from stripline construction in that stripline uses two ground-planes with a smaller main conductor sandwiched between them. Microstripline employs a single ground-plane. Microstripline is constructed by etching one side of the double-clad circuit board to a width corresponding to a desired characteristic impedance while the opposite, wider side acts as a ground-plane. If the width is narrow, the series inductance characteristic of the transmission line will dominate. Conversely, a wide width would cause the shunt capacitance of the line to dominate. By cascading microstripline sections of the proper widths, networks may be synthesized. This type of construction lends itself well for the application of VHF radio frequency amplifiers, where tab type transistors are used. Advantages of microstripline construction as contrasted with the use of all discrete components are improvements in amplifier performance, plus ease in design and fabrication of the amplifier. An amplifier for the 50 mHz region was designed, built, and tested using microstripline techniques and is described here. Methods of measuring circuit board parameters, criteria for amplifier network design, and results are given. The amplifier performed admirably. Results are discussed and analyzed and suggestions made for further study.

INTRODUCTION

Microstripline (MSL) is a method of realizing LC networks by cascading transmission line lengths of different characteristic impedance. The transmission line sections are made by etching a narrow conducting width on one side of a piece of double-sided, copper-clad, printed circuit board while the foil on the side directly opposite is either etched several times larger in width or left unetched. The wider side acts as a ground-plane for the other conductor.

MSL is a simplified version of "stripline" network synthesis. In stripline methods, two ground-planes are used with the narrowly etched center conductor sandwiched between them.

MSL is mostly applied in the microwave and ultra-high-frequency (UHF) regions (450 MHz and above) in either a complete network or commonly as part of a hybrid network also involving discrete components.

The purpose of the following study was to apply MSL to the design of a 50 MHz region, very-high-frequency (VHF) amplifier. The study was done to explore the feasibility of MSL at VHF and perfect design methods for use by experimenters who do not have expensive lab equipment.

The first part of this paper deals with methods used and data taken on several MSL widths. Once characteristic impedance and velocity data had been taken, it was tested for validity by building a low pass network also described.

The second part explains procedures used in the design and testing of a 52 mHz amplifier. This frequency was chosen because of availability of equipment to make measurements and test the amplifier.

G-10 epoxy-glass circuit board was used throughout with a dielectric thickness of 1 mm. The methods described in this study may be used to also establish line characteristics and design networks using other dielectrics and thicknesses.

An advantage to MSL is ease in constructing tuned networks using conventional printed circuit board methods. Once an accurate prototype is developed, it may be easily duplicated with no further tuning needed.

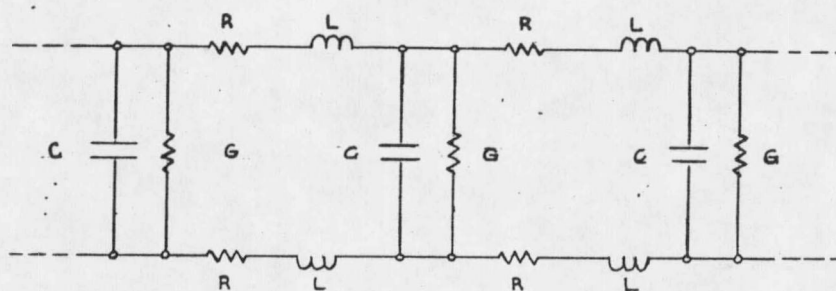
PROBLEM ANALYSIS

If MSL sections were to be used to synthesize networks, design equations had to be established. The logical place to begin was with the transmission line equations. Relationships were sought which would relate capacitance or inductance per unit length to measurable parameters. Such relationships would then lead to the design equations for the realization of networks.

MSL is similar to coaxial line in that the conductor surface is purposely smaller in width than its associated ground-plane. With that in mind, the coaxial transmission line model was used to begin the derivation of design equations. Figure 1a shows the general model for a parallel, balanced transmission line (Potter and Fich 1963). The model is modified in Figure 1b to that of a coaxial, unbalanced line. The model consists of series inductance (L) and ac resistance (R) shunted by capacitance (C) and leakage conductance (G). If the line may be assumed lossless, the model may be simplified to that shown in Figure 1c. Only the series inductance and shunt capacitance comprise the line characteristics.

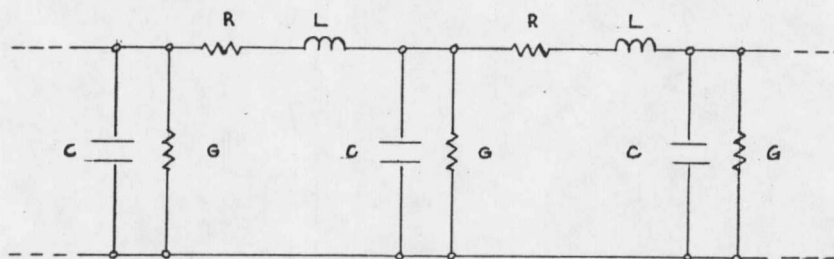
It was assumed the line was lossless initially. The lossless transmission line equations are then given by Potter and Fich (1963):

$$Z_0 = \sqrt{\frac{L}{C}} \quad (\text{Ohms}) \quad (1)$$



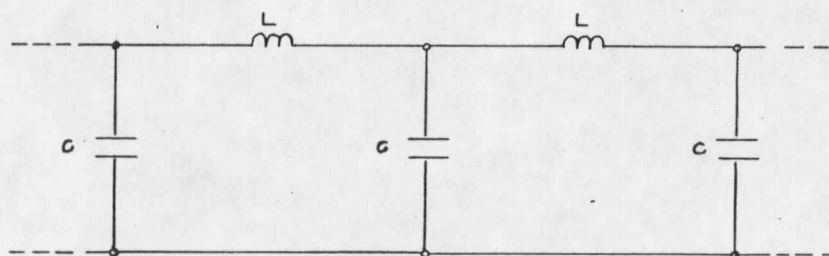
(a)

Parallel conductor transmission line with losses



(b)

Coaxial transmission line with losses



(c)

Lossless coaxial transmission line

Figure 1

Transmission Line Section Models

and

$$V = \frac{1}{\sqrt{LC}} \quad (\text{meters/second}) \quad (2)$$

where: L = Inductance per length (Henries/meter)

C = Capacitance per length (Farads/meter)

Z_0 = Characteristic line impedance (Ohms)

V = Phase velocity through the line (meters/second)

Solving for L and then C from equation (1):

$$L = Z_0^2 C \quad (\text{Henries/meter}) \quad (3)$$

and

$$C = \frac{L}{Z_0^2} \quad (\text{Farads/meter}) \quad (4)$$

If equation (3) is substituted into equation (2), the capacitance per unit length becomes:

$$C = \frac{1}{Z_0 V} \quad (\text{Farads/meter}) \quad (5)$$

Similarly, if equation (4) is substituted into equation (2), the inductance per unit length is:

$$L = \frac{Z_0}{V} \quad (\text{Henries/meter}) \quad (6)$$

As seen from equations (5) and (6), if the characteristic impedance (Z_0) and the phase velocity (V) could be established by experimental means, networks using series inductances and shunt capacitances could be synthesized by using the respective MSL lengths.

The next problem encountered was how to make the inductance in the transmission line dominate over the capacitance or vice-versa. Since the capacitance per unit length on the transmission line is static capacitance (Sams 1968), perhaps a clue could be found from the parallel plate capacitor equation.

Figure 2 illustrates the physical properties of an MSL section. The narrow main conductor of the line forms a parallel plate capacitor

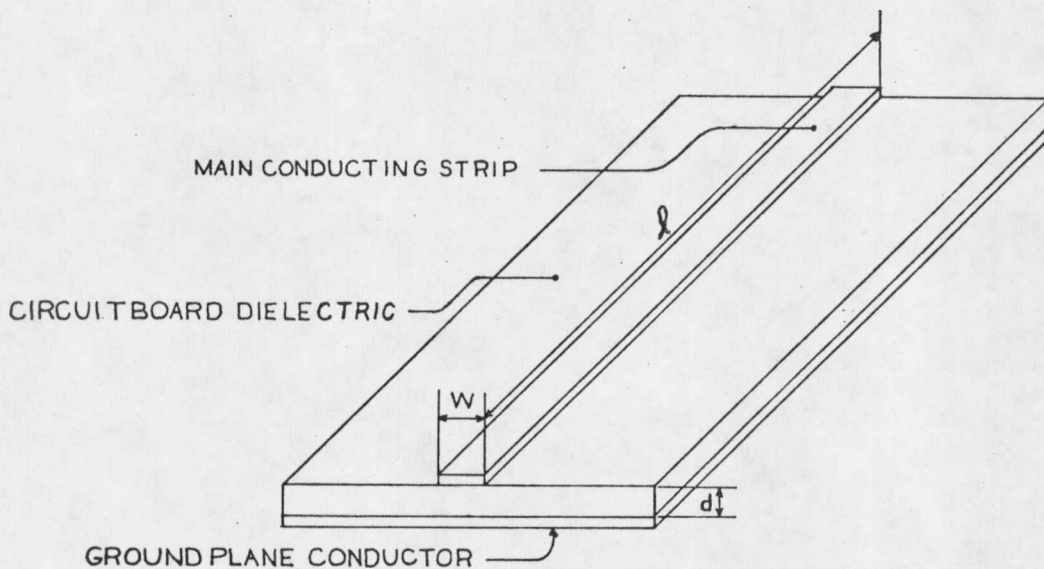


Figure 2

Geometry of an MSL Section

with the ground plane using the circuit board material as the dielectric between the plates. If fringing of the electric field

between the plates is neglected, the total capacitance is given by

Kraus and Carver (1973):

$$C' = \frac{\epsilon A}{d} \quad (\text{Farads}) \quad (7)$$

where: C' = Capacitance (Farads)

ϵ = Dielectric constant of the insulating medium (Farads/meter)

A = Surface area of the smallest plate (square meters)

d = Plate separation (meters)

The surface area (A) of the smallest plate is that of the main conductor. The area is given by:

$$A = \ell W \quad (\text{square meters}) \quad (8)$$

Substituting equation (8) into equation (7):

$$C' = \frac{\epsilon \ell W}{d} \quad (\text{Farads}) \quad (9)$$

Dividing through by ℓ the main conductor length:

$$\frac{C'}{\ell} = \frac{\epsilon W}{d} \quad (\text{Farads/meter}) \quad (10)$$

Since equation (10) shows units of Farads/meter, this is an expression for the capacitance per unit length. Then equation (10) becomes:

$$C = \frac{\epsilon W}{d} \quad (\text{Farads/meter}) \quad (11)$$

Examination of equation (11) reveals that if ϵ and d are constants, the capacitance per unit length is proportional to the

width of the conductor. Therefore, if the width is small, the capacitance per unit length will be also and the series inductance per unit length will begin to dominate the MSL characteristics.

At this point, the network synthesis appeared simplified. A method of arriving at the characteristic impedance and the phase velocity through the line for a given width needed to be established. If these two parameters were known, a direct conversion could be made using the previously derived equations for capacitance and inductance per unit length. If shunt capacitance was required to dominate, the width was made significantly wider than when series inductance was required.

In the above analysis, fringing was neglected for ease in calculations. It was found that fringing could not be generally neglected. More discussion is given later on this point.

PRELIMINARY DATA

Phase Velocity Measurements

The easiest parameter to measure was the phase velocity through the line. The experiment design began with the expression for velocity in terms of wavelength (Radio Society of Great Britain 1969):

$$V = f\lambda \quad (\text{meters/second}) \quad (12)$$

where: V = Phase velocity (meters/second)

f = Frequency (Hertz)

λ = Wavelength (meters)

When working in free space, V would be nearly the speed of light (3×10^8 meters/second) (Kraus and Carver 1973). In a dielectric material such as epoxy-glass circuit board, the velocity would be expected to be somewhat slower. The experiment design used both the expression for velocity, equation (12) above, and the fact that a quarter-wavelength transmission line acts as an impedance inverter (Atwater 1962). Sections of several conductor widths were etched using common circuit board etching techniques and cut to lengths of approximately .457 meter. The lines were fed using a female BNC coaxial connector as shown in Figure 3.

After an investigation of feeding or "launching" methods from coaxial lines to MSL at microwave frequencies (Farrell 1963), the

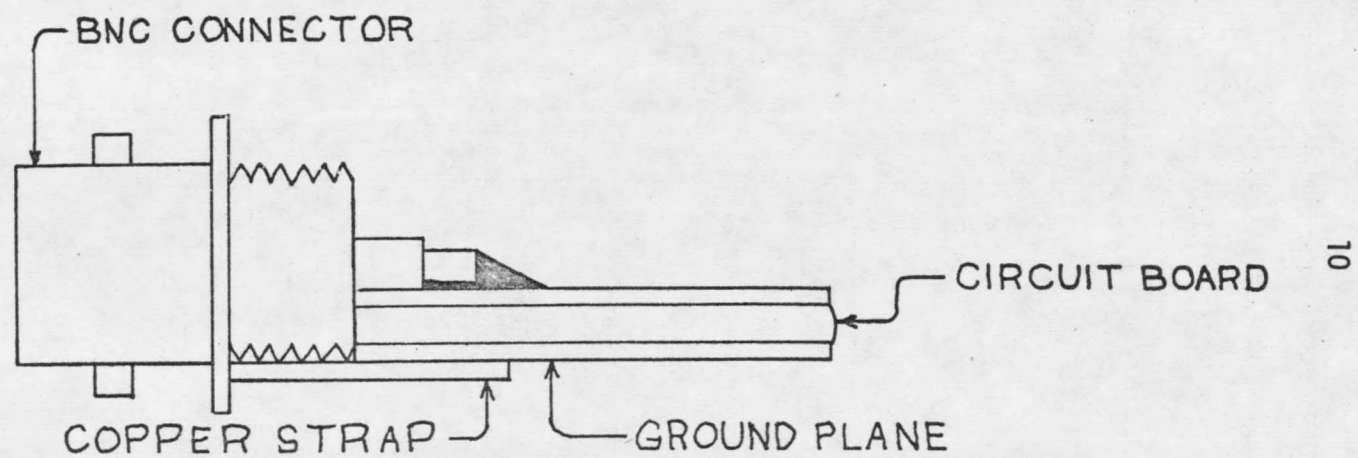


Figure 3

Coaxial Interfacing to an MSL Conductor
Using a BNC Connector

launching method used was assumed nearly lossless. This was due to the low frequency used (50 MHz region).

As previously stated, a quarter-wavelength transmission line section acts as an impedance inverter. This property also holds for odd multiples of a quarter-wavelength. If the output of a transmission line section was terminated in an open circuit, several frequencies could be found where the input of the line would appear nearly shorted. True input shorts would occur if an ideal transmission line were used in this configuration. The length of the line section would be equal to one quarter-wavelength of the lowest frequency at which this impedance inversion occurred. This lowest frequency was termed the quarter-wavelength resonant frequency (f_0).

The connector end of the open circuited MSL was connected to a Hewlett-Packard Model 4815A vector impedance meter and the frequency ranges were swept with increasing frequency from 500 kHz until the first impedance minimum was observed. The frequency where the impedance was a minimum and the impedance phase angle was zero degrees was noted as the quarter-wavelength resonant frequency. A frequency counter connected to the oscillator of the vector impedance was used to make the frequency measurements.

Once accurate length measurements were made of the transmission line sections, equation (12) was modified as shown below:

$$\ell = \frac{\lambda}{4}$$

where: ℓ = Conductor length (meters)

$$\lambda = 4\ell$$

and

$$V = 4f\ell \quad (\text{meters/second}) \quad (13)$$

Table 1 presents the data taken by this method.

Table 1

Experimental Measurement Results for MSL Phase Velocity
as a Function of Main Conductor Width

Width (mm)	f_0 (mHz)	ℓ (meters)	λ (meters)	V (meters/second)
.381	96.07	.4216	1.687	1.62×10^8
.762	87.15	.4580	1.832	1.59×10^8
1.580	83.43	.4580	1.832	1.53×10^8
2.540	80.21	.4580	1.832	1.47×10^8
3.810	76.85	.4580	1.832	1.41×10^8
6.350	71.80	.4570	1.829	1.31×10^8
12.700	65.49	.4320	1.727	1.13×10^8
25.400	53.62	.4310	1.723	9.24×10^7

An alternate method where a vector impedance meter is not available is shown in Figure 4. The results of this method were

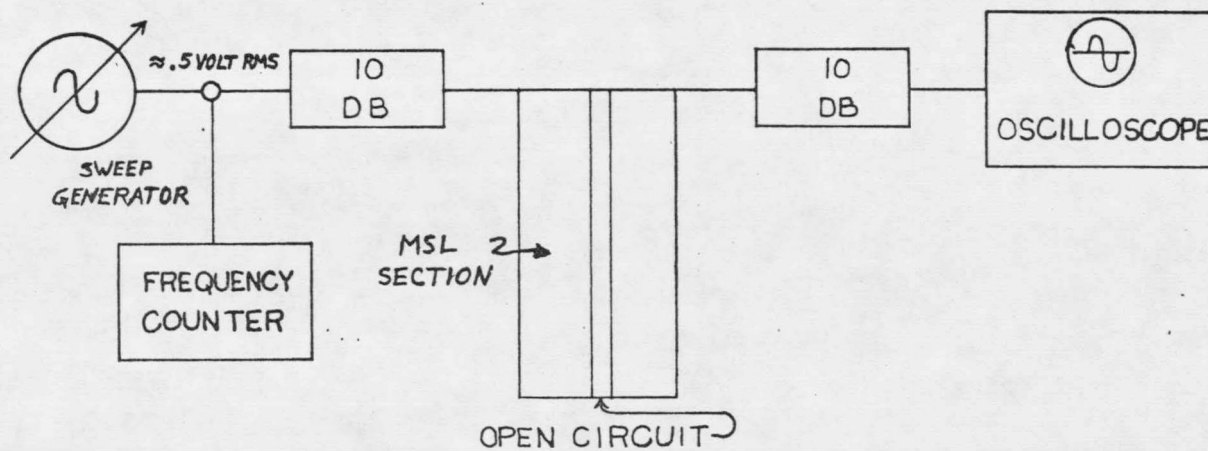


Figure 4

An Alternate Method of Measuring MSL Phase Velocity

in close agreement with those taken with the vector impedance meter. A sweep oscillator or signal generator, monitored by a frequency counter, fed a 10 dB attenuator for isolation purposes. The output of the attenuator was "tee-ed" with the coaxial input end of the open circuited MSL section and a second isolation 10 dB attenuator. The output of the second attenuator drove the vertical input amplifier of an oscilloscope. The signal source was carefully tuned in frequency until the first and lowest in frequency minimum voltage point was found as observed on the oscilloscope. At quarter-wavelength resonance, the open circuited MSL exhibited a short at the coaxially connected end of the MSL section. As a result, the voltage was at a minimum in the coaxial line. After noting the generator frequency and accurately measuring the MSL conductor length, equation (13) was used to determine the phase velocity (V) along the line. The system tested in this method used 50 Ohms as a characteristic impedance for convenience. Care was taken to keep coaxial interconnection lines short to prevent inaccuracies due to possible coaxial line resonance.

Characteristic Impedance Measurements

Once the phase velocity was established by the methods described previously, only characteristic impedance (Z_0) data was needed before network synthesis could take place.

The method used to determine line characteristic impedance was that of determining the geometric mean of the open circuit and short circuit impedances (Potter and Fich 1969). Characteristic impedance is given by:

$$Z_o = \sqrt{Z_{sc} Z_{oc}} \quad (\text{Ohms}) \quad (14)$$

where: Z_o = Characteristic impedance (Ohms)

Z_{oc} = Line input impedance with opposite end terminated in an open circuit (Ohms)

Z_{sc} = Line input impedance with opposite end terminated in a short circuit (Ohms)

The vector impedance meter was used again for these measurements.

Care had to be taken to obtain accurate data. When short circuiting the line section for short circuited impedance measurements, the entire width of the main conductor was connected to the ground-plane side. If a narrower connection were made, it could have appeared as a terminating inductance.

Data points were taken at three different frequencies to verify results. It was noted that these points had to be several megaHertz away from the quarter-wavelength resonance frequency to avoid resonance effects and, hence, error. For this reason, some higher frequency data points were thrown out because of the obvious effects of resonance.

The results of these measurements are presented in Table 2. An average of the data was taken and used as the characteristic impedance for each width. It was noted that the impedance phase angles were very close to zero degrees. That confirmed the previous assumption that such lines are nearly lossless.

Another method tried was to terminate the MSL section in actual carbon resistors. If the resistor used was near the characteristic impedance, the vector impedance meter could be swept over several decades of frequency with little change of the line input impedance phase from zero degrees. A small departure from an impedance phase angle of zero degrees with large frequency changes meant the carbon resistor value was close to the value of the characteristic impedance. This method was time consuming and cumbersome. Also, accurate results were only obtained for large main conductor widths. As a result, this method of impedance determination was abandoned.

Data Evaluation

After the characteristic impedances and velocities per width were established, a method was devised to test the equations for inductance and capacitance per unit length (equations (5) and (6)). The method used was to build a low pass filter by implementing the measurement data through equations (5) and (6). If a low pass

Table 2

Experimental Measurement Results for MSL Characteristic Impedance
as a Function of Main Conductor Width

Width(mm)	f(mHz)	Z_{oc} (Ohms)	Z_{sc} (Ohms)	Z_o (Ohms)
.381	20	298 $\angle -90^\circ$	40 $\angle 90^\circ$	109.2 $\angle 0^\circ$
	30	190 $\angle -90^\circ$	64 $\angle 90^\circ$	110.3 $\angle 0^\circ$
	40	134 $\angle -90^\circ$	95 $\angle 90^\circ$	112.8 $\angle 0^\circ$
				Z_o Ave. = 110.1 Ohms
.762	20	235 $\angle -90^\circ$	36 $\angle 90^\circ$	91.8 $\angle 0^\circ$
	40	102 $\angle -90^\circ$	85 $\angle 90^\circ$	93.1 $\angle 0^\circ$
	60	48 $\angle -90^\circ$	190 $\angle 90^\circ$	95.5 $\angle 0^\circ$
				Z_o Ave. = 93.5 Ohms
1.580	40	73.5 $\angle -90^\circ$	63 $\angle 90^\circ$	68.1 $\angle 0^\circ$
	60	32 $\angle -90^\circ$	140 $\angle 90^\circ$	67.2 $\angle 0^\circ$
	108	32 $\angle 90^\circ$	151 $\angle -90^\circ$	69.5 $\angle 0^\circ$
				Z_o Ave. = 68.3 Ohms
2.540	20	135 $\angle -90^\circ$	22.5 $\angle 90^\circ$	55.1 $\angle 0^\circ$
	40	56 $\angle -90^\circ$	52 $\angle 90^\circ$	54.0 $\angle 0^\circ$
	50	37 $\angle -90^\circ$	74 $\angle 90^\circ$	52.3 $\angle 0^\circ$
				Z_o Ave. = 53.8 Ohms
3.810	20	105 $\angle -90^\circ$	18.5 $\angle 90^\circ$	44 $\angle 0^\circ$
	40	44 $\angle -90^\circ$	44 $\angle 90^\circ$	44 $\angle 0^\circ$
				Z_o Ave. = 44.0 Ohms
6.350	20	74 $\angle -90^\circ$	14 $\angle 90^\circ$	32.2 $\angle 0^\circ$
	40	29.5 $\angle -90^\circ$	34 $\angle 90^\circ$	31.7 $\angle 0^\circ$
	50	17.5 $\angle -88^\circ$	46 $\angle 90^\circ$	28.4 $\angle 1^\circ$
				Z_o Ave. = 30.8 Ohms
12.700	15	61 $\angle -90^\circ$	6.8 $\angle 90^\circ$	20.4 $\angle 0^\circ$
	20	44 $\angle -90^\circ$	9.2 $\angle 90^\circ$	20.1 $\angle 0^\circ$
	30	26 $\angle -90^\circ$	14 $\angle 90^\circ$	19 $\angle 0^\circ$
				Z_o Ave. = 19.8 Ohms
25.400	15	31.5 $\angle -90^\circ$	4.6 $\angle 90^\circ$	12 $\angle 0^\circ$
	20	22 $\angle -90^\circ$	6.2 $\angle 90^\circ$	11.7 $\angle 0^\circ$
	30	11.7 $\angle -88^\circ$	9.5 $\angle 90^\circ$	10.6 $\angle 1^\circ$
				Z_o Ave. = 11.4 Ohms

filter could be synthesized using these parameters and the corner frequency be accurately predicted, then the design equations could be assumed correct.

The corner frequency of a low pass filter, which is also known as the half-power frequency, is the frequency where the output voltage is 70.7 percent the input drive voltage.

A low pass "Tee" filter was built to test the design equations and data. The schematic diagram of the filter is shown in Figure 5.

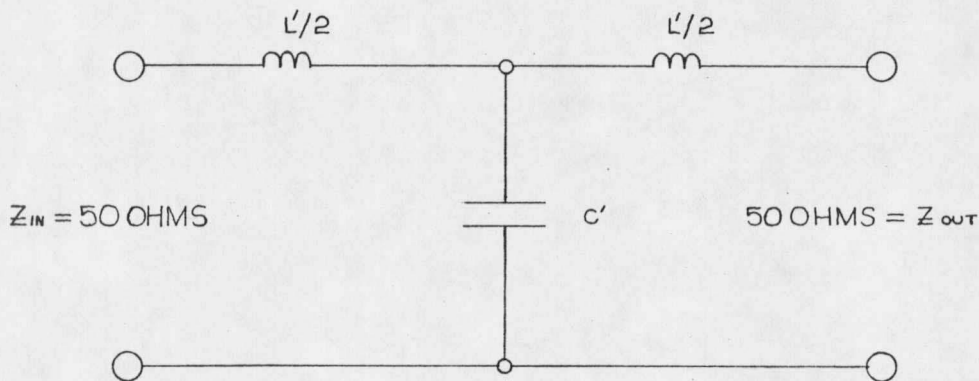


Figure 5

Schematic Diagram of "Tee" Filter Used to Test
Experimental Data and Design Equations

The inductance and capacitance sections in MSL are set apart only by abrupt transitions in section characteristic impedances. The larger the transition in characteristic impedance is between

cascaded line sections, the better each section will exhibit the desired characteristics whether inductive or capacitive.

There are limitations to the ratio of impedances between sections, however. One consideration is that of physical size of the MSL main conductor. If the main conductor is to be narrow to perform as a high impedance inductive section, there is a limitation on how narrow a circuit board foil may be etched. If the main conductor is to be wide and act as a low impedance capacitive section, the size could be too large and cumbersome to be practical.

In summary, for proper performance of the inductive and capacitive MSL sections, the ratio of characteristic impedances between cascaded MSL sections must be kept to a maximum within size and manufacturing tolerances (Howe 1974).

Though the impedance of MSL sections increases as the main conductor width narrows, this is not a linear function as will be discussed later. As a result, the ratio of impedances between sections should not be confused with the main conductor width ratio.

For the low pass "Tee" filter tested, two 93.5 Ohm (.762 mm width) sections were chosen as inductors and a 30.8 Ohm (6.35 mm width) section was chosen for the capacitor. The impedance ratio between sections was 3.

Solving for the inductance per unit length of the .762 mm line using equation (6):

$$\begin{aligned} L &= \frac{Z_0}{V} \quad (\text{Henries/meter}) \quad (15) \\ &= \frac{93.5 \text{ Ohms}}{1.6 \times 10^8 \text{ m/sec}} \\ &= .584 \text{ } \mu\text{H/meter} \end{aligned}$$

Likewise, for the 6.35 mm line, the capacitance per unit length using equation (5) was:

$$\begin{aligned} C &= \frac{1}{Z_0 V} \quad (\text{Farads/meter}) \quad (16) \\ &= \frac{1}{(30.8 \text{ Ohms})(1.31 \times 10^8 \text{ m/sec})} \\ &= 248 \text{ pF/meter} \end{aligned}$$

For a low pass "Tee" filter as shown in Figure 5, the design equations are given by Potter and Fich (1969):

$$L' = \frac{R_k}{\pi f_c} \quad (\text{Henries}) \quad (17)$$

and

$$C' = \frac{1}{\pi f_c R_k} \quad (\text{Farads}) \quad (18)$$

where: L' = Inductance (Henries)

C' = Capacitance (Farads)

f_c = Corner frequency of the filter (Hertz)

R_k = Characteristic impedance of the filter (Ohms)

$R_k = \sqrt{\frac{L'}{C'}}$

R_k was chosen as 50 Ohms for convenience in testing as all the test equipment available was 50 Ohms.

The corner frequency was chosen at 60 MHz because the filter pass-band covered the frequency range of interest.

Using equations (17) and (18) and noting each inductance section is $\frac{L'}{2}$, the required inductance and capacitance was found as shown:

$$\begin{aligned} \frac{L'}{2} &= \frac{R_k}{2\pi f_c} \quad (\text{Henries}) & (19) \\ &= \frac{50 \text{ Ohms}}{2\pi(6 \times 10^7 \text{ Hz})} \\ &= .133 \mu\text{H} \end{aligned}$$

and

$$\begin{aligned} C' &= \frac{1}{\pi f_c R_k} \quad (\text{Farads}) & (20) \\ &= \frac{1}{\pi(6 \times 10^7 \text{ Hz})(50 \text{ Ohms})} \\ &= 106 \text{ pF} \end{aligned}$$

At this point, the lengths of the low pass sections were calculated by dividing the results of equation (19) by the results of equation (15) and the results of equation (20) by the results of equation (16).

For inductance:

$$L_L = \frac{\frac{L'}{2}}{L}$$

$$= \frac{.133 \mu\text{H}}{.584 \mu\text{H/meter}}$$

$$= .228 \text{ meter}$$

For capacitance:

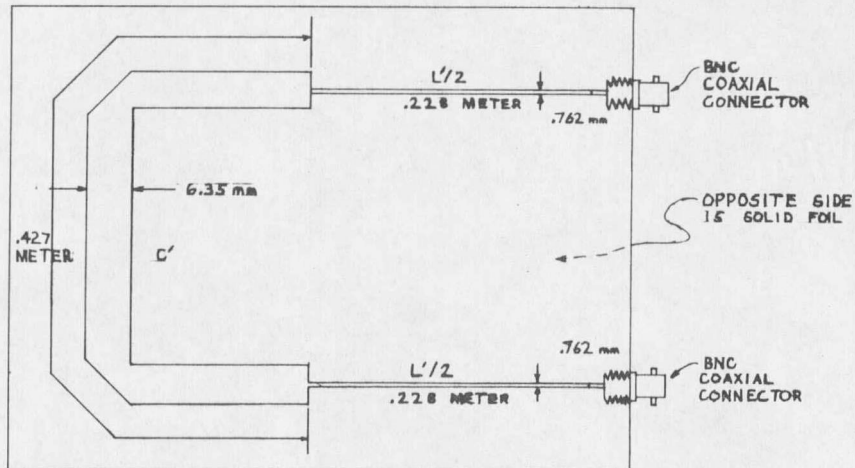
$$l_C = \frac{C'}{C}$$

$$= \frac{106 \text{ pF}}{248 \text{ pF/meter}}$$

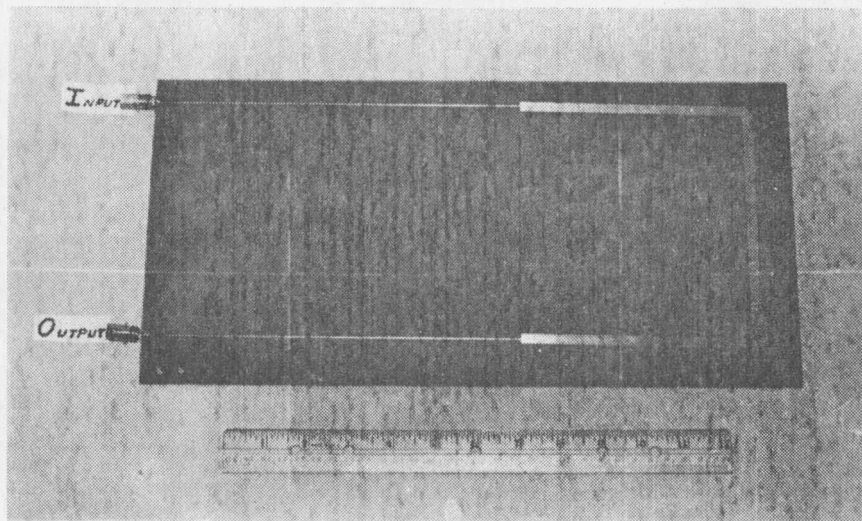
$$= .427 \text{ meter}$$

The low pass filter was constructed using standard photographic reduction, printed circuit board etching methods. The filter dimensions are illustrated in Figure 6a. A photograph of the finished filter is shown in Figure 6b.

As noted from Figures 6a and b, the filter was too long to etch in a straight line. For this reason, the MSL section was angled at 90 degrees to form a "U" shape to facilitate ease of construction. When right angle corners are made, the corners must be mitered at 45 degrees to decrease standing waves caused by boundary conditions at the corners (Howe 1974). Many references show only the outside corners mitered as shown in Figure 6a. Curved corners may be used but take up more area on the printed circuit board. Howe (1974) presents more data on this subject of mitered corners and standing waves at corners. No more discussion will be given here.



(a) Dimensional illustration



(b) Photograph

Figure 6

Dimensional Illustration and Photograph
of the Tested MSL "Tee" Filter

The filter was taped twice size on a piece of mylar. By using this method and reducing the taping to an actual size negative by photographic methods, any error made at twice size is also reduced in half. A larger layout would have been desirable to use due to this error reduction characteristic. However, the physical size of the 50 mHz filter would have made a larger layout too large to fit in the graphics camera available.

Once the printed circuit board was etched, the filter was tested by the method illustrated in Figure 7. Connections were made through female BNC coax connectors as previously shown in Figure 3.

A Hewlett-Packard Model 8554L spectrum analyzer was used as a measuring device. The spectrum analyzer was used because it is sensitive, has a broadband frequency response, and provides a 50 Ohm termination to outside connections.

The 10 dB, 50 Ohm attenuators were used as isolation to prevent loading of the generator and to provide proper filter termination. Care was taken to keep coax lengths short to prevent any effects of coax resonance from affecting the filter data.

Results of the measurements are given in Table 3. The actual corner frequency occurred at 67 mHz rather than 60 mHz as expected. That was an error of 12 percent across three elements in the filter. Upon remeasurement of the physical lengths of each MSL section, it

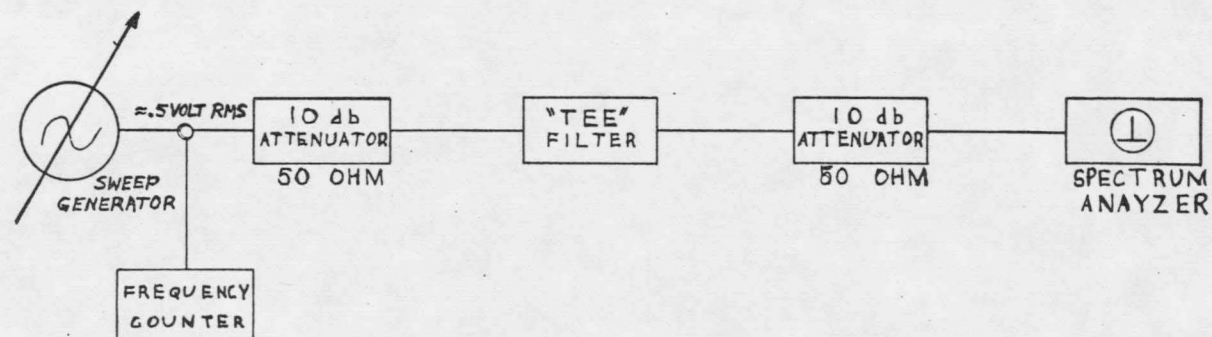


Figure 7

Method for Testing the MSL "Tee" Filter

Table 3

MSL "Tee" Filter Test Results

Frequency (mHz)	Attenuation (dB)
10*	0
20	0
40	-0.5
60	-2.6
60.5	-2.6
62	-2.6
64	-2.8
67	-3.0
70	-3.2
100	-7.0
150	-7.5
200	-4.0
250	-9.0
500	-8.0

*Frequencies less than 10 mHz showed 0 dB attenuation.

was observed that the capacitor conductor had lost some length and a small amount of surface area due to corner mitering. Both a shorter length and a loss of plate area resulted in less capacitance than expected causing a rise in the corner frequency. The inductor lengths had been constructed accurately.

The graph in Figure 8 illustrates the performance of the MSL "Tee" filter compared to a theoretical stripline and several actual stripline filters built and tested at microwave frequencies (Howe 1974). The MSL filter response was plotted with the dashed line. The stripline filter responses and theoretical response were plotted with solid lines. The MSL filter was a third order filter and the stripline filters were fifth order. For this reason, the MSL filter did not provide the attenuation in the stop-band as did the other stripline filters.

The ripple effect in the responses was due to using transmission line sections for synthesis. The abrupt impedance changes described earlier between the inductance and capacitance sections caused standing waves along the transmission line sections. As the drive frequency changed, the voltage maximas and minimas moved as a function of wavelength. As this occurred the output voltage correspondingly peaked and nulled. It was also noted that the ripple became less severe as the impedance ratios between line sections increased.

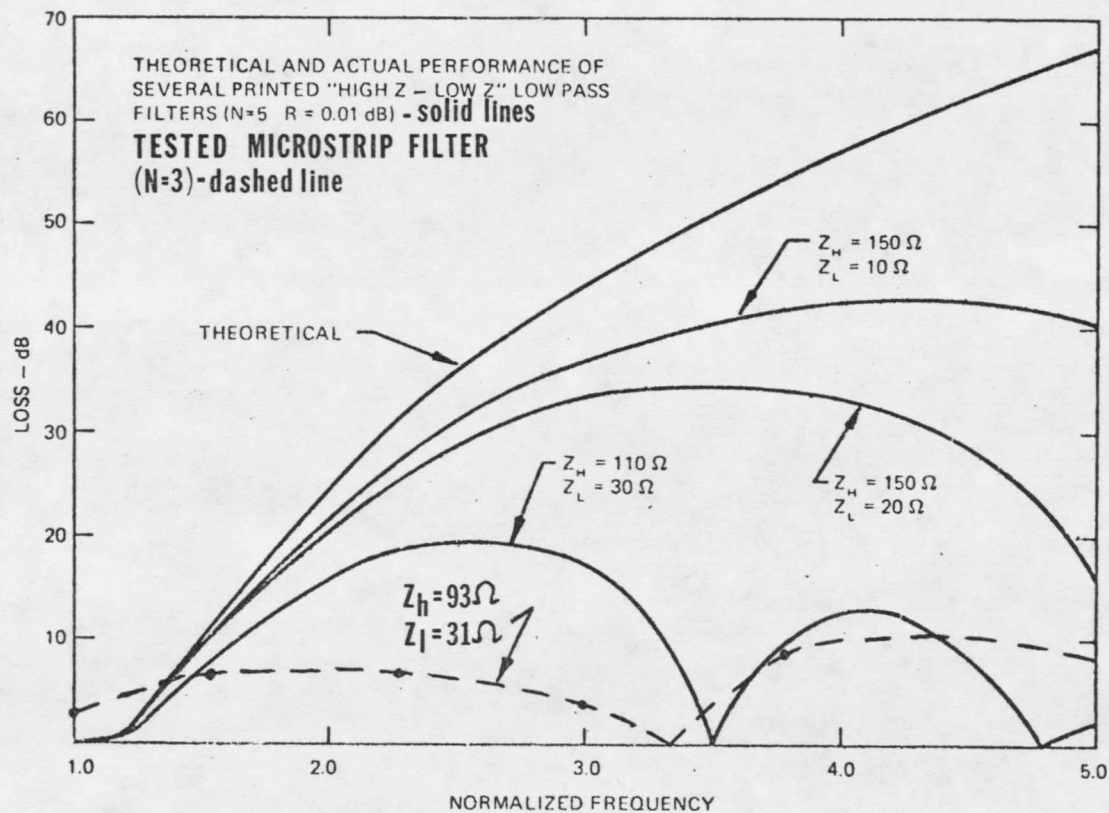


Figure 8

Results of the MSL "Tee" Filter Tests Compared to Several Stripline
and an Ideal Low Pass Filter Characteristics

An increase in impedance ratios between line sections in a given filter also caused the output response of that filter to better approximate the theoretical response. This aspect of impedance ratios between cascaded sections of line was noted to aid in further network synthesis.

The last test was to terminate the filter in 50 Ohms and measure the input impedance of the filter. If the characteristic impedance of the filter (R_k) was near 50 Ohms, the input impedance would be 50 Ohms also in the pass-band. The vector impedance meter was used to make the measurement and was swept from .5 mHz through 70 mHz. The pass-band impedance was 48 Ohms $\pm$ six degrees over the range from .5 mHz to 50 mHz. Above 50 mHz the impedance began to rise and the phase angle indicated the filter was becoming largely inductive on the input.

Before leaving the subject of parameter determinations, a final examination was made of the characteristic impedance (Z_0). Referring to equation (5) earlier, it was noted that the characteristic impedance could be derived in terms of the capacitance per unit length (C) and the phase velocity (V) through the line.

$$Z_0 = \frac{1}{CV} \quad (\text{Ohms}) \quad (21)$$

By substituting the expression for capacitance per unit length from equation (11) into equation (21) above, the characteristic

impedance is described as:

$$Z_0 = \frac{d}{\epsilon V W} \quad (\text{Ohms}) \quad (22)$$

where: Z_0 = Characteristic impedance (Ohms)

ϵ = Dielectric constant of the circuit board insulating material (Farads/meter)

V = Phase velocity of the wave through the line (meters/second)

w = Width of the main conductor (meters)

All of the variables in equation (22) were known except ϵ . To determine ϵ , several parallel plate capacitors were made by cutting several sizes of rectangles from double-sided unetched printed circuit board and measuring their capacitances on a Hewlett-Packard Model 4260A universal RLC bridge. If the area of the plates was large with respect to the spacing of the plates (the thickness of the printed circuit board dielectric), then it was assumed fringing could be neglected in the measurements. Equation (7) was used to calculate the data presented in Table 4.

The relative dielectric constant (ϵ_r) for this printed circuit board material is given as:

$$\epsilon_r = \frac{\epsilon}{\epsilon_0} \quad (\text{dimensionless})$$

where: ϵ_r = Relative dielectric constant

ϵ = Dielectric constant of the material (Farads/meter)

ϵ_0 = Dielectric constant of free space (Farads/meter)

Table 4

Measurement Data for the Dielectric Constant of G-10 Epoxy-glass
Printed Circuit Board of 1 mm Thickness by use of
Parallel Plate Capacitors at 1 kHz

Plate Area (sq. meters)	Capacitance (Farads)	Dielectric Constant (Farads/meter)
.0474	1.29×10^{-9}	2.71×10^{-11}
.0146	4.25×10^{-10}	2.91×10^{-11}
.0161	4.55×10^{-10}	2.83×10^{-11}
.0094	2.69×10^{-10}	2.86×10^{-11}
.0054	1.63×10^{-10}	3.03×10^{-11}
Average =		2.87×10^{-11}

Table 5

Comparison of Calculated Characteristic Impedance
to Measured Characteristic Impedance
for Several MSL Widths

Width (mm)	Measured Characteristic Impedance (Ohms)	Calculated Characteristic Impedance (Ohms)	Error (%)
.762	92	258.0	180
2.540	55	84.0	53
6.350	32	37.5	16

Then:

$$\epsilon_r = \frac{2.87 \times 10^{-11} \text{ Farads/meter}}{8.85 \times 10^{-12} \text{ Farads/meter}}$$

$$= 3.24$$

The previous measurements on the unetched rectangular printed circuit board capacitors were made at one kiloHertz. Further measurements made at microwave frequencies confirmed the relative dielectric constant as in the neighborhood of three and one half. Manufacturer specifications on the same circuit board dielectric, however, claimed an ϵ_r of five at one megaHertz. Because the measured data at both one kiloHertz and at microwave frequency agreed closely, the measured relative dielectric calculated above was used.

Once ϵ was determined, equation (22) was used to find the characteristic impedance of MSL sections with widths of .762, 2.54, and 6.35 mm. These calculated results are shown in Table 5 with the measured MSL section impedances for comparison. The error in calculation became larger as the width of the MSL main conductor decreased. From this data the conclusion was drawn that fringing may not be generally neglected. As the main conductor becomes narrow, fringing has more of an effect. No attempt was made to modify equation (22) to compensate for fringing.

AMPLIFIER DESIGN METHODS

Amplifier Design

The original goal of the project was to construct a power amplifier for the 50 MHz region using MSL sections to synthesize the input and output networks. Up to this point, the study had been directed toward derivation of design equations for MSL networks and testing their validity through experimental means. Since the measured data resulting from the previously described "Tee" filter experiment indicated the design equations to be useful, the design and construction of the power amplifier began.

The amplifier was used as the power output stage for an amateur radio transmitter operating on 52.525 MHz. The drive source for the amplifier was a commercially made transceiver which delivered 10 watts output as drive for the power amplifier. The mode used was narrowband FM.

The amplifier was designed to operate as a class C (non-linear) amplifier for three reasons. First, the amplification of FM requires no preservation of linearities in amplitude variations as do AM type signals. Second, class C operation would provide a greater theoretical collector efficiency than linear types of amplifiers. Third, class C is biased such that the quiescent operating point is below transistor cutoff.

The aspect of the quiescent point being below cutoff allowed the transistor base to emitter junction diode to be used as a method of bias generation. The transistor would remain cutoff until enough signal voltage appeared across this junction diode to forward bias it. For a silicon transistor, the forward bias voltage would be about .6 volt. Once forward biased, the junction diode would continue to rectify the input signal to provide base drive for the transistor.

The transistor to be used was a silicon BM70-12 manufactured by Communication Transistor Corporation (CTC). This transistor was chosen because it contained internal matching for easier interfacing with networks, it was constructed for use in MSL circuits in the VHF region, and it was immediately available. A photograph of the BM70-12 is shown in Figure 9. The BM70-12 was designed to operate from a 12 volt power source as a large signal amplifier with an output power of 70 watts. Although it was designed to operate in the 200 mHz region, no problems could be foreseen in using it at 52 mHz. Complete specifications for the transistor are given in Figures 10 and 11. The common emitter mode was chosen for operation of the transistor.

Consideration was given to three main areas during the design procedure. Discussed in the following sections are the methods

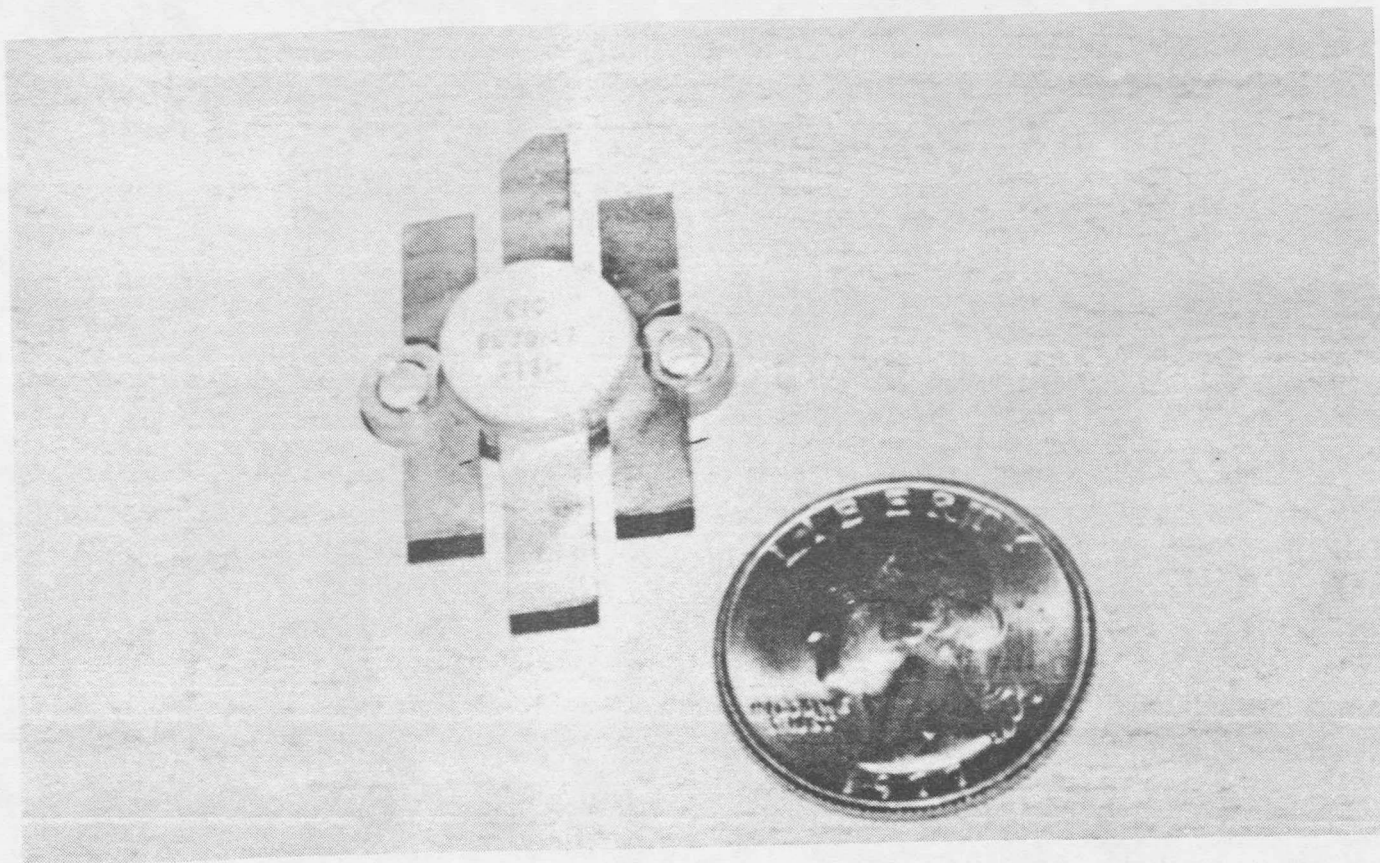


Figure 9

Photograph of the BM70-12 Transistor
Used in a 52 mHz Amplifier

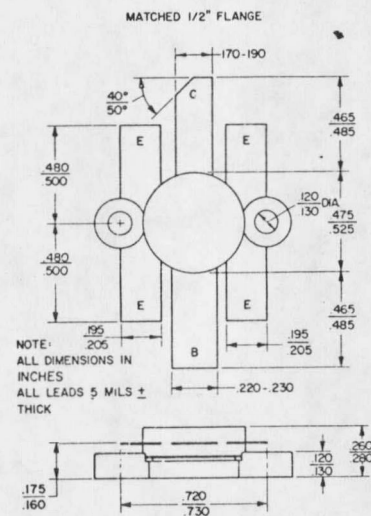
COMMUNICATIONS TRANSISTOR CORPORATION BM70-12 • 70 WATT

175 MHz • 12 VOLT RF POWER TRANSISTOR WITH INTERNAL MATCHING

GENERAL DESCRIPTION-Specifically designed to meet rugged mobile or portable radio requirements. Internal matching simplifies circuit design and gives excellent broadband performance.

FEATURES

- MAXIMUM RELIABILITY DUE TO SINGLE CHIP CONSTRUCTION.
- GUARANTEED TO WITHSTAND ∞ VSWR AT ALL PHASE ANGLES WHEN OPERATED AT RATED POWER WITH A 16V DC SUPPLY.
- EXCELLENT HERMETIC SEAL FOR EXTENDED LIFE IN HOSTILE ENVIRONMENTS.
- LOW THERMAL RESISTANCE -- 0.8° C/W.
- HIGH POWER GAIN.
- INTERNAL MATCHING FOR HIGHER INPUT IMPEDANCE AND LOWER Q.



ELECTRICAL CHARACTERISTICS

ABSOLUTE MAXIMUM RATINGS

MAXIMUM TEMPERATURES

Storage Temperatures	-65°C to +200°C
Operating Junction Temperatures	200°C
Lead Temperature (Soldering 8 seconds time limit)	260°C
≤ 1/32" from Ceramic	

MAXIMUM POWER DISSIPATION (Note 1)

Total Power Dissipation at 25°C Case Temperature	220 W
--	-------

MAXIMUM VOLTAGES AND CURRENT

BV _{CES}	Collector to Emitter Voltage	36 V
BV _{EBO}	Emitter to Base Voltage	4 V
LV _{CEO}	Collector to Emitter Voltage	18 V
I _C	Collector Current	20 A

Figure 10

General and Maximum Specifications of the BM70-12 Transistor

COMMUNICATION TRANSISTOR

BM70-12

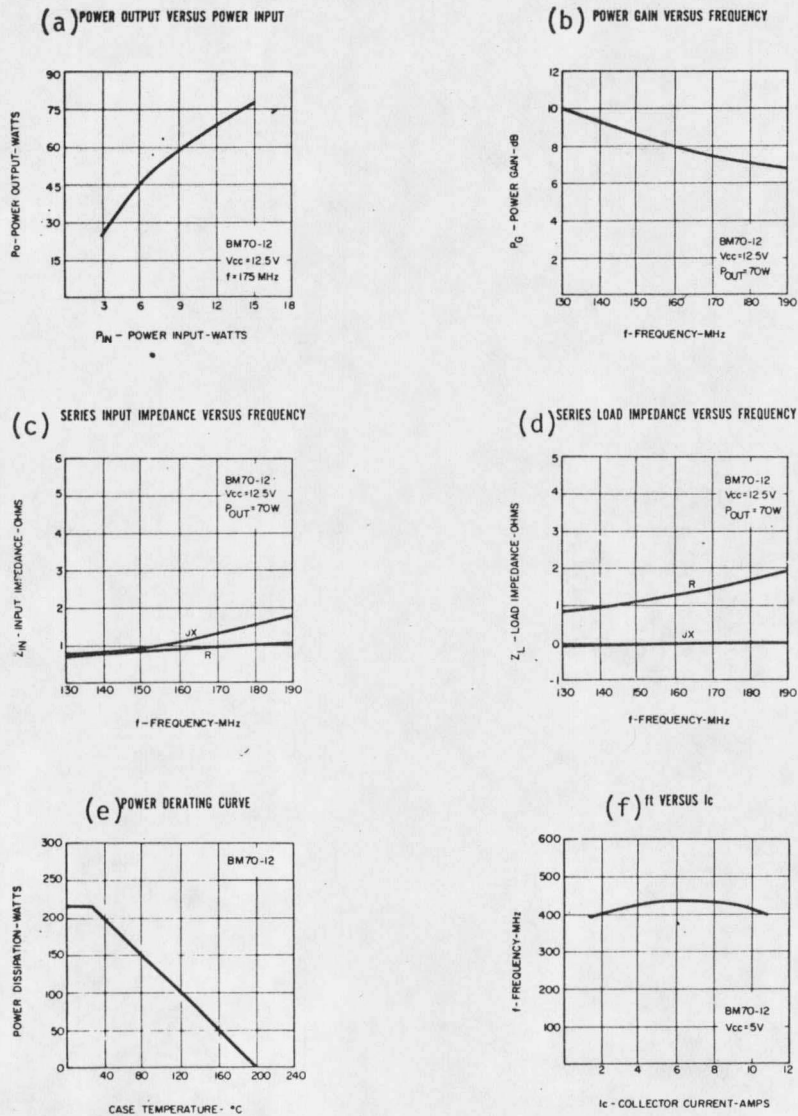


Figure 11

Amplification Characteristics of the BM70-12 Transistor

and reasoning used in dealing with the areas of network design, power dissipation, and stability of the amplifier.

Network Design

Rather than go through the steps of network synthesis for the transistor input and output networks, a Motorola application note (Davis AN-267) was used which gave computer solutions to several networks. These solutions were for matching a known series or parallel impedance through one of six networks to a non-reactive, 50 Ohm load with a specified loaded circuit "Q" (Q_L). Q_L is given by:

$$Q_L = \frac{f}{BW} \quad (\text{dimensionless})$$

where: Q_L = Q of the tuned network with all loads connected (dimensionless)

f = Resonant frequency of the tuned network (Hertz)

BW = Half-power bandwidth of the tuned network (Hertz)

Before use could be made of the computer solutions, the transistor base input impedance and the transistor collector output impedance had to be known. The input and output impedances of a large signal, class C amplifier, however, were hard to establish.

The general recommendation of the references consulted on this point was to apply power to the device and measure the full input drive current and voltage to the transistor base. Similarly, with the full drive power applied to the transistor base circuits, the

collector output voltage and current into a known load resistance would be measured. Once the respective voltage and currents were known, the base input and collector output impedances could be calculated. Such measurements required specialized equipment, however. A radio frequency signal generator with enough drive power, a known source impedance, and capability of tuning 52 mHz was needed. In addition, radio frequency current meters with dependable accuracy in the 52 mHz range were required. Because such test devices were not readily available, this method of impedance determination was not used.

The only other impedance data available was that shown in Figures 11c and 11d. Although the parameter curves shown were not extended below 130 mHz in the figures and were perhaps class A (linear) parameters, it was decided to try and make use of them. The reactive parts of the impedance curves in these figures dropped slightly with decreasing frequency as did the real parts. The real parts were assumed as one Ohm at 50 mHz, however, in both input and output cases for ease in later calculations. The reactive parts were projected from the curves to 50 mHz. As a result, a transistor base input impedance of $1 + j.25$ Ohms and a transistor collector output impedance of $1 - j.25$ Ohms were assumed.

Three of the previously mentioned six networks taken from the Motorola application note were eliminated because they were designed to match into a parallel source impedance. The transistor impedances determined above were series source impedances. Figure 12 illustrates the three remaining networks and component values calculated for the transistor collector output impedance using procedures from the Motorola application note. Similarly, the same networks (not shown) were calculated for the transistor base impedance. Solutions of these networks are detailed in the application note and will not be described here. A summary of network component values is given in Table 6 below.

Table 6

Amplifier Network Component Values Calculated
to Match a 50 Ohm Terminating Impedance
for a Loaded Q of 10

Component	Calculated Value
L_1	.032 μ H
L_2	.030 μ H
C_1	271 pF
C_2	271 pF
C_3	60 pF
C_4	60 pF

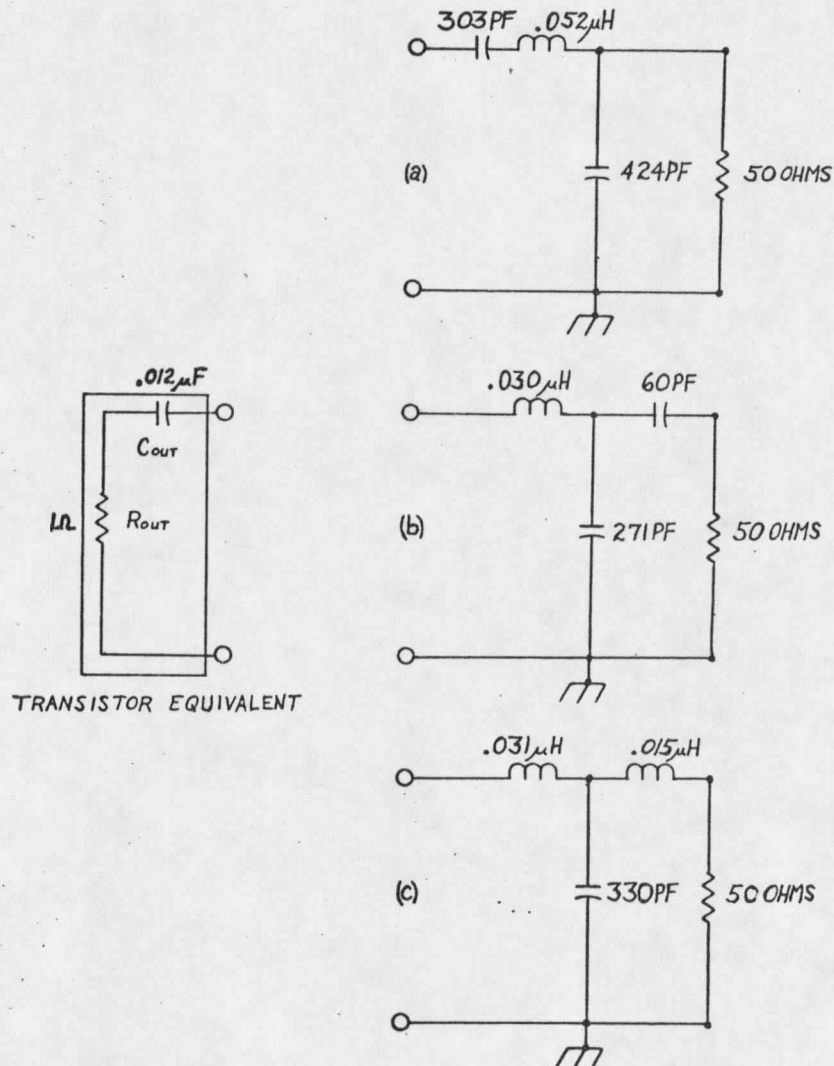


Figure 12

Possible Transistor Collector Matching Network Designs
for the 52 mHz Amplifier

Referring to Figure 12, networks a and b were the best choices as they presented capacitive output loading and tuning. This was an advantage over the inductive output circuit in c because capacitor errors could be compensated through addition of discrete trimmer capacitors if necessary. Circuit b was decided upon because the series capacitor in circuit a was larger than that in b. In the long run, the smaller capacitor in circuit b would contribute to a smaller physical printed circuit board size for the completed amplifier.

The amplifier schematic diagram using networks patterned after circuit b in Figure 12 is illustrated in Figure 13.

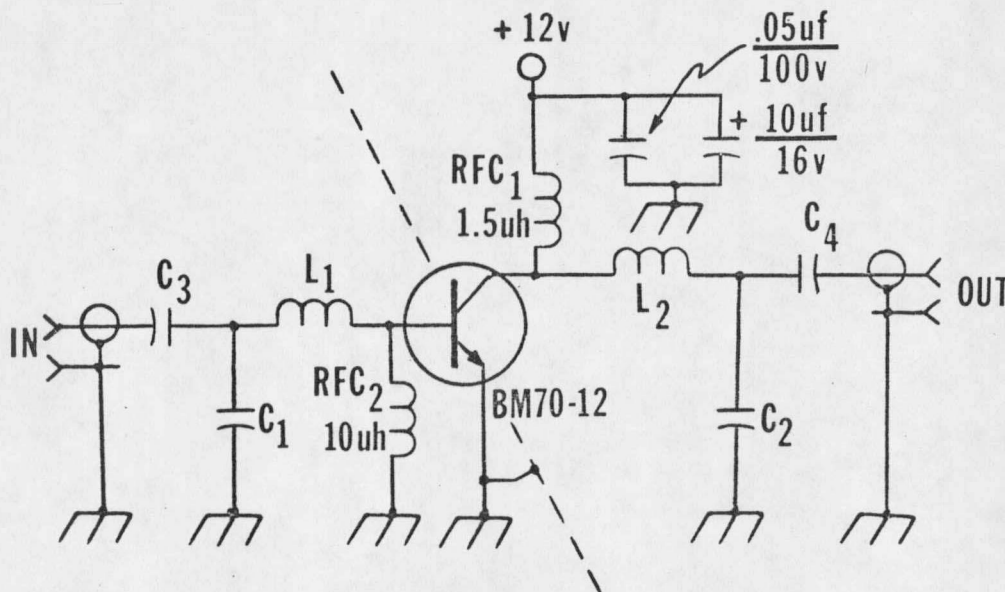


Figure 13

Schematic Diagram of the 52 MHz Amplifier

The input and output networks were to be constructed as MSL synthesized networks. Equations (5) and (6) were now used to calculate line lengths for the MSL sections. Only the series inductors and shunt capacitors could be calculated by this method. The calculations for the series capacitors are discussed later.

Recalling the earlier discussion on characteristic impedance ratios between cascaded MSL sections, it was discovered that impedance ratios from section to section produced the best network performances. For this reason, the networks for the amplifier were designed with a large characteristic impedance ratio between sections. An MSL main conductor width of .381 mm corresponding to a line characteristic impedance of 110.1 Ohms was chosen for inductor sections while an MSL main conductor width of 24.5 mm corresponding to a line characteristic impedance of 11.4 Ohms was chosen for capacitor sections. An impedance ratio between the capacitive and inductive MSL sections of 66.7 resulted from these choices of MSL sections. Due to examination of earlier measurements and reference data, such an impedance ratio was expected to perform well.

Using equations (5) and (6) in addition to the phase velocity and impedance data shown in Tables 1 and 2 respectively, the capacitance (C) and the inductance (L) per unit length were calculated:

$$\begin{aligned}
 L &= \frac{Z_0}{V} \\
 &= \frac{110.1 \text{ Ohms}}{1.62 \times 10^8 \text{ meters/second}} \\
 &= .680 \text{ } \mu\text{H/meter}
 \end{aligned}$$

and

$$\begin{aligned}
 C &= \frac{1}{Z_0 V} \\
 &= \frac{1}{(11.4 \text{ Ohms})(9.24 \times 10^7 \text{ meters/second})} \\
 &= 949 \text{ pF/meter}
 \end{aligned}$$

This data made the series inductance and shunt capacitance line lengths easy to calculate:

$$\begin{aligned}
 \ell_{L_1} &= \frac{L_1}{L} \\
 &= \frac{.032 \text{ } \mu\text{H}}{.680 \text{ } \mu\text{H/meter}} \\
 &= .047 \text{ meter}
 \end{aligned}$$

and

$$\begin{aligned}
 \ell_{L_2} &= \frac{L_2}{L} \\
 &= \frac{.030 \text{ } \mu\text{H}}{.680 \text{ } \mu\text{H/meter}} \\
 &= .044 \text{ meter}
 \end{aligned}$$

and

$$\begin{aligned} l_{C_2} &= l_{C_1} = \frac{C_1}{C} \\ &= \frac{271 \text{ pF}}{949 \text{ pF/meter}} \\ &= .286 \text{ meter} \end{aligned}$$

The series capacitors were made by etching two areas of foil on opposite sides of the printed circuit board to form parallel plate capacitors. Since the dielectric of the epoxy-glass was determined earlier, the foil sizes could be easily calculated. Referring to equation (7), the plate spacing and dielectric of each capacitor were constants implying the capacitance would be determined by the plate area. The plate area was calculated as shown below:

$$\begin{aligned} A_{C_3} &= A_{C_4} = \frac{dC}{\epsilon} \\ &= \frac{(.001 \text{ meter})(6 \times 10^{-11} \text{ Farads})}{(2.87 \times 10^{-11} \text{ Farads/meter})} \\ &= 2.09 \times 10^{-3} \text{ square meter} \end{aligned}$$

The input and output circuits were etched on a circuit board in such a way that a small piece of wire soldered through the board was used to connect the main conductor of C_1 and C_2 to the bottom plates of C_3 and C_4 , respectively. The inductance and capacitance sections were folded and mitered to conserve space in the layout. Care was taken to assure that the lengths were accurate through mitering to

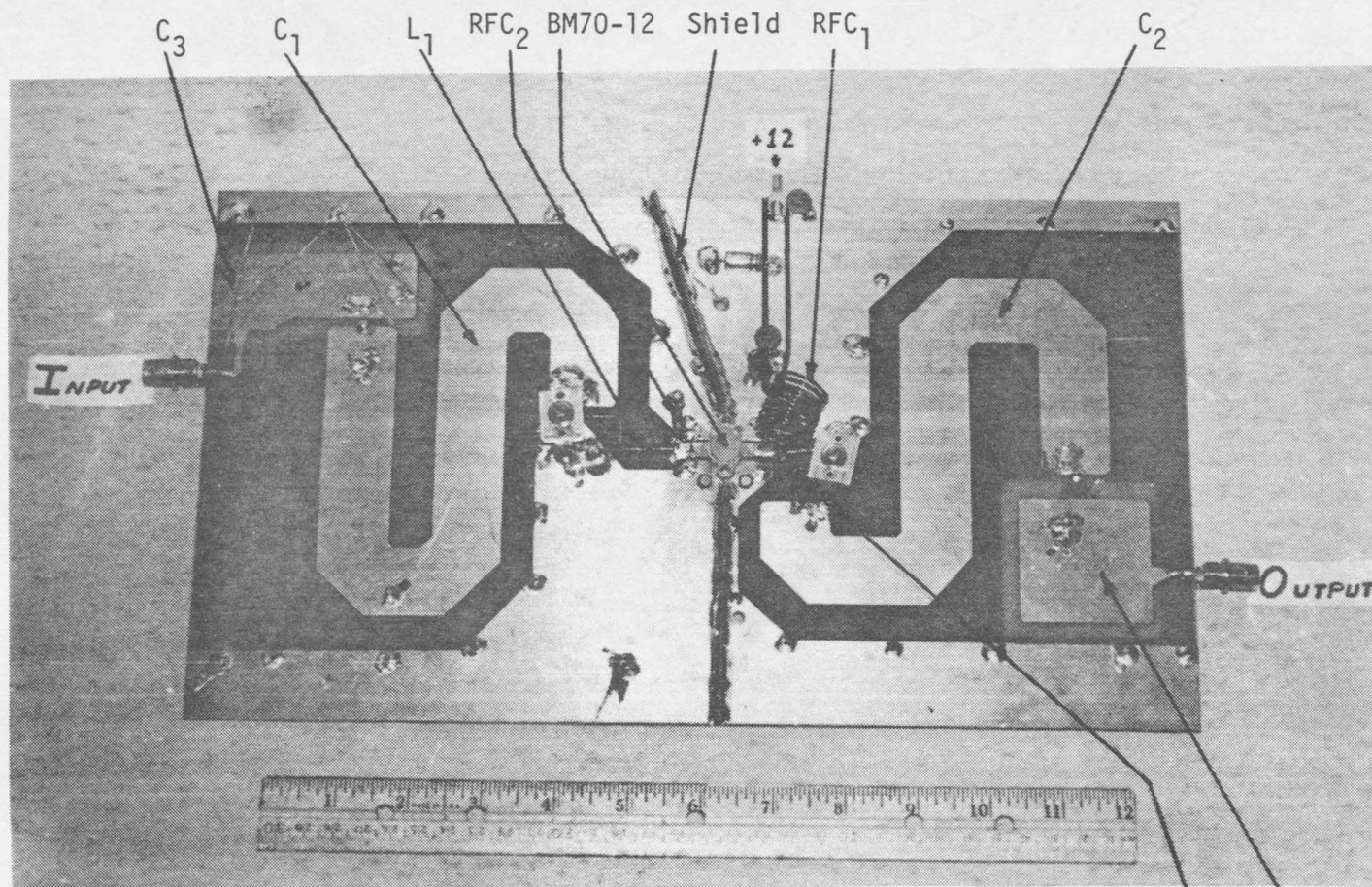
avoid the error encountered in the earlier MSL "Tee" filter experiment. Figure 14 presents a photograph of the etched amplifier networks with the transistor in place and the heat sink removed. The opposite side is solid foil with the exception of the areas under C_3 and C_4 which are the capacitor areas connected to the input and output connectors, respectively.

The foil area around the MSL sections is grounded to the foil side beneath by means of several small wires soldered through the board at small length intervals.

Before the transistor was placed in the circuit, the capacitor values were measured with a Hewlett-Packard Model 4260A RLC bridge. No easy method could be thought of to measure the inductance sections. Table 7 depicts the results.

Table 7
Amplifier Capacitor Measurements Compared
to the Calculated Values

Capacitor	Measured Capacitance (pF)	Calculated Capacitance (pF)
C_1	223	271
C_2	201	271
C_3	60	60
C_4	64	60



47

Figure 14

Photograph of the 52 mHz Amplifier
with Heat Sink Removed

The results of the measurements indicated C_3 as being the only capacitor to be accurate. Errors in the other measured capacitor values compared to those calculated were disturbing.

The difference between the measured and calculated values of C_4 was the easiest to explain. While remeasuring for physical length and size errors, C_4 was found to be slightly too large in plate area. Nothing was done to change C_4 at that time pending the actual operation of the amplifier. At that time some "fine tuning" was expected to be done.

The discrepancies found in the MSL capacitors, C_1 and C_2 , were not so easily explained. After reconsulting several references, a reason for the problems became evident. The corner mitering had been done incorrectly in both this circuit and the previously tested MSL "Tee" filter. Both inside and outside corners in these circuits had been mitered and no method had been used to govern the miter angles or lengths. Only the outside corners should have been mitered according to methods backed by experimental data given by Howe (1974). As a result of this oversight, the transmission lines became mismatched at the corners causing the lines to change characteristics. The result was performance other than expected in these capacitor sections. These sections were not adjusted until the amplifier was tested.

Some doubt was raised at this point as to the accuracy of the inductor MSL sections. As previously stated, no good method of measuring the inductors could be produced. As a result, it was decided to test the amplifier without further investigation of these sections.

Stability

Many papers and books were referred to for clues on how to predetermine the stability of the amplifier. Since the transistor was a large signal device, several problems presented themselves.

First, the "S" parameters, which are commonly used in design of VHF circuits, were eliminated because they are small signal parameters.

Second, most methods of calculating stability such as described in Linvill and Gibbons (1961) used "Y" parameters which are also small signal parameters.

Third, the amplifier was to operate in class C which is extremely nonlinear. This aspect added doubt to all parameter sets because transistor parameters are usually measured in the linear or active region of operation.

Without any method of predetermining stability, it was decided to construct the amplifier in such a way as to avoid effects which would cause instability. These effects could have

been field coupling between inductors from output to input or feedback through the reverse parameter of the power transistor.

As stated previously, large signal parameters are not readily available for transistors such as the BM70-12 used in this experiment. For this reason, feedback through the reverse direction of the amplifier transistor was assumed not to be a factor in the amplifier stability. Provisions for the addition of neutralizing components to the circuit board were provided, however, in case this assumption proved inaccurate during amplifier testing.

The remainder of the stability effort was spent in preventing network coupling between input and output. Specifically, L_1 and L_2 posed the greatest problems in terms of field coupling.

An experiment was conducted using two 10 cm lengths of MSL with conductor widths of 1.58 mm. The lengths were run parallel with 2 mm spacing between them. A 50 mHz signal source was connected to one line and an oscilloscope to the other line. The output voltage to the oscilloscope from the second line was 10 dB below the input voltage to the first line. These results illustrated that there is poor coupling between parallel MSL sections.

The results of this parallel line experiment seemed to indicate that if MSL lines would not cross-couple well at close spacing, then extreme spacing of lines would further reduce mutual coupling

possibilities. With this in mind, the input and output networks of the amplifier were placed opposite each other with as much spacing between the inductors, L_1 and L_2 , as possible.

To further cut down on inductor coupling a shield was constructed. Soldered to the top ground foil was a double-clad circuit board wall the height of the transistor which extended the entire width of the board. This strip provided shielding of the input network from the output network to aid in stability. The transistor separated one section of this shield from the other as may be seen from Figure 14.

Power Dissipation

Since the output power of the transistor was expected to be fairly large, the power dissipation likewise would be expected to be large. Some sort of heat sink had to be developed to transfer the heat to ambient air and cool the transistor so junction heat damage would not be sustained. Ambient temperature of the air was assumed at 25°C.

Calculations were made using methods described by Alley and Atwood (1973) and are presented in Appendix I.

The formulation of the correct heat sink size presented a problem. The method used resulted in a required heat sink area larger in size than the entire amplifier. This seemed unreasonable

(.45 square meter). A heat sink made from 3.2 mm bright aluminum approximately one-fourth the size calculated (.1 square meter) was used instead on a trial basis. If it became too hot, a larger heat sink would be tried.

The heat sink was attached to the transistor with a thin layer of silicon heat compound between them to aid heat transfer.

Amplifier Circuit Description

Referring to the completed schematic diagram in Figure 15, C_1 , C_3 , and L_1 match the 50 Ohm input drive source to the base of

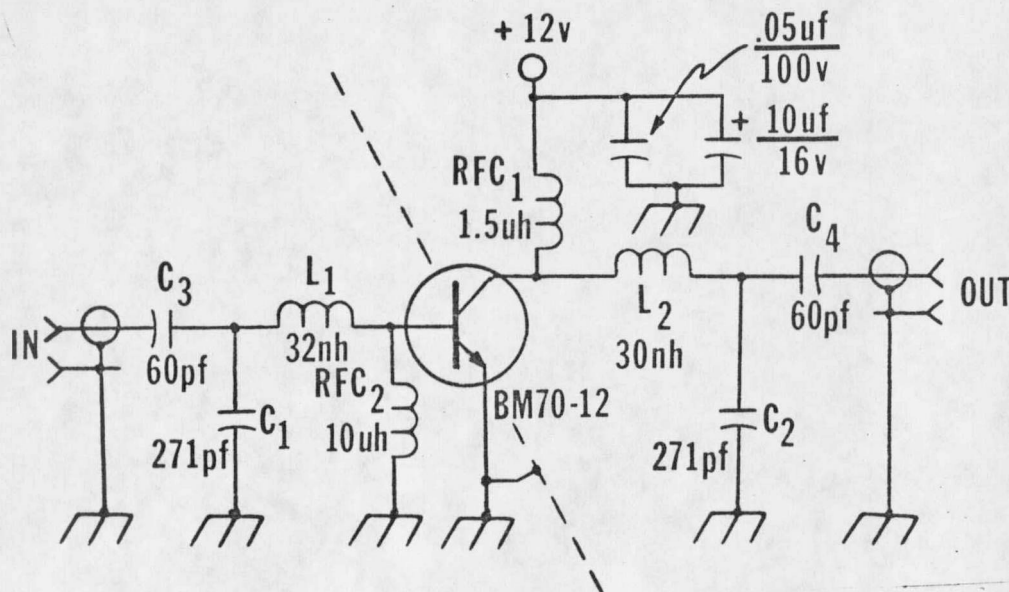


Figure 15

Completed Schematic Diagram of the 52 mHz Amplifier

the transistor. C_1 and L_1 resonate the circuit while C_3 tunes out any unwanted reactance at the input.

C_2 , C_4 , and L_2 match the 50 Ohm load impedance to the collector impedance of the transistor. C_2 and L_2 resonate here with C_4 matching the reactive component at the output terminals.

Because the transistor was to run in class C, no bias was applied to the transistor. A dc return for the transistor base was provided by a 10 μ H, commercially manufactured, radio frequency choke (RFC₂). This return was necessary due to the dc component of the class C wave form. No radio frequency signal was allowed to pass to ground due to the choke action.

Another radio frequency choke (RFC₁) was used to supply dc voltage to the transistor collector. It prevented radio frequency from entering the power supply. Any unwanted radio frequency voltage leaking through this choke was shorted to ground by the bypass capacitors on the power supply end of the choke. This choke was constructed from 12 turns of #18 AWG magnet wire about 25.4 mm long and 20 mm in diameter. The reactance of this choke measured 480 Ohms at 52 mHz or about 1.5 μ H.

RESULTS

The first test of the amplifier used a tunable VHF signal generator supplying about .5 watt through a variable 50 Ohm attenuator as a drive source to the input. The frequency of the generator was monitored by a frequency counter.

The amplifier output was terminated by a 50 Ohm, 20 dB power attenuator feeding a Hewlett-Packard Model 8554A spectrum analyzer. The attenuator provided a matched load impedance to the amplifier output terminal. Figure 16 illustrates the test set-up.

Twelve volts was applied to the amplifier and with a small amount of signal applied to the input, the frequency was swept in the neighborhood of 52 mHz. Only one resonance was observed at 57.328 mHz. A higher frequency was expected because the capacitor measured a bit small as was shown in Table 7. The network frequency error overall was nine percent over six MSL and printed circuit board components.

The loaded networks together produced a half-power bandwidth of 2 mHz or a Q_L of 28.7.

The final test was made by replacing the signal generator and variable attenuator by a 10 watt output, Model IC-6F, FM transceiver as a drive source. The output frequency was at 52.525 mHz. The

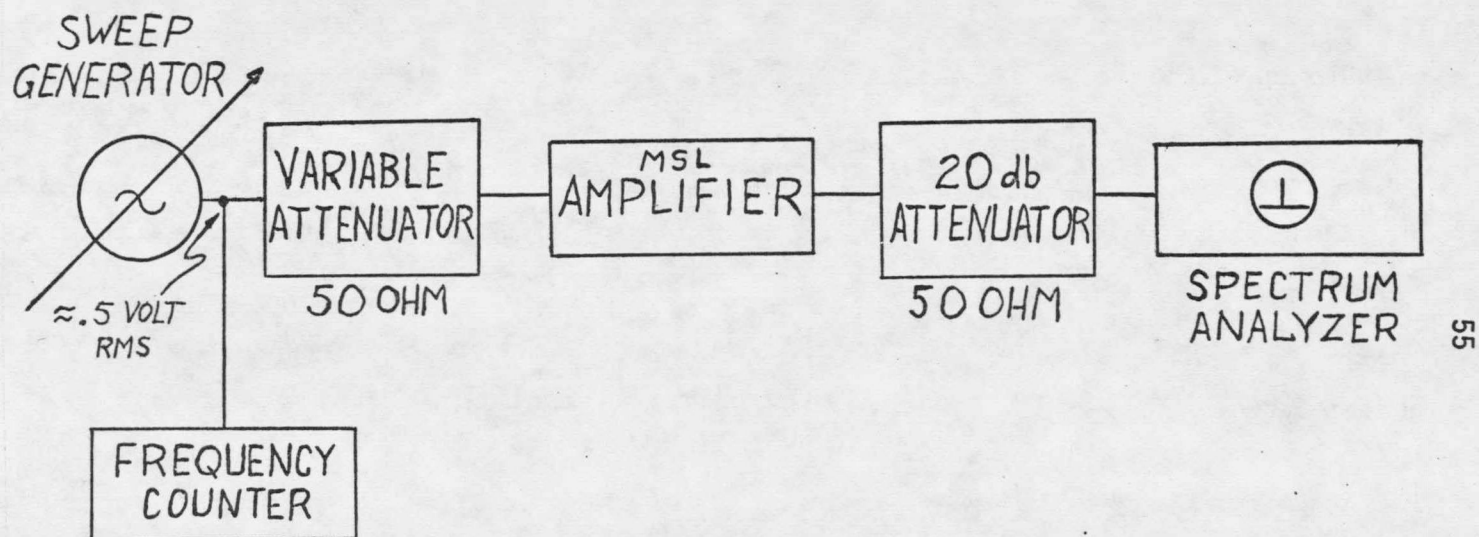


Figure 16

Initial Testing Method for the 52 mHz Amplifier

20 dB 50 Ohm power attenuator and spectrum analyzer remained connected to the output as before. Figure 17 illustrates the final test connections. Two wattmeters were inserted in series with the input and output of the amplifier so power gain could be measured.

Without tuning the networks, 12 volts dc was applied to the collector circuitry. With an input drive power of 10 watts, the amplifier output was 28 watts. This was decidedly not optimum performance.

After adding ARCO 461 compression trimmer capacitors, the amplifier was fine tuned. These capacitors paralleled C_1 and C_2 and were installed from the MSL section junctions of L_1-C_1 and L_2-C_2 to ground. With 12 volts dc to the collector and drive applied, the trimmers were tuned for maximum power output. This resulted in 35 watts of output for 10 watts of drive.

The ARCO 461 trimmers were removed and measured. The capacitor on the input measured 21 pF while that on the collector measured 65 pF. Referring to Table 7, both the sum of the ARCO 461 base capacitor plus C_1 and the sum of the ARCO 461 collector capacitor plus C_2 came out slightly less than the calculated 271 pF. This fact pointed to the inductor sizes being slightly too large. Any inductor error noted here was probably due to the mitering problems discussed earlier.

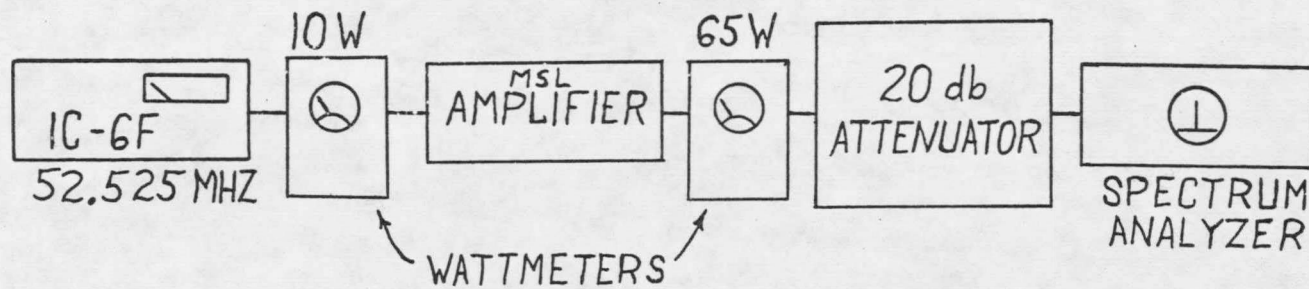


Figure 17

Final Testing Method for the 52 mHz Amplifier

To further increase the output power, the standing wave ratios on both input and output were tested using the wattmeters. There ideally should have been no reflected power on either the input or output. The output had no power reflected. The input, however, displayed nearly two watts reflected with 10 watts in the forward direction. Since C_3 was the matching capacitor, a small ARCO 460 compression trimmer capacitor was paralleled with it. The resulting reflected power at the input was greater even at minimum capacitance. This indicated that the capacitance of C_3 was too large to begin with. The ARCO 460 trimmer was removed and area of the C_3 foil on the circuit board was decreased by cutting diagonals through the foil. As seen from equation (7), decreasing the capacitor plate area decreased the capacitance of the capacitor.

As C_3 was decreased, the input reflected power began to drop as expected. The process of decreasing the plate area was continued until with 10 watts forward, only .1 watt was reflected.

With the input matched, optimum performance was achieved. With an input of 10 watts, the output power was measured at 65 watts (P_{out}) for a power gain of 8.13 dB. The input current was 9.5 amperes (I) at a collector voltage of 12 volts (E). The average input power (P_{in}) to the amplifier was:

$$P_{in} = IE \quad (\text{watts})$$

$$= (9.5 \text{ A})(12 \text{ V})$$

$$= 114 \text{ watts}$$

Efficiency of the amplifier is given by:

$$E_{ff} = 100 \times \frac{P_{out}}{P_{in}}$$

$$= 100 \times \frac{65 \text{ watts}}{114 \text{ watts}}$$

$$= 57 \text{ percent}$$

The expected efficiency of a class C amplifier is somewhat higher than this. However, there was no real attention given to the quiescent bias point of the transistor which contributed to the lower efficiency.

Figure 18 presents a photograph of the finished amplifier with the heat sink in place, the ARCO 461 trimmers across C_1 and C_2 shown, and the cuts in the top foil of C_3 to tune the input visible.

The final data taken was harmonic data from the spectrum analyzer. The data is plotted in Figure 19. Good reduction in amplitude of the harmonics was noted.

The only problem encountered was oversight in the collector circuit design. The inductor section, L_2 , was physically too small (.365 mm) to handle the high power radio frequency currents. It overheated and broke just after the final data was taken. The melting of L_2 was a case in point on why characteristic impedance

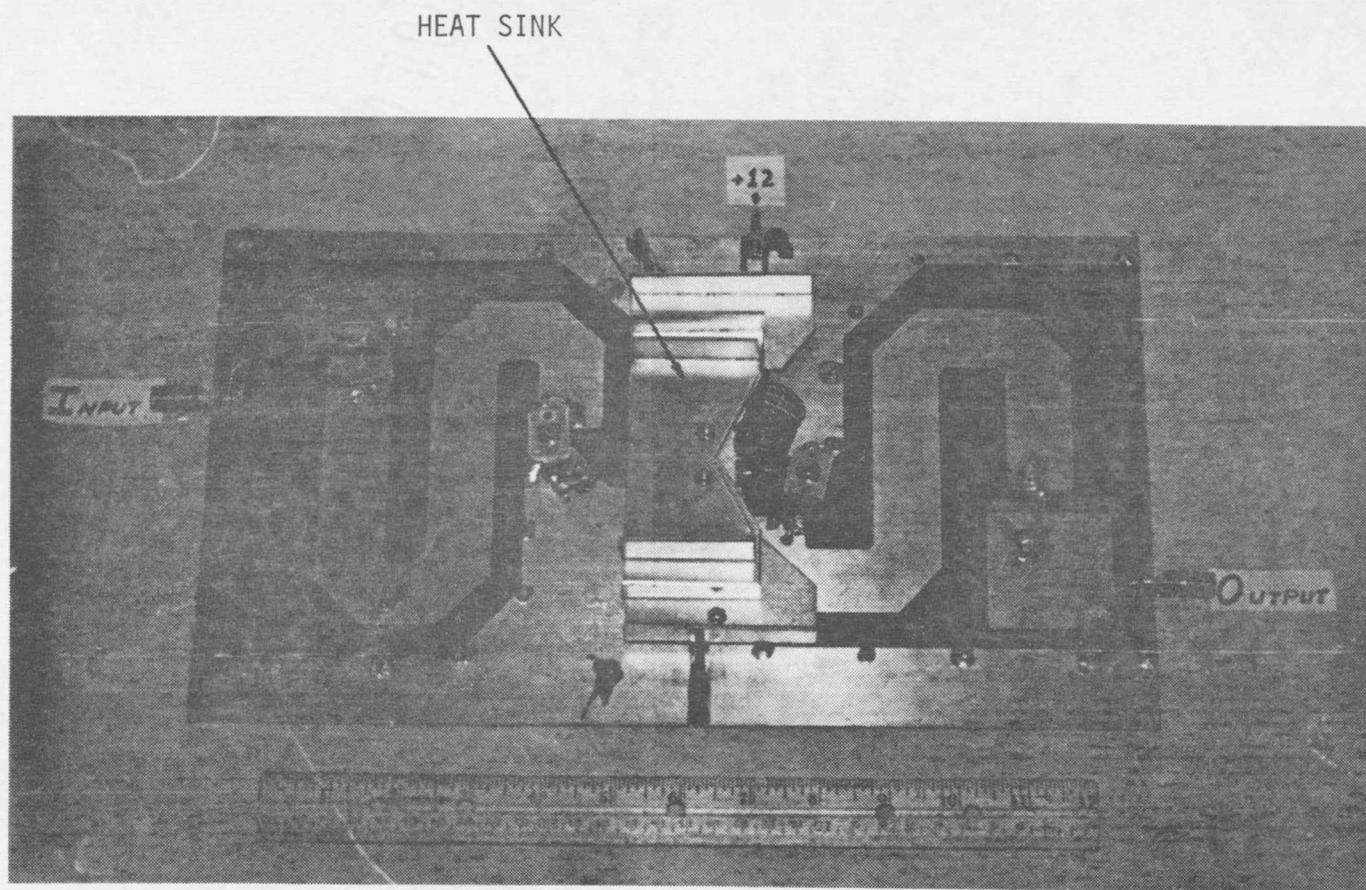


Figure 18

Photograph of the 52 mHz Amplifier
with Heat Sink in Place

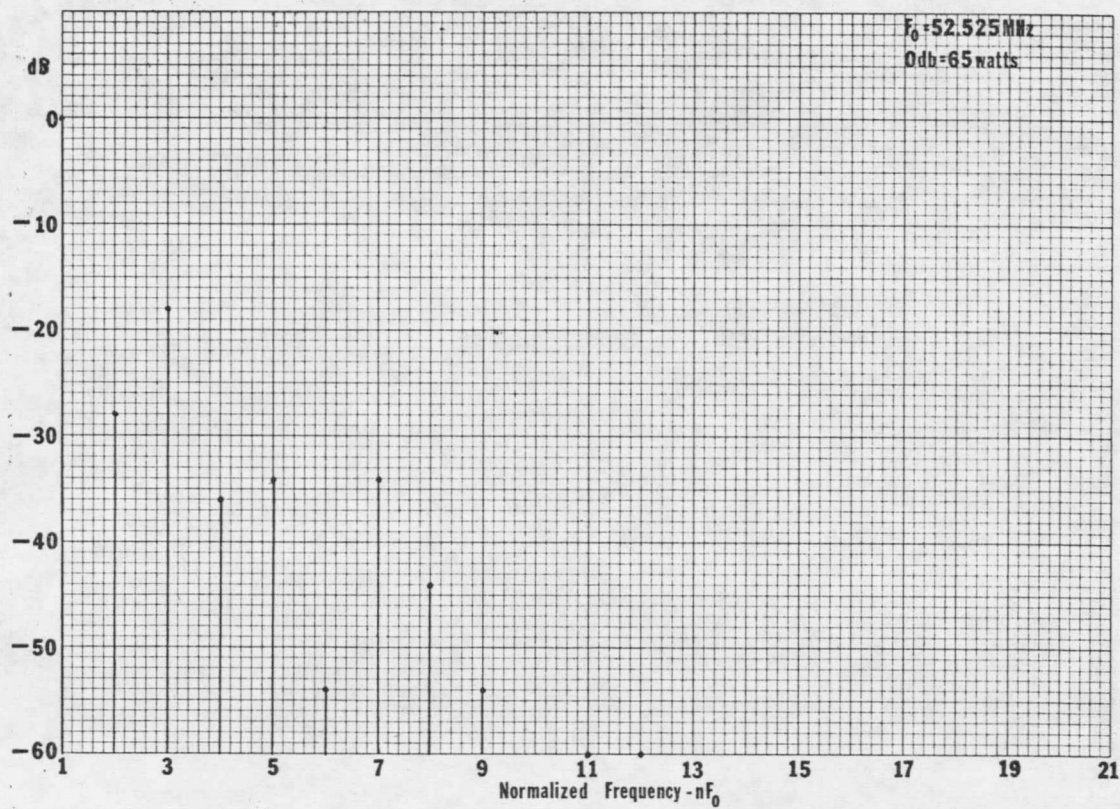


Figure 19
Harmonics Measured at the Output
of the 52 mHz Amplifier

ratios between MSL sections cannot exceed practical size considerations as discussed previously. It was now obvious that the impedance ratio of 66.7 used here was too large for practical construction although the networks performed admirably while they lasted.

The heat sink choice made earlier turned out well. Even after several minutes of transmission at full power, the heat sink became only warm to the touch.

CONCLUSION

This project was undertaken to explore the feasibility of MSL at VHF and to perfect design methods for use by experimenters who do not have expensive lab equipment. This was done through the design of a 52 mHz amplifier.

Overall, the project was a success. It was demonstrated that MSL techniques used in the UHF and microwave regions work equally well at 52 mHz. The main consideration proved to be the physical size of the resulting circuits. The finished amplifier measured 21 cm x 35 cm.

Size may be reduced by bending the MSL sections with properly mitered outside corners as previously referenced or by curving the corners with the optimum curve radius being three MSL main conductor widths (Howe 1974). Mitering corners takes less area than does curving with essentially the same results if proper dimensions are used. Neither type of corner, when properly made, has line characteristics distinguishable from a straight section of line. A further study could be done in this area to compare mitered corners with radial corners.

The amplifier results were good although problems in mitering were discovered which made it necessary to add discrete trimmer capacitors to tune the networks to resonance.

C_3 had to be severely reduced in size in order to reduce reflected power at the input. Apparently the assumption of $1 + j.25$ Ohms for the transistor base input impedance was incorrect. The assumed transistor collector output impedance of $1 - j.25$ Ohms seemed correct, however, as no reflected power was measured at the amplifier output.

The MSL methods used to design the matching networks for the amplifier were based on equations derived from the lossless coaxial transmission line model. It was found that the lossless coaxial transmission line model approximates MSL characteristics quite well. Once the characteristic impedance and phase velocity for a given width of main conductor in an MSL section was known, series inductances and shunt capacitances were readily synthesized. Cascading of these MSL sections may be used to synthesize networks.

The characteristic impedance and phase velocity data was measured by methods described earlier. Once the data was accumulated, it was tested for validity by constructing and testing a low pass "Tee" filter out of MSL sections. Ignoring mitering and length problems in the filter itself, the MSL data proved accurate. A summary of the MSL design equations and a graph of the experimental data taken for phase velocity and characteristic impedance is given in Appendix II. The data given is for G-10 epoxy-glass printed circuit board of 1 mm thickness.

These equations and the data are given for reference by the experimenter. Although a simple method for finding phase velocity with simple apparatus was given, characteristic impedance is a much harder parameter to measure.

Fringing may not be neglected in MSL designs. The narrower the main conductor, the more fringing effects are noticed. For this reason, the characteristic impedance of a line section may not be easily calculated. The quickest method is to measure these impedances. If a vector impedance meter is not available, a reactance bridge could be used. The characteristic impedance is the geometric mean of the open and shorted impedances of the line. Care must be taken to keep leads short. Also, the frequency of such measurements must be several megaHertz away from quarter-wavelength measurements. A suggested further study would be to find an accurate modified equation for characteristic impedance which compensates for fringing.

An additional observation made concerns the actual cascading of MSL sections to synthesize networks. The conclusion here was that the characteristic impedance ratio between cascaded MSL sections must be kept as high as possible for proper inductive or capacitive operation of the line sections. The greater this ratio, the less transmission line type voltage rippling occurs in the

network over a given frequency range. The point was well demonstrated by the MSL "Tee" filter results and comparisons. In this filter, the impedance ratio between MSL sections was 3 and performance in the stop-band was not very impressive.

There is a practical limit to this impedance ratio, however. This was demonstrated in the collector output circuit of the amplifier when the collector inductor (L_2) melted due to its physical size being too small to handle the output current. The characteristic impedance ratio between the MSL inductor and capacitor sections was 66.7. From these findings, a good estimated value for the impedance ratio of cascaded sections would be 10. As a rule, if larger ratios may be used without hindrance of circuit performance, the network will perform better.

Finally, MSL is not readily affected by the proximity of metallic objects. An aluminum sheet was used and placed in the immediate area of the networks for the 52 mHz amplifier. No effects were noticed in performance unless the plate was touching one of the network components.

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APPENDICES

Appendix I

This section presents heat sink calculations made during the amplifier design stage.

Using the power model from Alley and Atwood (1973), the heat sink size was determined. This model is shown in Figure 20.

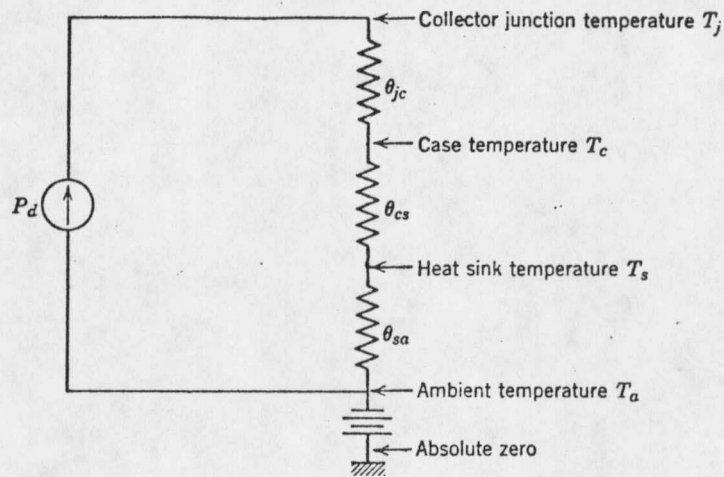


Figure 20

Heat Sink Power Model

By rearrangement of the equations given by Alley and Atwood, the heat sink thermal resistance was given by:

$$\theta_{sa} = \frac{T_{\max} - T_{\text{amb}}}{P_d} - \theta_{jc} - \theta_{cs} \quad (^\circ\text{C/watt}) \quad (23)$$

where: θ_{sa} = Thermal resistance of the heat sink ($^\circ\text{C/watt}$)

θ_{jc} = Thermal resistance of the transistor-junction to case ($^\circ\text{C/watt}$)

θ_{cs} = Thermal resistance between the transistor case and heat sink ($^{\circ}\text{C}/\text{watt}$)

T_{max} = Maximum temperature of the transistor junction ($^{\circ}\text{C}$)

T_{amb} = Environmental ambient temperature ($^{\circ}\text{C}$)

P_d = The maximum allowed transistor power dissipation (watts)

Figure 10 in the text shows the maximum junction temperature (T_{max}) as 200°C . Ambient temperature was chosen as room temperature or 25°C .

Figure 10 rates the power output of the transistor at 70 watts. Though a class C amplifier should operate in the range of 75 percent efficiency, for conservative reasons the efficiency was assumed to be 50 percent. As such, 70 watts would be dissipated (P_d) by the transistor at 70 watts output from the amplifier.

The junction to case thermal resistance (θ_{jc}) was given as $.8^{\circ}\text{C}/\text{watt}$ in Figure 10. The thermal resistance between the heat sink and transistor case (θ_{cs}) was that of the silicon heat sink compound used between them. It was estimated at $.1^{\circ}\text{C}/\text{watt}$.

By substituting the above data into equation (23), the heat sink thermal resistance was calculated:

$$\theta_{sa} = \frac{T_{\text{max}} - T_{\text{amb}}}{P_d} - \theta_{jc} - \theta_{cs}$$

$$\begin{aligned} &= \frac{200^{\circ}\text{C} - 25^{\circ}\text{C}}{P_d} - .8^{\circ}\text{C/watt} - .1^{\circ}\text{C/watt} \\ &= 1.6^{\circ}\text{C/watt} \end{aligned}$$

The only heat sink data available at the time was that shown in Figure 21 (Alley and Atwood 1973). The line shown is for 3.2 mm (.125 inch) thick bright aluminum. If this were used, .45 square meter (700 square inches) of aluminum would have to be used.

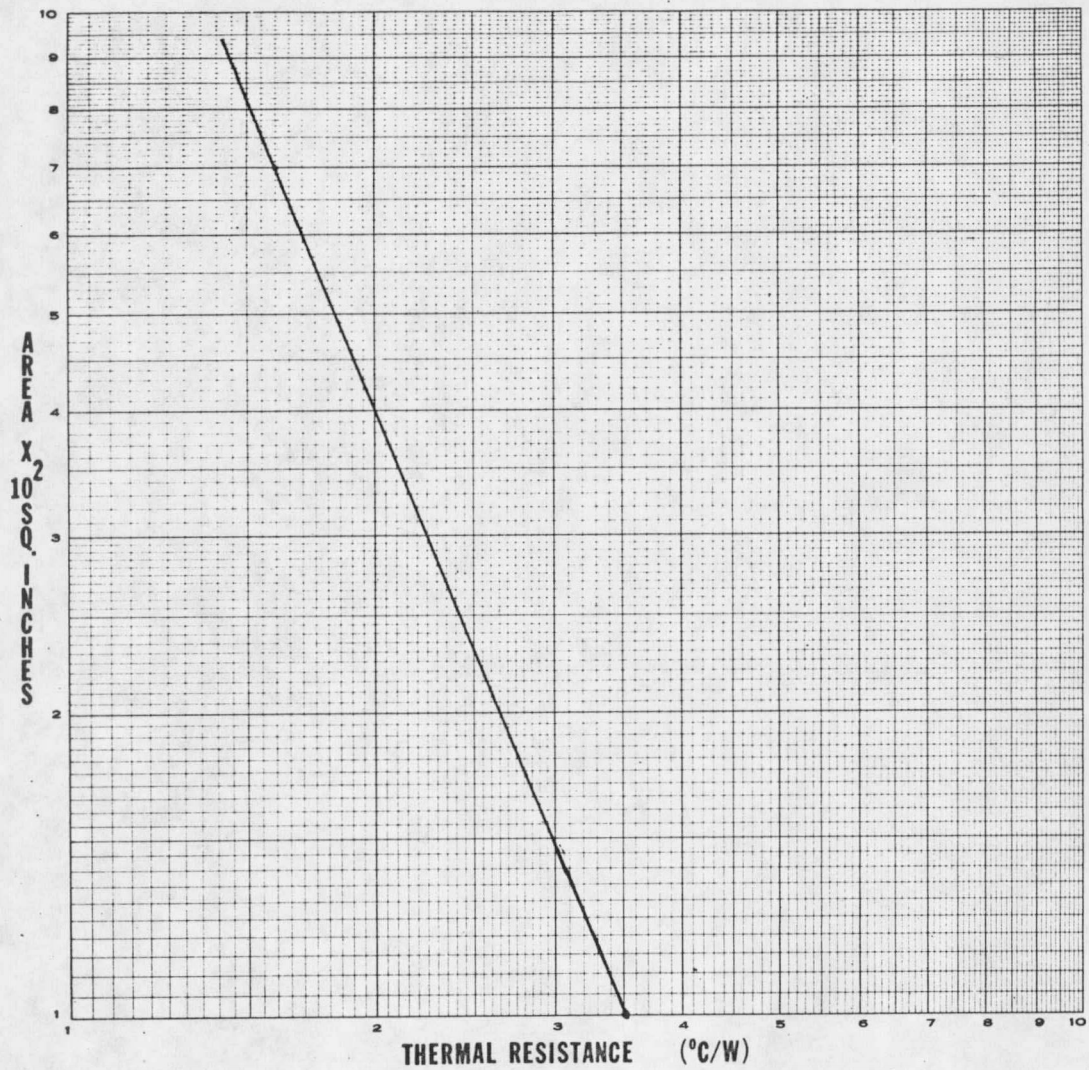


Figure 21

Heat Sink Thermal Resistance and Area Graph

Appendix II

This section contains a summary of MSL design equations and a graph of experimental data measured using G-10 epoxy-glass printed circuit board having a thickness of 1 mm.

The inductance per unit length of an MSL section is given by:

$$L = \frac{Z_0}{V} \quad (\text{Henries/meter})$$

where: L = Inductance per unit length (Henries/meter)

Z_0 = MSL section characteristic impedance (Ohms)

V = MSL section phase velocity (meters/second)

The capacitance per unit length of an MSL section is given by:

$$C = \frac{1}{Z_0 V} \quad (\text{Farads/meter})$$

where: C = Capacitance per unit length (Farads/meter)

Z_0 = MSL section characteristic impedance (Ohms)

V = MSL section phase velocity (meters/second)

To find the MSL section length for a given component value, simply divide the capacitor values by C and the inductor values by L . The resulting units will be length in both cases.

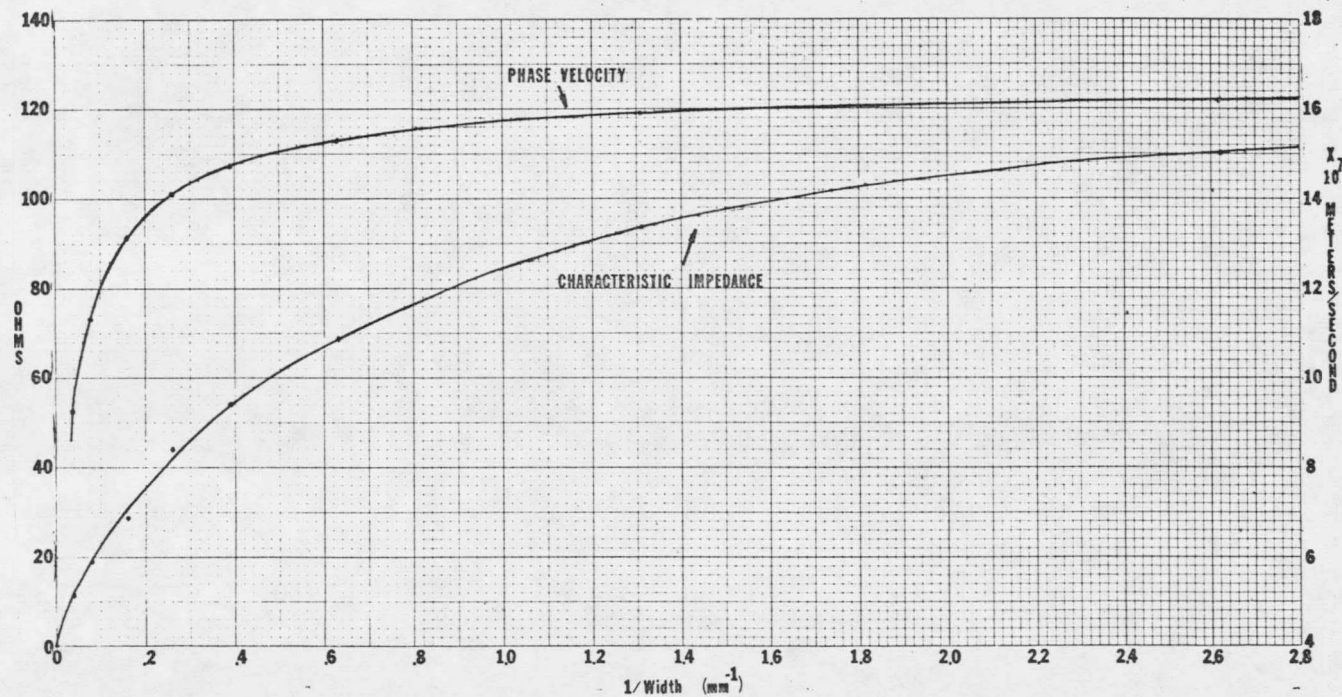
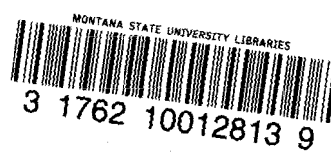


Figure 22

Experimental Phase Velocity and Characteristic Impedance Data
 Plotted for the Reciprocal of Various MSL Section Widths
 Using the G-10 Epoxy-glass Printed Circuit Board
 of 1 mm Thickness




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