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FINAL REPORT
DESIGN AND FABRICATION
OF
COMPACT FOUR-PORT TRIPLEXING FILTERS

by
R. J. Wenzel
April 1970

Prepared Under Contract No. NAS 12-2133

by
The Bendix Corporation
Research Laboratories
Southfield, Michigan 48075

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ABSTRACT

A series of four-port triplexing filters for broadband harmonic doublers is described. The filters are designed to achieve specified performance in a structure of minimum complexity, size, and weight. The measured performance for four complete multiplexer units covering L-band through K_u -band is given.

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1. INTRODUCTION

This report describes the performance achieved in a series of four-port triplexing filters constructed for use in broadband harmonic doublers operating from L-band through K_u-band. These filters were designed to provide precision characteristics with a "best" combination of minimum complexity, size, and loss.

This report will concentrate on the overall design philosophy and presentation of the measured characteristics of the experimental triplexing networks. A detailed manual for the design, construction, and alignment of these filters has been proposed as part of an extension of the present contract.¹

2. DESIGN APPROACH

As design goals, it was desired that the four-port triplexing filters have the following ideal characteristics:

- (1) A common input port (1) matched over a nominal 50 percent frequency band about ω_0 and $2\omega_0$, with an input impedance whose real part is ≥ 30 ohms for all frequencies below the upper frequency edge of the $2\omega_0$ band (below $2.5 \omega_0$ for 50 percent bandwidth). The network is constrained such that little or no power is to be dissipated in the $3\omega_0$ and $4\omega_0$ bands, requiring the input impedance at port (1) to be reactive above $2.5 \omega_0$.

- (2) Low-loss transmission from port (1) to port (2) in the ω_0 band, with a minimum of 25-dB isolation for the $2\omega_0$ band and 15-dB isolation for the $3\omega_0$ and $4\omega_0$ bands.
- (3) Low-loss transmission from port (1) to port (3) in the $2\omega_0$ band, with a minimum of 25-dB isolation for the ω_0 band, and 15-dB isolation (where possible) for the $3\omega_0$ and $4\omega_0$ bands.
- (4) A fourth port (4) to provide a termination for all input frequencies at port (1) below $2.5 \omega_0$ and not contained in the ω_0 or $2\omega_0$ bands, such that isolation to ports (2) and (3) is a minimum of 10 dB.
- (5) ω_0 to be 0.938, 1.875, 3.75, and 7.50 GHz for each respective filter network.

The following sections present general guidelines for selecting the necessary filter to meet the requirements stated above, with the added goals of achieving minimum size, insertion loss, and complexity. The intended application of the filter structures in harmonic multipliers for space communications places a premium on achieving low insertion loss and small size beyond that usually encountered.

2.1 Tradeoffs Between Insertion Loss and Isolation in Broadband Contiguous-Band Diplexers

From past experience in the construction of microwave multiplexing filters, it has been found that the insertion loss for compact broadband TEM microwave structures in the passband frequency region near

band edge is approximately 0.1 dB per section, with lower achievable values near band center.* If the component filters in a bandpass-bandstop contiguous-band diplexer have monotonic stopband attenuation characteristics (as is often the case), then minimum isolation occurs at band edge, where the passband insertion loss is approximately 0.1 dB per section. If the diplexer is well matched, a simple calculation shows that a minimum isolation of approximately 17 dB will add 0.1 dB to the corresponding passband loss. Consequently, the decrease in passband loss achieved by increasing the isolation beyond 17 dB by use of additional filter sections is normally matched or exceeded by the increased passband dissipation caused by the additional filter elements. The advantage of increasing the complexity of filter structures of the type being considered, to obtain isolation values greater than approximately 17 dB, is the increased isolation achieved. The price paid is an increase in size, complexity, and passband insertion loss. It is emphasized that the 0.1-dB loss per section and the corresponding 17-dB isolation are representative of typical compact wideband TEM filter structures and are dictated by the achievable element Q . If large, high Q , cavity-type elements were used, the dB loss per section value would be lower, and the tradeoff isolation value correspondingly higher.

* For low-frequency filters, loss is typically somewhat less than 0.1 dB per section. However, the 0.1 dB value has been found to be a good estimate for C-band and higher frequencies.

In summary, increasing isolation beyond 17 dB in a harmonic multiplier filter, using typical compact TEM filter structures, will normally result in decreased or unchanged efficiency near band edge due to the increase in passband dissipation. If a large isolation requirement must be met, a larger, more complex and costly structure of lower efficiency must be accepted. A minimal reduction in efficiency with increased isolation can be achieved by use of filter structures employing high-Q elements. These structures are typically very large, creating serious junction effects. It is frequently not possible to compensate these effects when broadband performance is required.

2.2 Tradeoffs Between Bandwidth and Physical Form

Physical realizations of commensurate linelength microwave filters could be divided into two basic categories: (1) stub-type realizations and (2) coupled-line realizations. In addition to obvious physical differences, the type of realization to be employed in a given filter application is often dictated by the specified filter bandwidth. The reason for discrimination based on bandwidth is that the achievable range of characteristic impedance values is limited. These values normally lie in the range of 5 ohms to 200 ohms with 15 ohms to 150 ohms being the maximum range usually desired.

Stub-type filter realizations normally require impedance values in the 15-ohm to 150-ohm range for filter bandwidths on the order of 2:1 (67 percent) to 5:1 (133 percent). On the other hand, coupled-line realizations are practical for bandwidths of a fraction of a percent

to 3:1 (100 percent), or so. Therefore, for bandwidths of less than 5:1, there is a division of practical realizability between stub-type and coupled-line structures that occurs in the vicinity of 2:1 bandwidth. Structures narrower than octave band normally employ coupled-line realizations and those greater than octave band normally employ stub-type realizations. A mixture of both types of realizations is used in the transition region.

Stub-type realizations are often very compact because multiple stub geometries, employing both series and shunt elements, are practical. On the other hand, the narrowband capability of coupled-line bandpass structures is achieved by the use of redundant or superfluous network elements which do not contribute to the response, but act primarily as impedance transformers to provide practical line impedance values. The interdigital filter form is an example of a typical structure that provides a compact realization for narrowband-bandpass filters. Most coupled-line geometries for narrowband bandstop filters require a cascade of transmission line sections or unit elements, each one-quarter wave long at the center frequency of the bandstop region. The overall size of most coupled-line realizations of bandpass-bandstop contiguous-band diplexers, particularly at low microwave frequencies, is considerably larger than the overall size of a stub-type realization.

This section will conclude with examples applicable to the type of filters required in the harmonic multiplier network. The practical limitations previously discussed will be demonstrated.

To demonstrate practical realizability restrictions on stub-type contiguous-band diplexing networks, consider the three-section prototype network shown in Figure 1(a). The element values are those of 0.5-dB ripple $|Y_{12}|^2$ designs which, when properly scaled, provide a pseudo-complementary response with a maximum input VSWR given to a good approximation^{2,3} by

$$\text{VSWR}_M \approx 10^{\frac{\alpha}{10}} = 1.12 \quad , \quad (1)$$

where

$\alpha = 0.50$ dB, ripple value in this example.

The scaled element values⁴ for 100 percent and 50 percent bandwidths are given in Figures 1(b) and 1(c), respectively. The element values have been scaled such that the specified bandwidth occurs between the bandpass filter cutoff frequencies rather than the 3-dB crossover frequencies. A possible (and compact) stub-type realization of the above prototype network is shown in Figure 1(d), where the correlation between prototype element values and line impedances is indicated in the figure. The impedance values for the 100 percent and 50 percent bandwidth cases for a 50-ohm termination level are shown in a plan view in Figures 1(e) and 1(f), respectively, where the use of teflon dielectric ($\sqrt{\epsilon_r} = 1.42$) was assumed for support and foreshortening of the series stub elements.

As is evident from Figures 1(e) and 1(f), all impedance values in the 100 percent bandwidth design lie in the desired 15-ohm to 150-ohm range, whereas the 229-ohm series stub in the 50 percent bandwidth design

is not practical. As is also clear from the two designs, the realizability of this particular filter prototype in stub form (and, in general, for all similar prototypes) is dependent on the realizability of the first series stub in either filter. The optimum situation occurs when these stub impedances are equal (giving minimum stub impedance), a condition which obviously occurs for a bandwidth between the 100 percent and 50 percent values chosen. Finally, the minimum bandwidth that can be achieved with practical structures lies between 100 percent and 50 percent bandwidth.

Next, assume that a coupled-line realization of a three-section pseudo-complementary filter pair is desired to achieve a practical structure with 50 percent bandwidth. A compact coupled-line configuration can be constructed using an interdigital bandpass filter and a stub-type filter as shown in Figure 2(a). Approximate design procedures for this type of diplexer applicable to narrow or moderate bandwidth designs are given in Matthaei and Cristal,⁵ and broader-band exact designs can be obtained using the procedures described by Wenzel.^{3,6} As has been shown,⁷ and L-C ladder is not an exact prototype network for an interdigital filter, and for a comparable number of filter sections, the performance of the interdigital is somewhat poorer than that achievable with a ladder prototype. However, for the 50 percent bandwidth design under consideration, the difference in performance is not large. Note also in Figure 2(a) the introduction of two redundant-line elements (resulting in a five-line structure to realize a three-section response) in the interdigital realization to obtain practical impedance levels, and the introduction of

redundant-line elements in the bandstop network to allow spacing of the shunt stubs. Optimum filter theory⁸ might be employed to allow the redundant-line elements in the bandstop filter to contribute to the response. However, the advantage gained by including contributing unit elements in a 50 percent bandstop design is not substantial. The bandstop filter might also employ size reducing configurations as shown in Figures 2(b) and 2(c). However, a direct comparison of the stub-type realization (neglecting physical realizability for the moment) of Figure 2(d) with that of Figures 2(a) through 2(c) shows the size advantages of the stub-type network. For example, even the most compact structure employing an interdigital filter would be at least 6 inches long at 1.0 GHz ($\lambda/4 \approx 3.0$ inches), whereas the stub realization would be but 3 inches long. Finally, the stub realization, when practical-line impedances can be achieved, can be less costly to design, construct, and align than the coupled-line counterpart and it offers performance that theoretically exceeds (only slightly for the case given) that of the comparable coupled-line structure for a given number of filter elements.

2.3 Estimate of Filter Complexity

As an estimate of required component filter complexity, the number of sections required to achieve 25-dB isolation between 50 percent bandwidth channels using commensurate linelength pseudo-complementary filters with center frequencies of ω_0 and $2\omega_0$ was computed. In the calculations, $|Y_{12}|^2$ designs with 0.5-dB ripple were used which give a maximum input VSWR of approximately 1.12 and a theoretical power transmission

ripple of less than 0.02 dB. The 50 percent bandwidth region was chosen between the ripple bandwidth instead of the 3-dB crossover bandwidth to reduce filter complexity as described below. The computed number of sections was 3.03 for the ω_0 filter-pair and 4.38 for the $2\omega_0$ filter-pair. The use of three sections for the ω_0 design and four sections for the $2\omega_0$ results in isolation over a 50 percent band of approximately 25 dB and 21.7 dB, respectively. The reason for the increased complexity of the $2\omega_0$ filter relative to the ω_0 filter for identical percentage bandwidths is that the ω_0 filter has poles of attenuation at $2\omega_0$ whereas the $2\omega_0$ filter has its poles of attenuation at dc and $4\omega_0$.

Another factor, not considered in the above calculations, is that the complexity of the filters required to achieve a given isolation is dependent on the order in which the filter-pairs are cascaded. This is readily demonstrated as follows. Since most of the power incident at the first filter-pair in the design bandpass region is routed through the bandpass filter, the energy going through the bandstop complement in the removed band is already attenuated. The second filter in the cascade then needs to supply only an additional amount of isolation beyond that provided by the first filter-pair to meet the required specifications, thus reducing the complexity of the second filter-pair. In the above case, if the ω_0 band is removed first with a three-section filter, the $2\omega_0$ filter can be reduced to a three-section design to provide 20-dB to 25-dB isolation in both bands. The above reduction in complexity is dependent on designing the component filters with the specified bandwidth

between the prototype ripple values rather than the 3-dB values. If a 3-dB bandwidth is used, the maximum attenuation provided by the first filter-pair at band edge is 3 dB, which is not sufficient to allow reduction of complexity in the second filter-pair. When the ripple bandwidth is used, an attenuation determined by the ripple value is achieved, which in the above case is approximately 10 dB, allowing the elimination of one section in the second filter-pair.

In view of the tradeoff between isolation and dissipation loss discussed in Section 2.1, the use of a minimum complexity filter structure that provides greater than 17-dB isolation over a specified band will result in a more efficient multiplier than can be obtained with more complex filters. Furthermore, size, weight, and cost will be reduced by using the minimum complexity structure consistent with minimum passband loss.

Based on the use of a cascade of a three-section ω_0 design with a three-section $2\omega_0$ design, it is possible to compute the frequency band between $1.25 \omega_0$ and $1.50 \omega_0$ in which isolation to the ω_0 and $2\omega_0$ ports is simultaneously 10 dB or greater. (See characteristic (4), page 2.) Direct calculation shows that simultaneous achievement of 10-dB isolation to both ports is not achieved at any frequency within the $1.25 \omega_0$ to $1.50 \omega_0$ band. Therefore, for minimum complexity networks that essentially satisfy the 25-dB isolation requirement and provide minimum passband loss, the 10-dB isolation requirement cannot be simultaneously achieved at any frequency in the $1.25 \omega_0$ to $1.50 \omega_0$ band. 10-dB isolation over parts of

the band can be achieved by using more complex filter designs at the cost of increased size, weight, unit cost, and an increase in passband dissipation near band edge of approximately 0.1 dB per added filter section. Network complexity would have to be doubled to terminate a significant portion of this band at port (4). It appears that a reasonable compromise would be to make the ω_0 and $2\omega_0$ passbands contiguous and eliminate the 10-dB requirement between $1.25 \omega_0$ and $1.50 \omega_0$. The above possibility will be explored in greater detail in a subsequent network configuration.

2.4 Spurious Passbands

Due to the periodic frequency response of commensurate line-length networks, i.e., passbands at ω_0 , $3\omega_0$, etc., overall network requirements cannot be achieved by a cascade of two complementary or pseudo-complementary filter-pairs. Referring to Figure 3, a cascade of 50 percent bandwidth ω_0 and $2\omega_0$ filter-pairs is shown in which significant repeating passband regions are designated. The bands shown are those for ideal infinite complexity filters. Practical structures will have significant overlap due to finite crossover characteristics. Elimination of spurious passbands requires the addition of a $2.5 \omega_0$ lowpass filter at port (1) or additional lowpass filtering at ports (2) and (4). It is sometimes simpler to add two lowpass structures at the output ports rather than to use a $2.5 \omega_0$ lowpass filter at the input port due to reduced selectivity and size requirements of the former configuration. The use of two lowpass filters can also result in lower dissipation loss in the $2\omega_0$ band.

2.5 Possible Network Interconnection

Many possible interconnections were investigated for achieving the desired network characteristics including various arrangements of bandpass-bandstop complementary filters, highpass-lowpass complementary filters, and common junction arrangements containing bandpass and lowpass structures. All required the use of additional filtering to suppress repeating passbands and resulted in similar network complexity. After serious consideration, the use of coupled-line bandpass-bandstop filter-pairs was ruled out for the following reasons:

- (1) Size and Weight - As demonstrated in Section 2.2, coupled-line configurations are large in size, particularly at low microwave frequencies. For example, two three-section filter-pairs at 1.0 and 2.0 GHz would lead to minimum length structures of 6.0 and 3.0 inches, respectively. If more conventional bandstop designs were used, see Figure 2(a), very large structures would result. The network size for coupled-line geometries was considered prohibitive for the low-frequency designs.
- (2) Cost and Complexity - The design, construction, and alignment of suitable coupled-line structures could be costly due to the network complexity and resulting junction effects.
- (3) Performance - Based on the tradeoff studies discussed above, no clear cut performance advantage for the use of coupled-line structures over the network described in the following section could be found.

3. PREFERRED FILTER NETWORK CONFIGURATION

As concluded in the previous section, significant advantages in size, weight, and cost can be achieved in the desired four-port network if a stub-type realization rather than a coupled-line network is employed. The preferred network described below utilizes a stub-type realization made possible by a careful choice of filter bandwidth. It was felt that this network offered the best overall solution to the specified filter and multiplier problem.

3.1 Basic Design Considerations

Due to the approximate 0.1-dB loss per section near band edge for filter elements of the type required, increasing isolation beyond 17 dB at band edge will result in an increase in passband loss and a lowering of multiplier efficiency as well as an obvious increase in size, weight, and cost. Consequently, filter complexity was limited to that required to achieve a minimum of 17-dB isolation throughout a 50 percent band about ω_0 and $2\omega_0$. Based on the selectivity calculations described in Section 2, the use of component filters with a minimum of three sections was indicated. Three-section filters do not allow significant isolation from port (1) to ports (2) and (3), in the $1.25 \omega_0$ to $1.50 \omega_0$ band. It was, therefore, decided to delete this characteristic to avoid a significant increase in the complexity of the component filters. This deletion allows the choice of a crossover frequency lying anywhere within the $1.25 \omega_0$ to $1.50 \omega_0$ band. Finally, since a commensurate linelength band-pass filter centered at ω_0 has poles of attenuation at dc and $2\omega_0$, the

heart of the multiplexing filter network was chosen to be a three-section pseudo-complementary design centered at ω_0 with simpler filter structures being required to suppress repeating passband responses and duplex remaining widely separated frequency bands.

Since a three-section pseudo-complementary diplexer with 50 percent bandwidth (measured between the ripple values) was shown in Section 2 to be impractical to construct in stub form, the bandwidth had to be increased to obtain a realizable structure. An appropriate bandwidth or cutoff frequency was found by forcing the 25-dB isolation bandwidth about ω_0 to equal the 25-dB isolation bandwidth about $2\omega_0$ for a three-section pseudo-complementary filter pair centered at ω_0 . Calculations show that when the bandwidth between the ripple values of the band-pass filter was near octave, the 25-dB isolation bandwidth was approximately 40 percent and the 17-dB isolation bandwidth was approximately 50 percent for both ω_0 and $2\omega_0$ bands. The above design is seen to offer minimum passband loss with practical element Q's, and to give the least degradation in multiplier efficiency due to an optimum tradeoff between filter dissipation and isolation leakage.

3.2 Final Filter Network

The block diagram for the four-port filter with all important spurious passbands is shown in Figure 4(a), and pertinent filter design parameters are shown in Figure 4(b). Other network connections for suppressing the spurious responses are possible. However, the configuration shown was the one used and will be the only configuration discussed.

Network operation is as follows: The primary passbands at ω_0 and $2\omega_0$ are separated with minimum loss in a three-section stub-type diplexer. The widely separated dc and $2\omega_0$ bands at the output of the bandstop portion of the first diplexer are then separated in another three-section pseudo-complementary diplexer. The bandwidth and complexity of this diplexer were chosen to provide required isolation and to minimize loss in the $2\omega_0$ band. The wide separation between the dc and $2\omega_0$ bands is significant in allowing the achievement of this result. Suppression of transmission at frequencies in the $3\omega_0$ and $4\omega_0$ bands is accomplished by two simple lowpass filters as shown in Figure 4. In all cases, filter complexity was chosen to provide minimum size and passband loss consistent with specified performance. The equivalent circuit of the complete multiplexer is shown in Figure 5(a), and a pictorial drawing of a practical physical realization is shown in Figure 5(b).

3.3 Expected Performance of the Final Network

Plots of theoretical performance for the network of Figure 4 and the ideal specifications of Section 2 are shown in Figure 6. Actual characteristics are plotted wherever they do not exceed the given specifications. Line impedances for all filter structures were investigated and found to be practical to construct.

Based on the above design, original estimates of realizable network performance are given below.

Passband VSWR at port (1) in the

50 percent bandwidth ω_0 and $2\omega_0$

bands:

$\leq 1.50:1$

Maximum VSWR at port (1) below

$2.5 \omega_0$ at all frequencies not in

the ω_0 or $2\omega_0$ bands: $\underline{\leq 2.00:1}$

Passband loss in the ω_0 and $2\omega_0$

bands: 0.3 - 0.5 dB maximum

Isolation: Isolation values should be very close (with a few dB) to those given in Figure 5. Isolation is not a particularly critical or sensitive parameter.

These estimates are included to indicate the close correlation between original expectations and the measured performance of the experimental filters described in the following section.

4. MEASURED PERFORMANCE

4.1 General Comments on Design and Assembly

Four complete filter units were constructed using the four ω_0 design frequencies. The component filters of each succeeding unit were scaled by a factor of two in length and by approximately $1/\sqrt{2}$ in ground plane spacing (gps). The above ground plane scaling results in an increase in the effect of parasitic elements at higher frequencies but tends to give similar passband dissipation loss in all units. Use of a gps scaling factor of two would maintain the same relationship of parasitic elements to desired elements in all units. However, the cross-sectional dimensions would vary by a factor of $2^3 = 8$, and would lead

to impractical physical dimensions as well as a substantial increase in loss at high frequencies.

The lowest frequency unit used a 0.5-inch gps and the highest frequency unit a 0.1875-inch gps. Parasitic effects were minor in all units at frequencies below X-band. They become important at X-band and above in both the third and fourth units. The greatest problems were caused by the use of shunt open-circuited stubs at high frequencies. These did not work well in X-band and were unpredictable in K_u band, especially if the stub impedances were low. Both the series open-circuited and short-circuited stubs worked well throughout all frequency bands, although proper length adjustment was critical at high frequencies. A practical technique used to realize and tune the series stubs is shown in Figure 7.

The most difficult part of obtaining good performance in a pseudo-complementary filter-pair at high frequencies has been found to be the poor performance of shunt open-circuited stubs. Good performance in the 15-GHz diplexer of the fourth unit could only be achieved by using a physical realization for the bandstop filter that consisted primarily of series short-circuited stubs. In all high-frequency filters, it was found necessary to couple all shunt stub elements to vertical side walls and to use reduced rectangular channels for all connecting lines to insure suppression of higher-order modes and spurious resonances.

In all units, each component, filter or diplexer was individually constructed and aligned before final assembly. The first three units provided good performance after the first complete assembly, with no

internal modifications being required. The fourth and highest frequency unit required additional internal walls to suppress higher-order modes. In the construction of all component filters, no change in cross-sectional dimensions was required to obtain acceptable performance. Proper alignment was achieved only by the adjustment of line lengths and by junction compensation.

The cases for all four units were aluminum as were all internal filter elements that did not require solder connections. Soldered parts were machined brass, or were lengths of standard wire in the case of the series stub center conductors. All series stubs were encased in teflon to foreshorten their length. Before beginning construction of the component filters of each unit, separate connector transitions were constructed and measured to insure good performance over the operable band ($\omega \leq 2.5 \omega_0$) of the unit. The measured performance of the four complete units follows.

4.2 Unit No. 1 - $\omega_0 = 0.938$ GHz

The measured responses for all component filters of the first unit are shown in Figures 8 through 10 and the respective filters are shown in Figures 11 through 13. The performance of the individual filters is shown to indicate typical component performance and correlation with complete unit performance. The measured performance of the complete first unit is shown in Figure 14. The isolation response agrees well with the predicted performance. Dissipation loss is less than 0.15 dB in the ω_0 band and is less than 0.2 dB in the $2\omega_0$ band. Maximum passband VSWR is

approximately 1.2:1, with a maximum VSWR of 1.53:1 in the second crossover region. The complete unit is shown in Figure 15. Total unit volume is 12.5 cubic inches, and the unit weighs 14 ounces.

4.3 Unit No. 2 - $\omega_0 = 1.875$ GHz

The measured responses of the second unit component filters are shown in Figures 16 through 18 and the respective filters are shown in Figures 19 through 21. Note the nearly identical response characteristics of the second unit diplexers when compared to the response of the first unit filters. The measured response of the complete second unit is shown in Figure 22. The isolation response is in close agreement with the theoretical performance. Dissipation loss is slightly less than 0.2 dB in the ω_0 band and is about 0.2 dB in the $2\omega_0$ band. These values are better than those originally expected. VSWR is less than 1.30:1 over most of the band. A maximum VSWR of 1.54:1 occurs in the $\omega_0 - 2\omega_0$ crossover region. The interior of the complete unit is shown in Figure 23, and the packaged unit is shown in Figure 24. Total unit volume is 4.4 cubic inches, and the unit weighs 6 ounces.

4.4 Unit No. 3 - $\omega_0 = 3.75$ GHz

The measured responses of the third unit component filters are shown in Figures 25 through 27, and the third unit primary diplexer is shown in Figure 28. (No individual photograph of the third unit secondary diplexer was taken. The third unit lowpass filter is shown in Figure 21.) The measured performance of the complete third unit is shown in Figure 29. The isolation response is excellent and common port

VSWR is less than 1.43:1 at all frequencies below cutoff ($2.5 \omega_0$). Primary passband loss is about 0.3 dB and the second harmonic passband loss is approximately 0.5 dB. Due to a skewing of the response of the $2\omega_0$ diplexer in the 15 GHz region, an additional lowpass filter was required at the $2\omega_0$ port to achieve the specified higher harmonic rejection. This filter contributes approximately 0.1 dB loss to the second harmonic passband. The skewed response is due to the use of shunt open-circuited stubs in the component diplexers. By using a series stub configuration as discussed in Section 4.1, it is felt that this additional lowpass filter could be eliminated. Furthermore, if this filter network is used in a multiplier chain, the band limiting of the input filter of the next higher unit will also eliminate the need for this filter. The interior of the complete unit is shown in Figure 30, and the packaged unit is shown in Figure 31. Total unit volume is 2 cubic inches and the unit weighs 3 ounces.

4.5 Unit No. 4 - $\omega_0 = 7.50$ GHz

The measured responses of the fourth unit component filters are shown in Figures 32 through 34, and the component diplexers are shown in Figures 35 and 36. The measured performance of the complete fourth unit is shown in Figure 37. Passband VSWR is less than 1.5:1 almost everywhere, and the isolation characteristics are as desired. Primary passband loss is typically 0.3 dB with secondary passband loss being 0.5 dB. The complete unit uses short lead-in lines, thus providing a significant reduction in dissipation loss from the sum of the component

loss values. The complete unit weighs 2 ounces and unit volume is 1.1 cubic inches. The interior of the unit is shown in Figure 38, and the completed package is shown in Figure 39.

4.6 Summary

A summary of the performance of all four complete units is shown in Figure 40 for reference. Referring to the original expected performance given in Section 3.3, the final results are seen to meet or exceed these values in almost all respects.

5. RECOMMENDATIONS FOR FUTURE WORK

From discussions with members of the NASA-ERC technical staff, requirements for additional filter networks and detailed design manuals have resulted in submission of a proposal¹ for additional work and extension of the present contract.

6. REFERENCES

1. Bendix Letter CA-27885, Proposed Additional Scope to Contract NAS 12-2133.
2. R. J. Wenzel, "Application of Exact Synthesis Methods to Multichannel Filter Design," IEEE TRANSACTIONS ON MICROWAVE THEORY AND TECHNIQUES, Vol. MTT-13, January, 1965, pp. 5-15.
3. R. J. Wenzel, "Some General Properties of Commensurate Linelength Complementary and Pseudo-Complementary Microwave Filters," IEEE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, Vol. MTT-16, No. 3, March 1968, pp. 191-2.

4. R. J. Wenzel, "Exact Design of TEM Microwave Networks Using Quarter-Wave Lines," IEEE TRANSACTIONS ON MICROWAVE THEORY AND TECHNIQUES, Vol. MTT-12, January, 1964, pp. 94-111.
5. G. L. Matthaei and E. G. Cristal, "Multiplexer Channel-Separating Units Using Interdigital and Parallel-Coupled Filters," IEEE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, Vol. MTT-13, May, 1965, pp. 328-34.
6. R. J. Wenzel, "Printed Circuit Complementary Filters for Narrow Bandwidth Multiplexers," IEEE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, Vol. 16, No. 3, March, 1968, pp. 147-57.
7. R. J. Wenzel, "Exact Theory of Interdigital Band-Pass Filters and Related Coupled Structures," IEEE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, Vol. MTT-13, No. 5, September, 1965, pp. 559-75.
8. M. C. Horton and R. J. Wenzel, "General Theory and Design of Optimum Quarter-Wave TEM Filters," IEEE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, Vol. MTT-13, May, 1965, pp. 316-27.

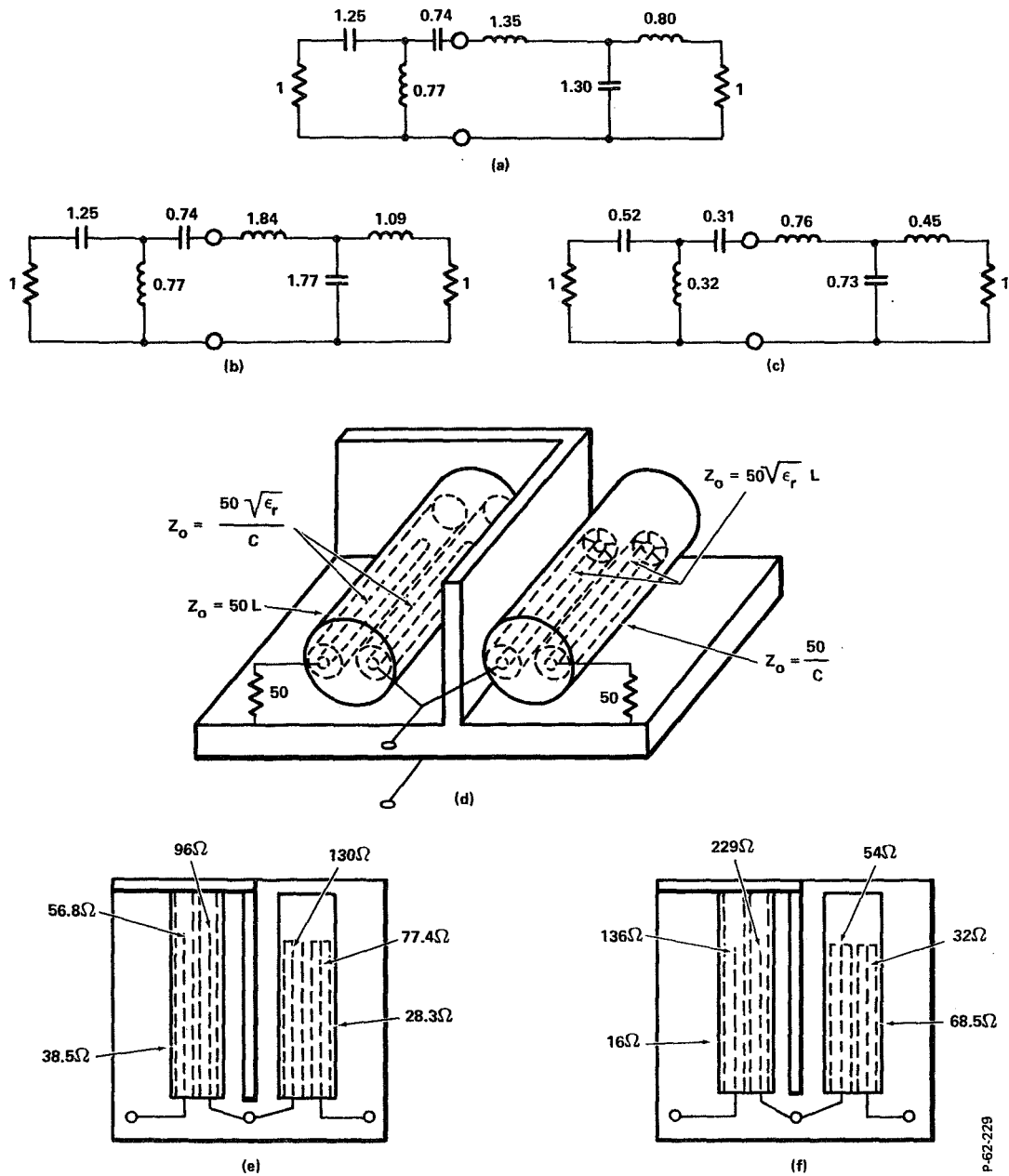
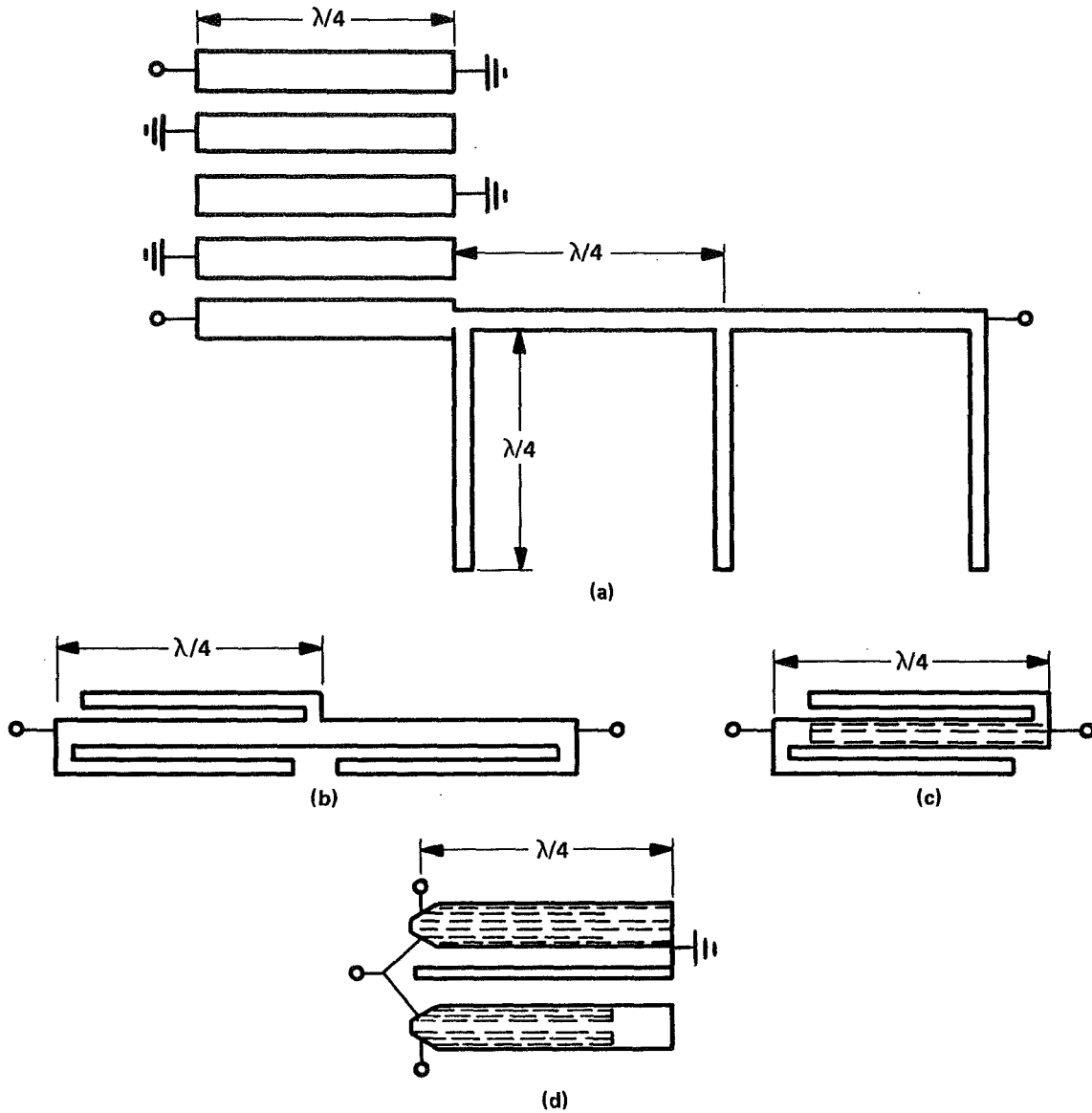


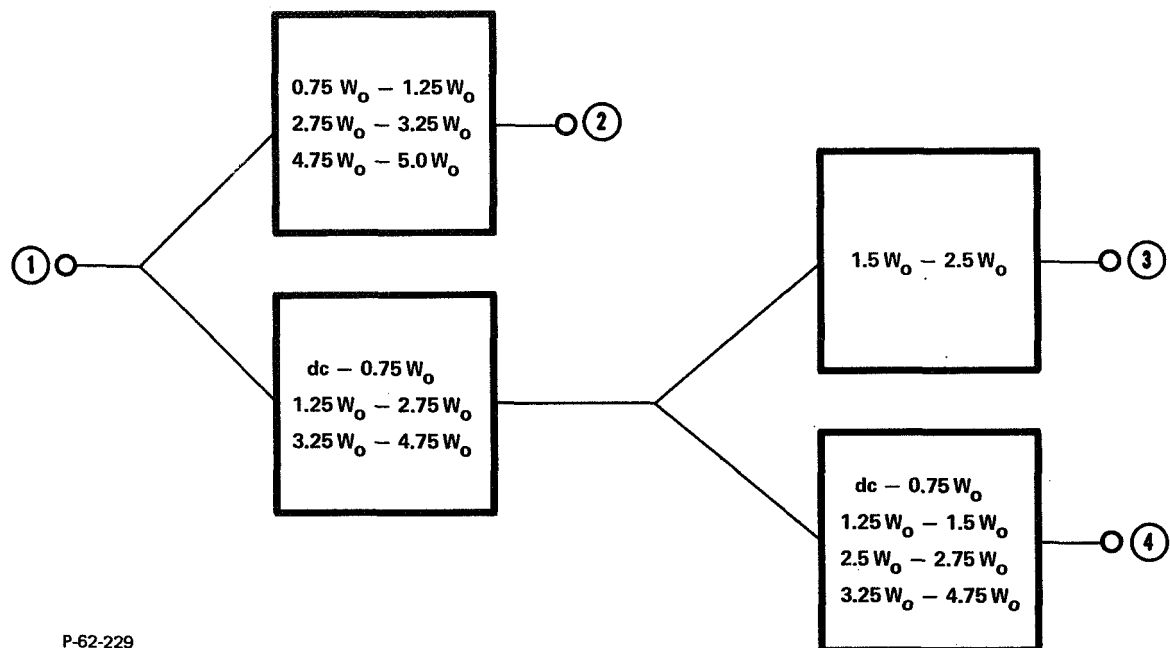
Figure 1 - Stub Type Pseudo-Complementary Diplexer



- (a) Interdigital Bandpass and Shunt Stub Bandstop Realization
- (b) Compact Bandstop Realization Using Folded Stubs
- (c) More Compact Bandstop Realization Using a Series Stub
- (d) Stub Type Realization of the Type Described in Figure 1 for Size Comparison. Theoretical Performance Exceeds that of the Structures Shown in (a)-(c).

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Figure 2 - Coupled Line Realizations for Contiguous Band Diplexers and Comparison with Stub-Type Realization



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Figure 3 - Significant Spurious Passbands for a Commensurate Line-length Complementary Filter Cascade

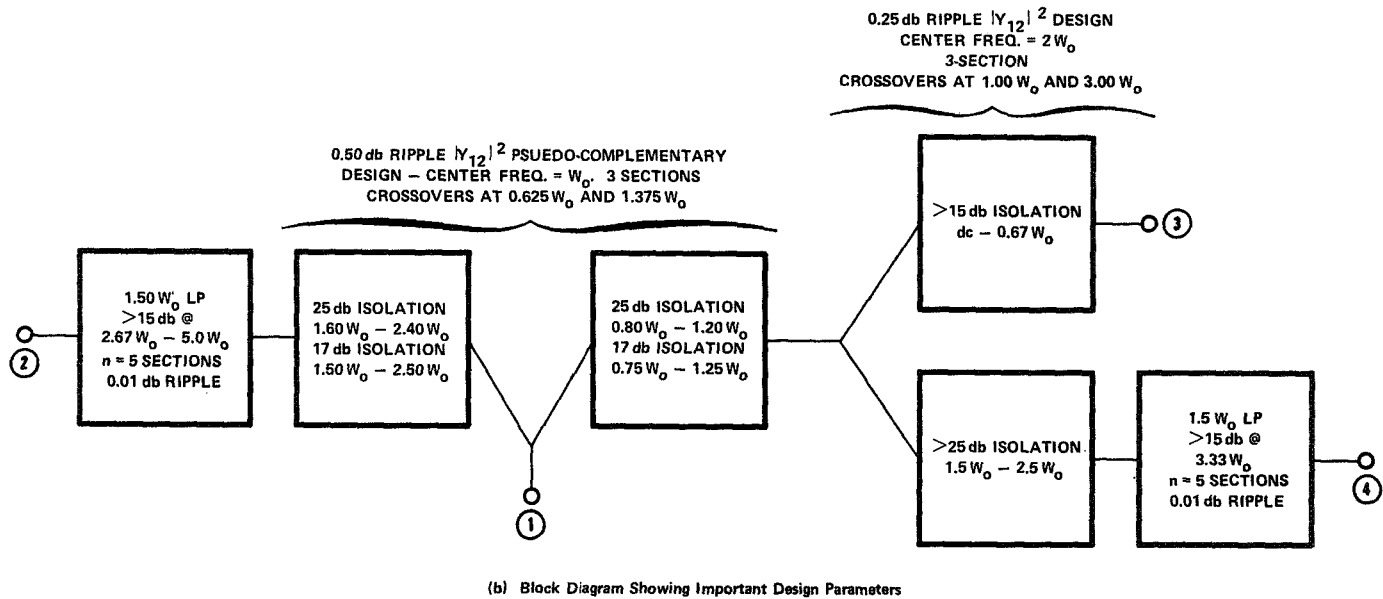
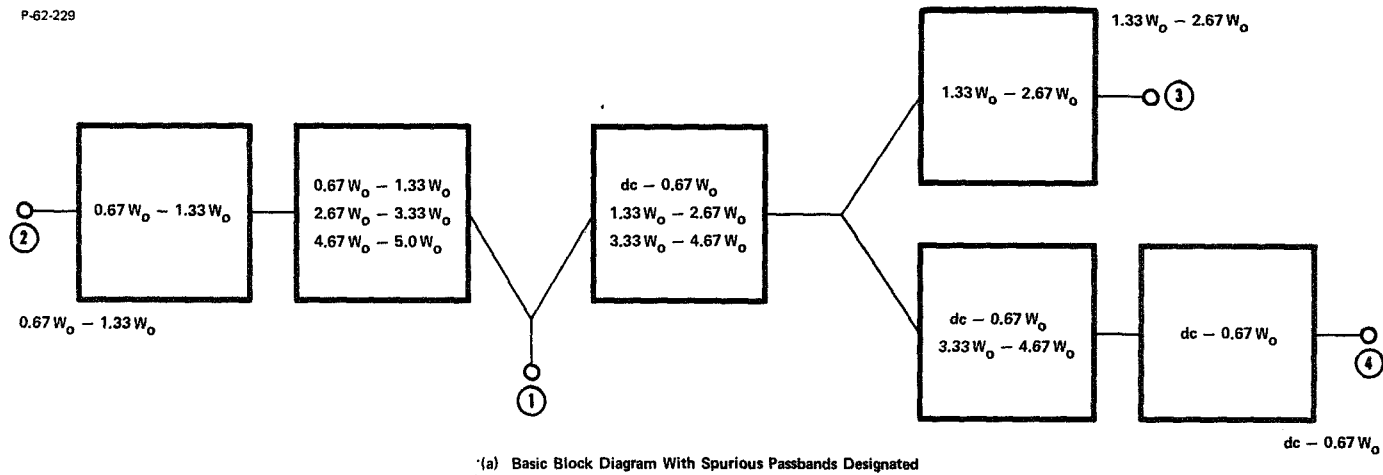


Figure 4 - Block Diagram of Multiplexing Network

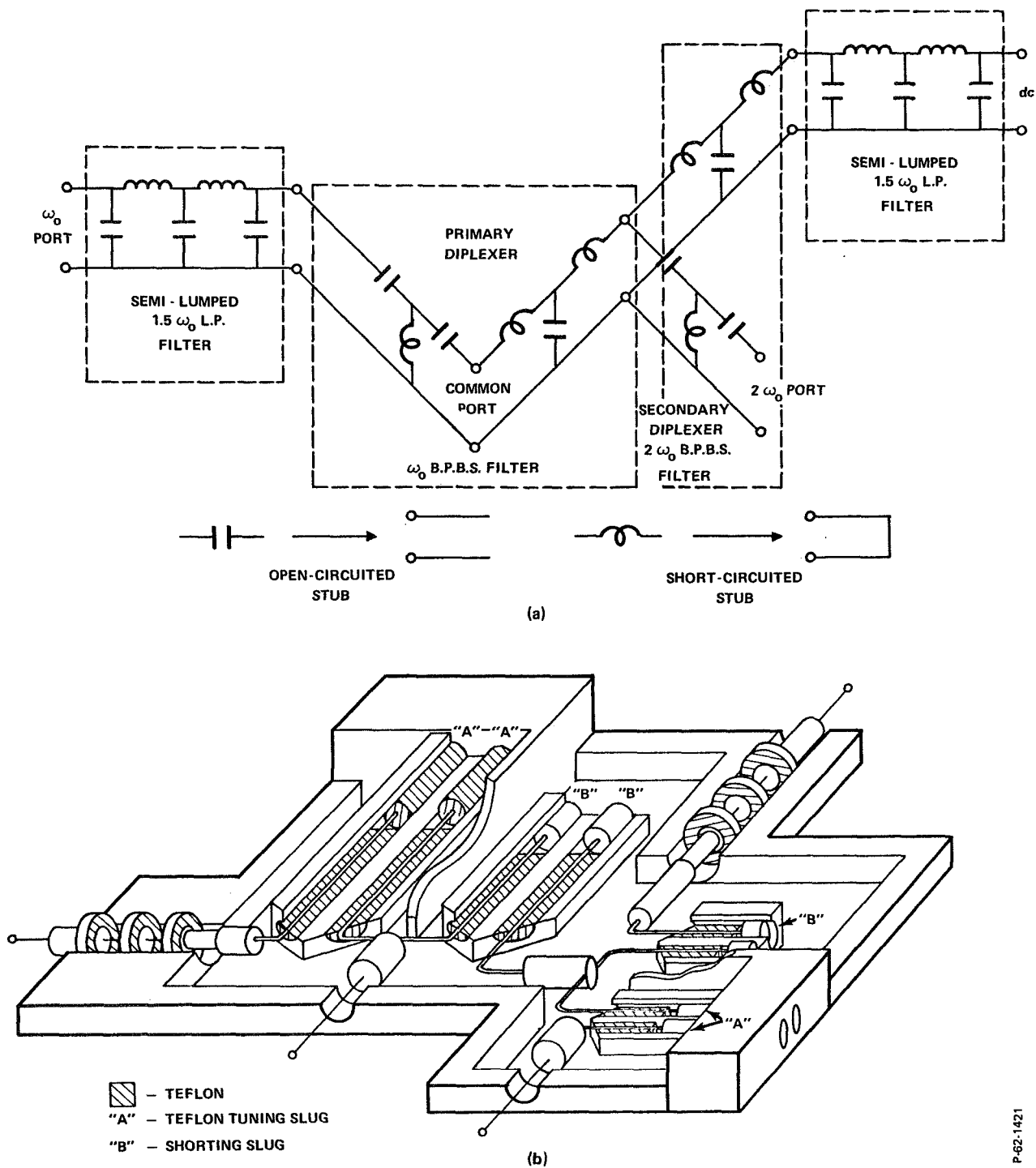


Figure 5 - Equivalent Circuit and Practical Physical Realization

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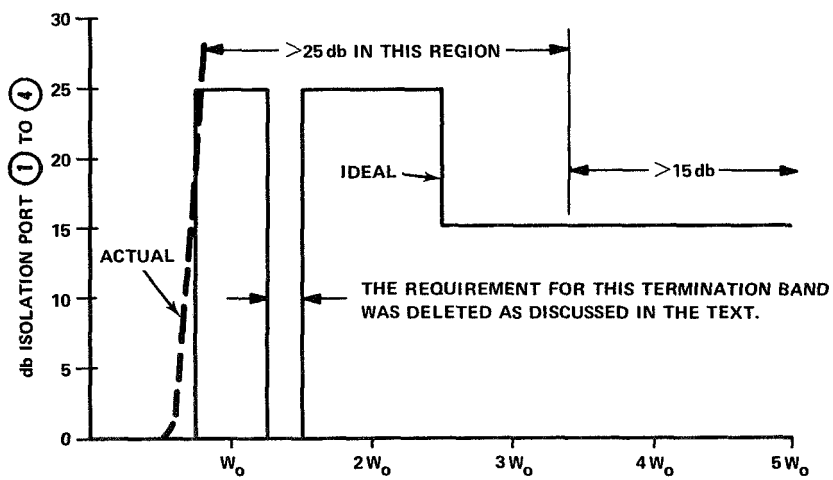
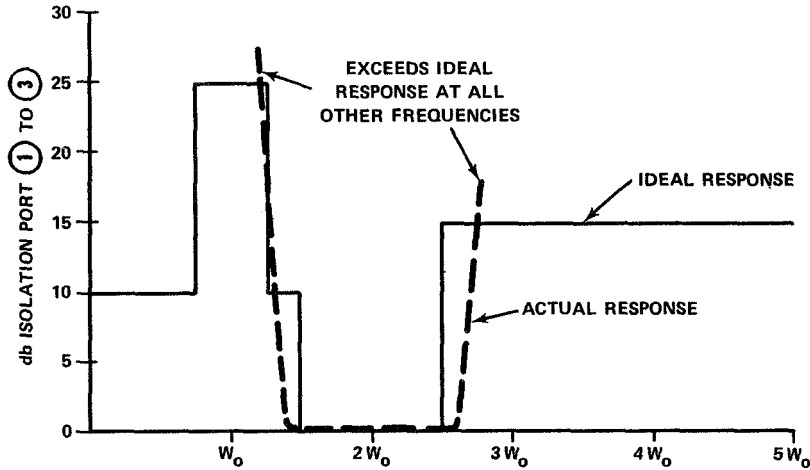
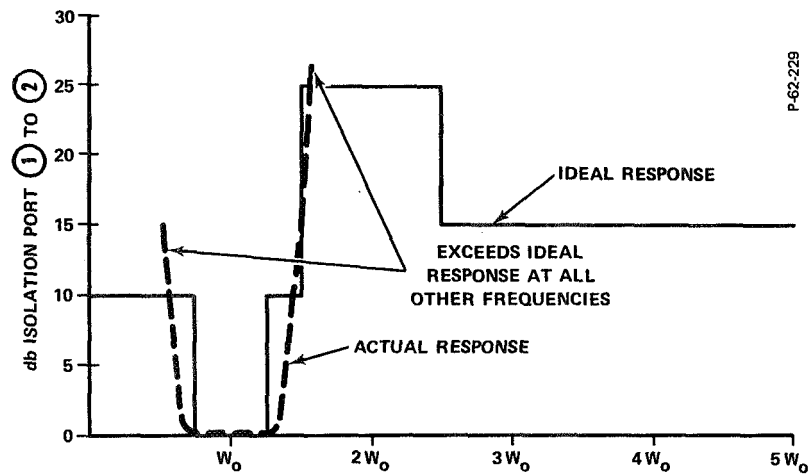
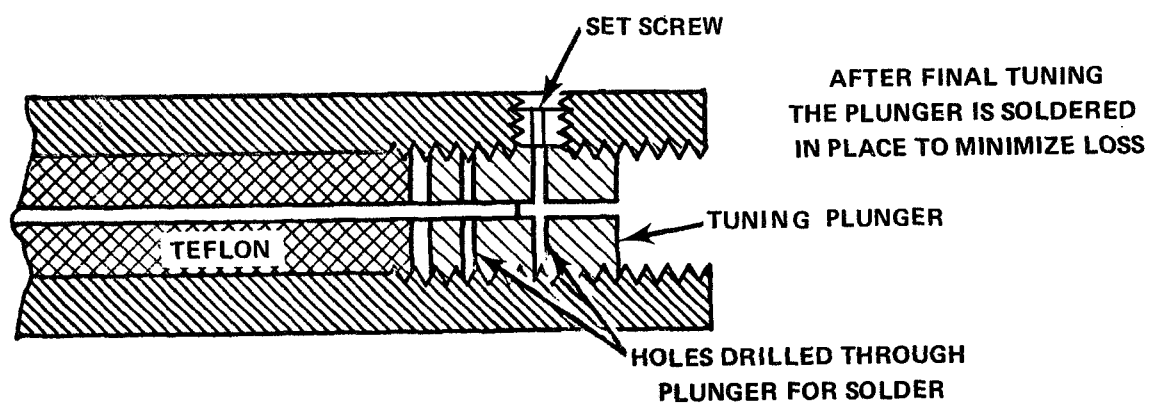
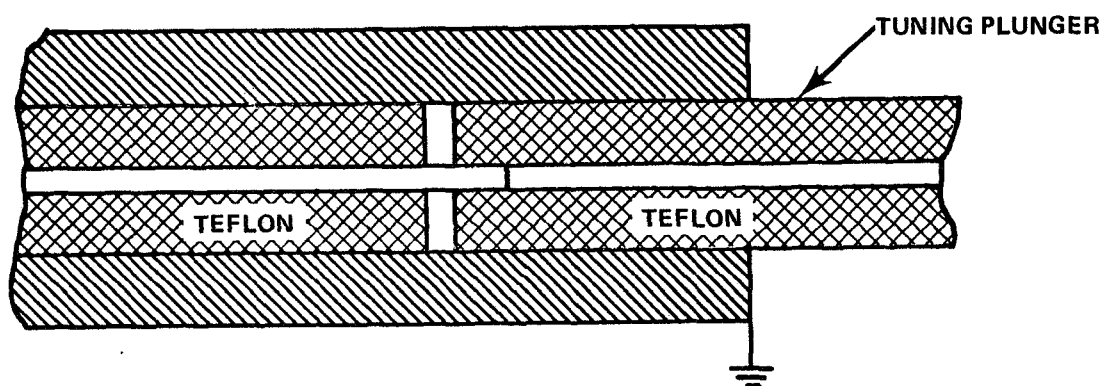


Figure 6 - Theoretical Performance of Final Network Versus Ideal Response Specified



(a) Tuning Mechanism For Short-Circuited Series Stubs



(b) Tuning Of Open Circuited Series Stubs

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Figure 7 - Construction and Tuning of Series Stubs

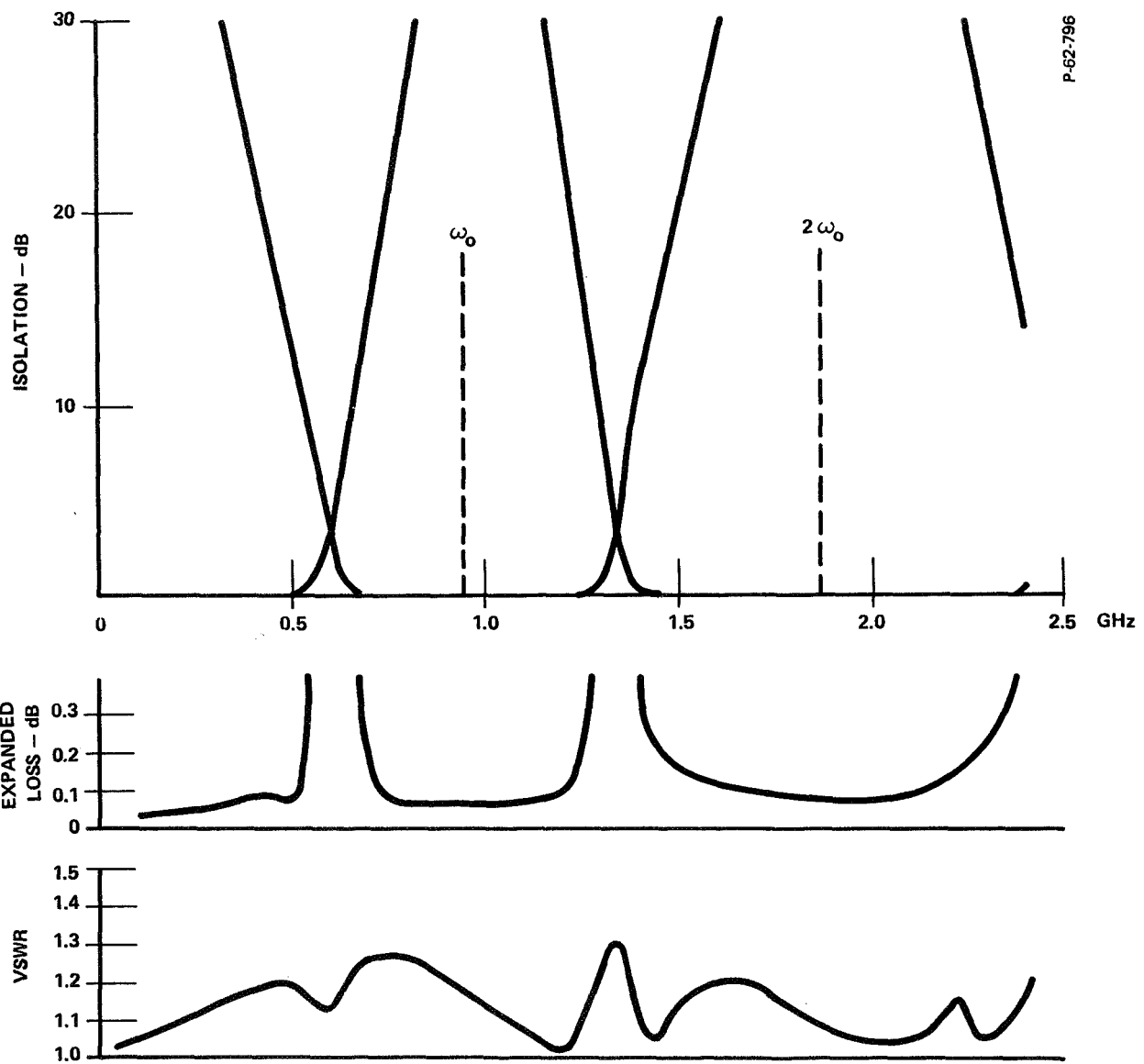


Figure 8 - Performance Characteristics of Primary Filter
for 1-2 GHz Triplexing Filter

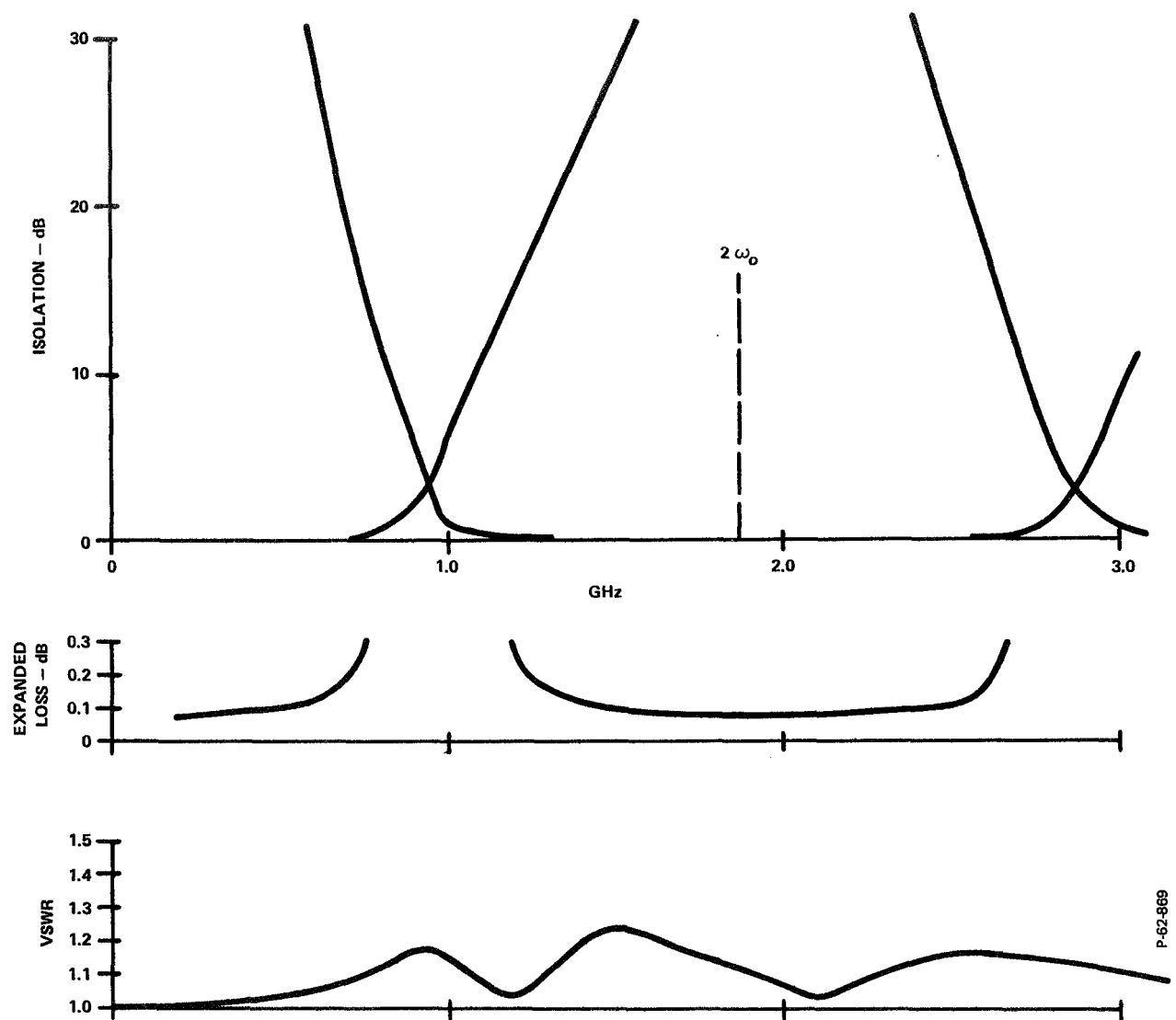


Figure 9 - Measured Performance of $\omega_0 = 1.875$ GHz
First Unit Diplexer

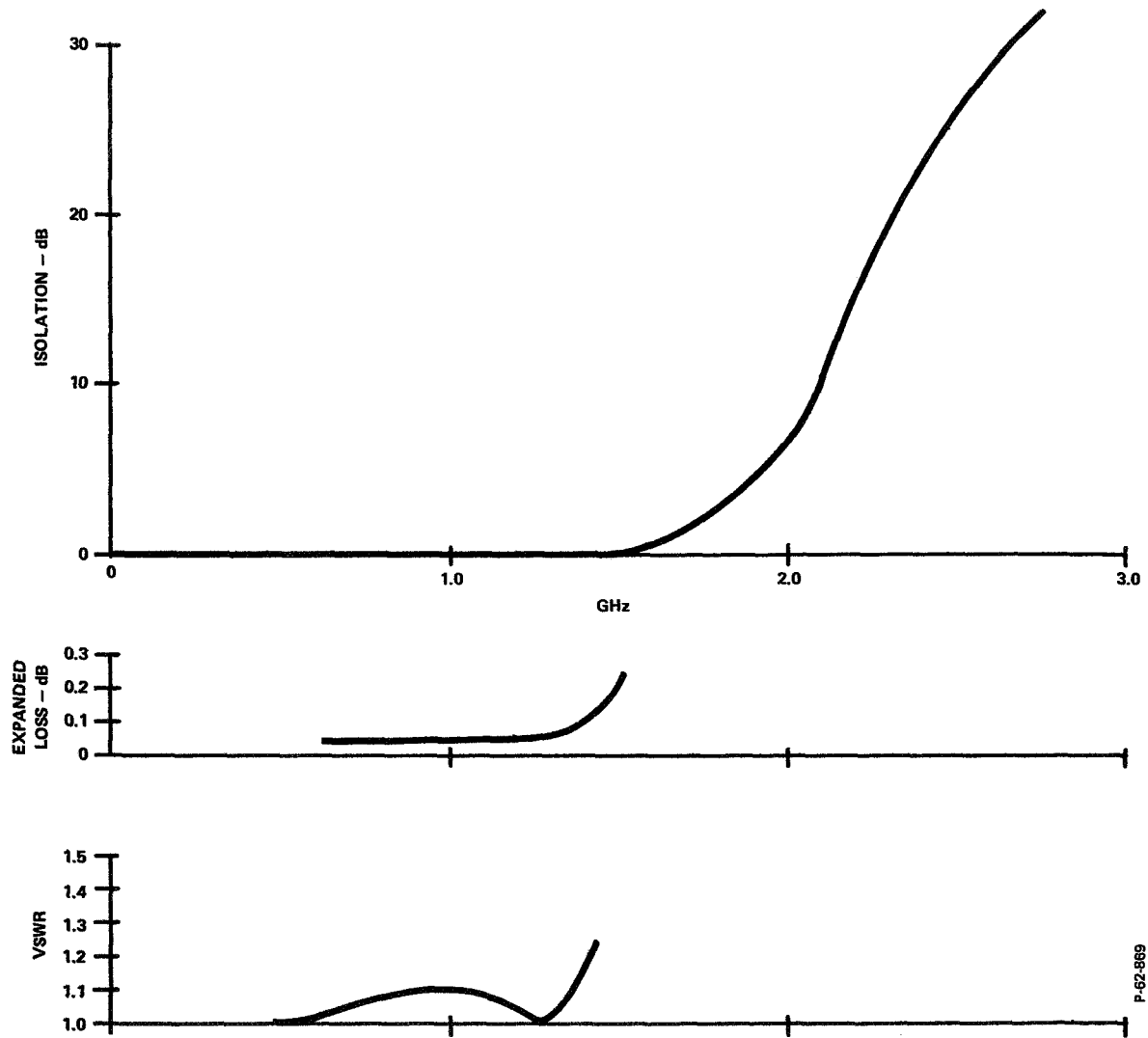


Figure 10 - Measured Performance of First Unit Lowpass Filter

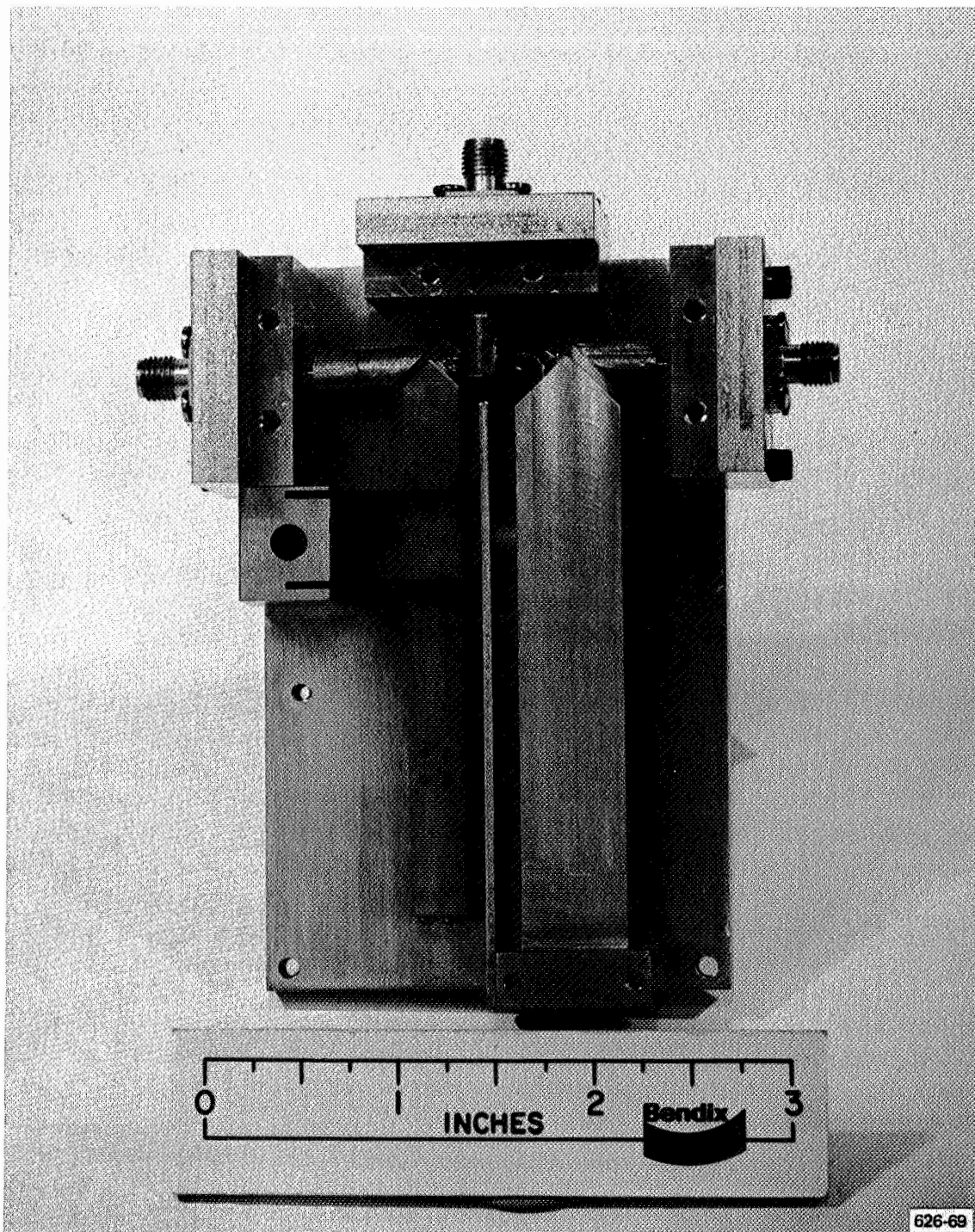


Figure 11 - Primary Diplexer for 1-2 GHz Triplexing Filter

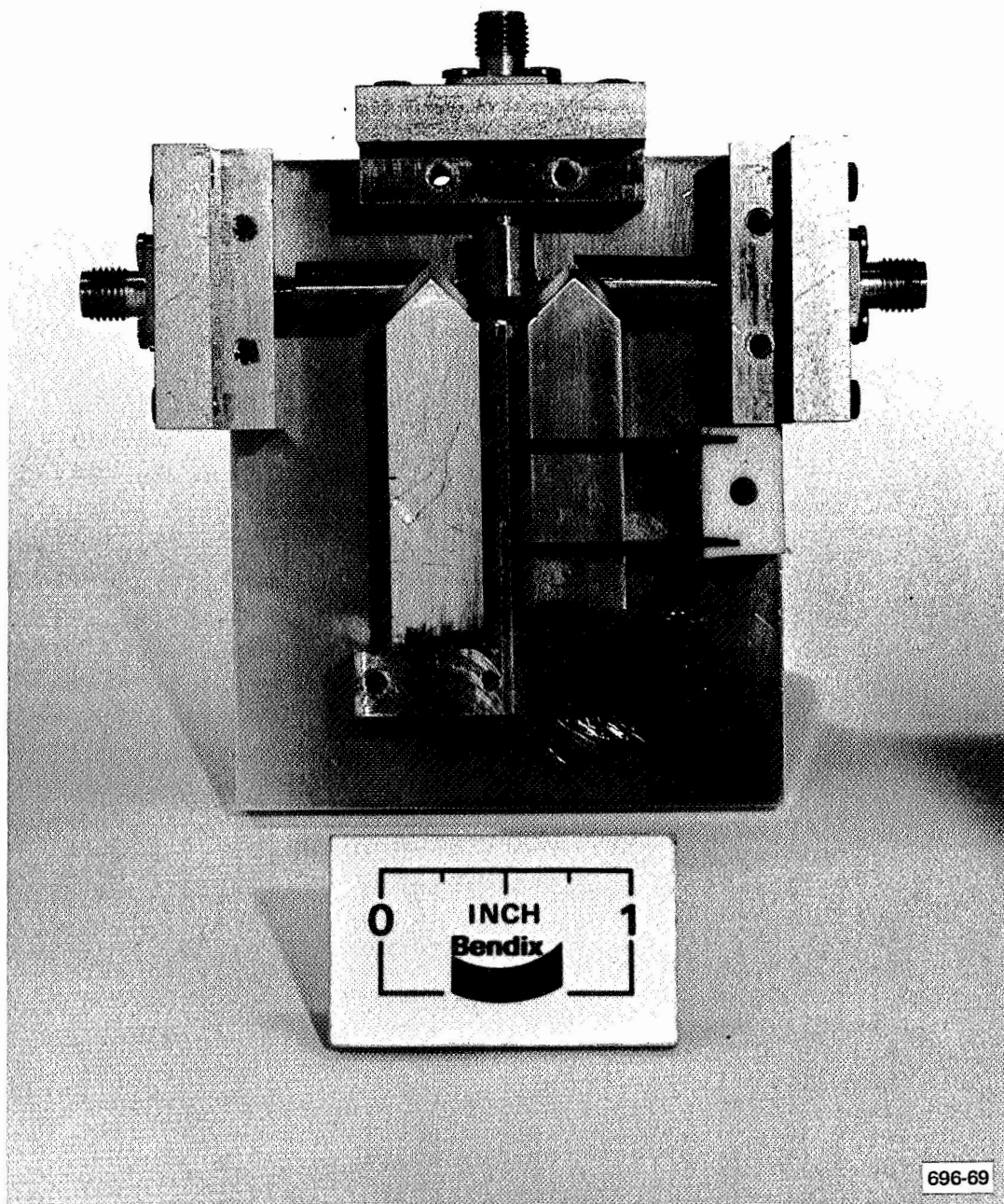


Figure 12 - First Unit Diplexer $\omega_0 = 1.875$ GHz

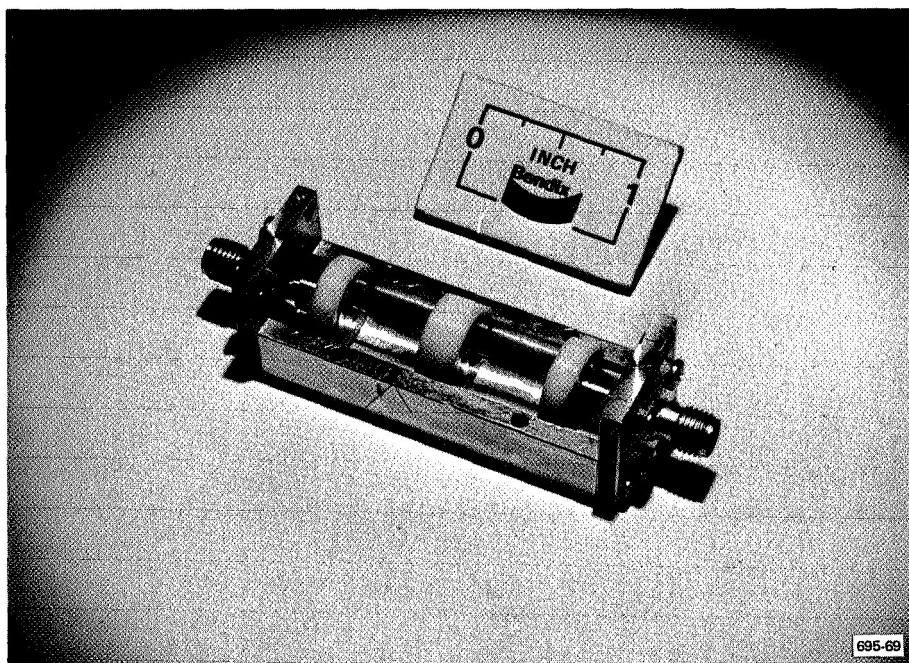


Figure 13 - First Unit Low Pass Filter

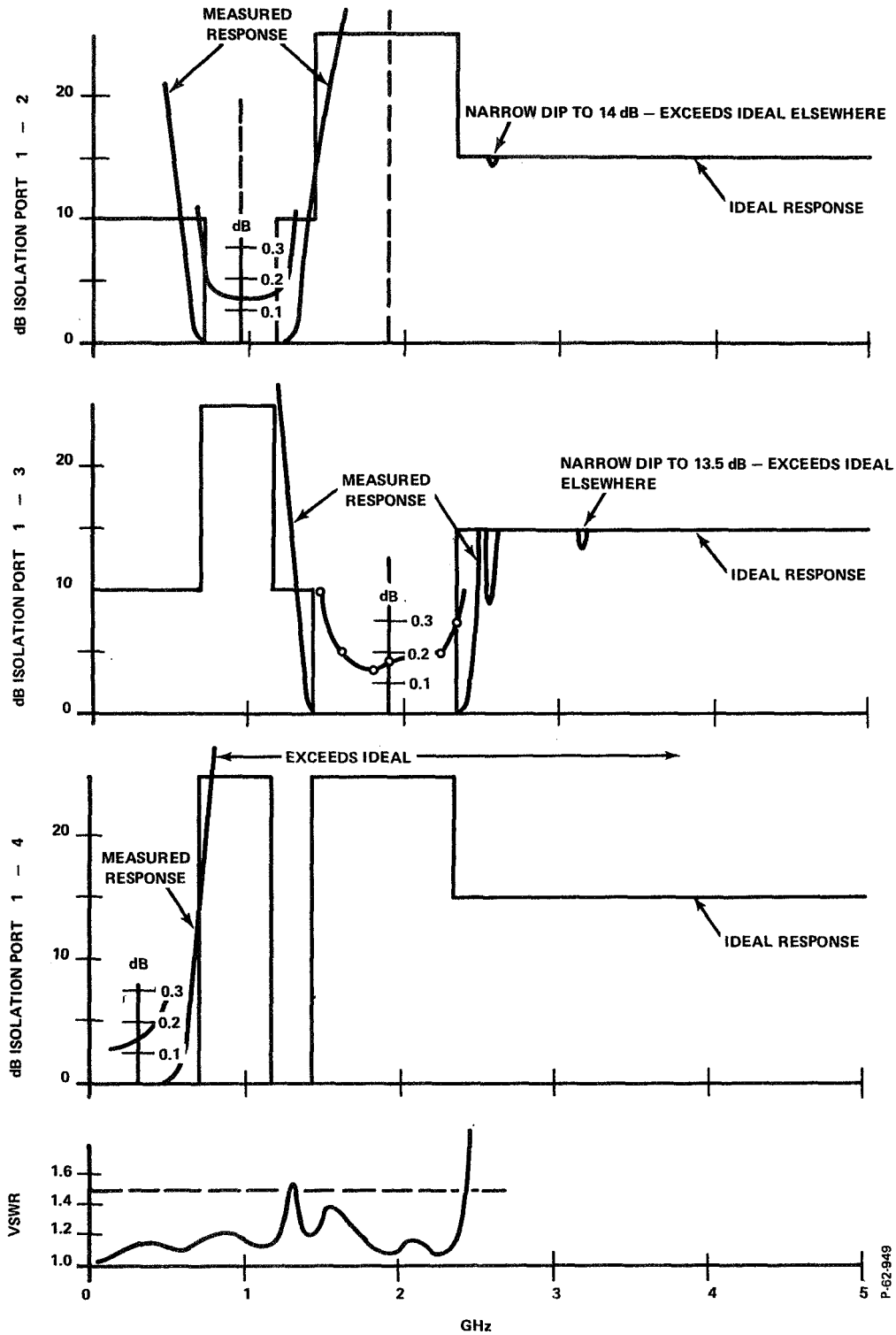
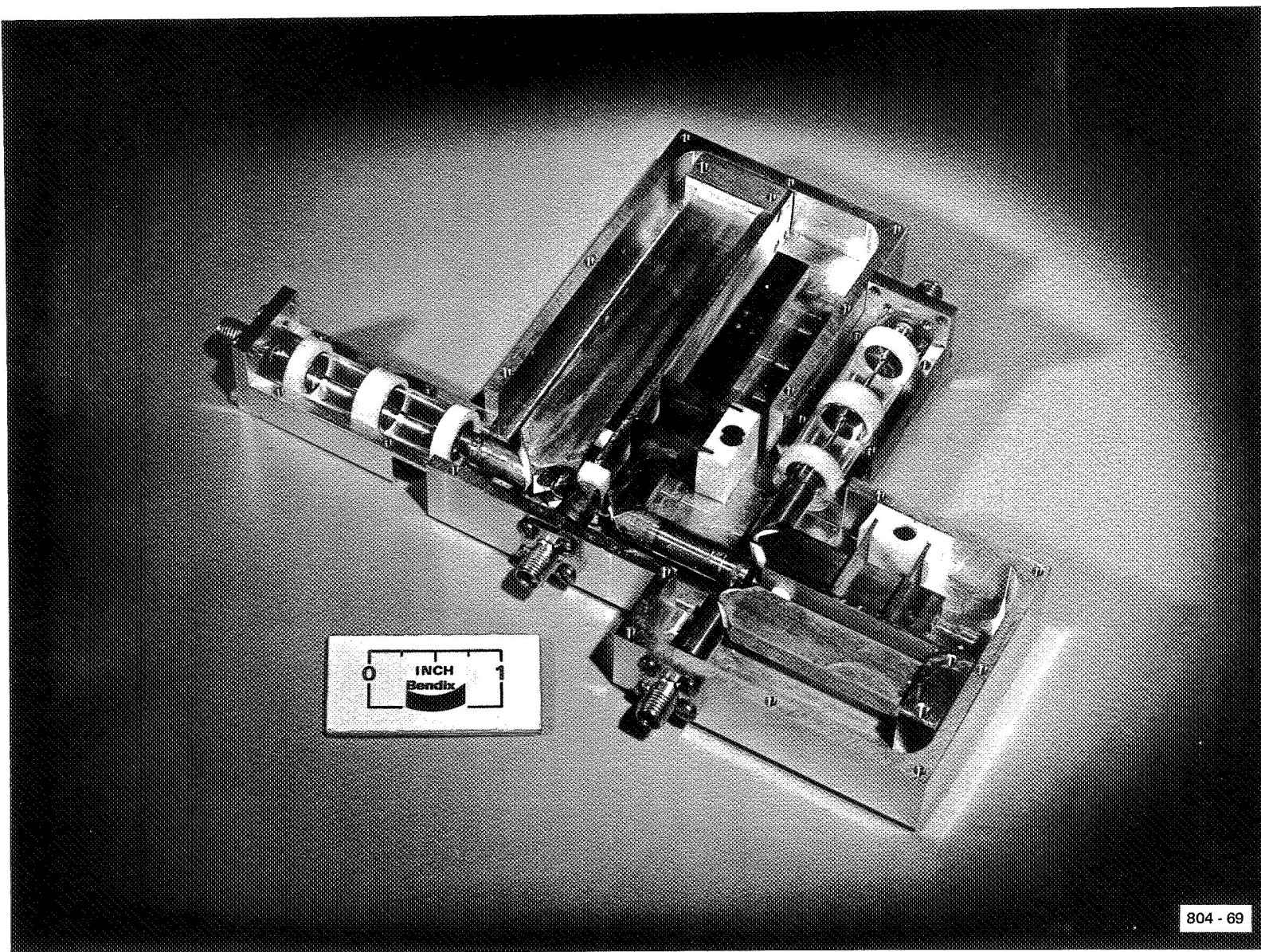


Figure 14 - Measured Response of Complete First Unit Four-Port Triplexing Filter



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Figure 15 - Four-Port Triplexing Filter - First Unit

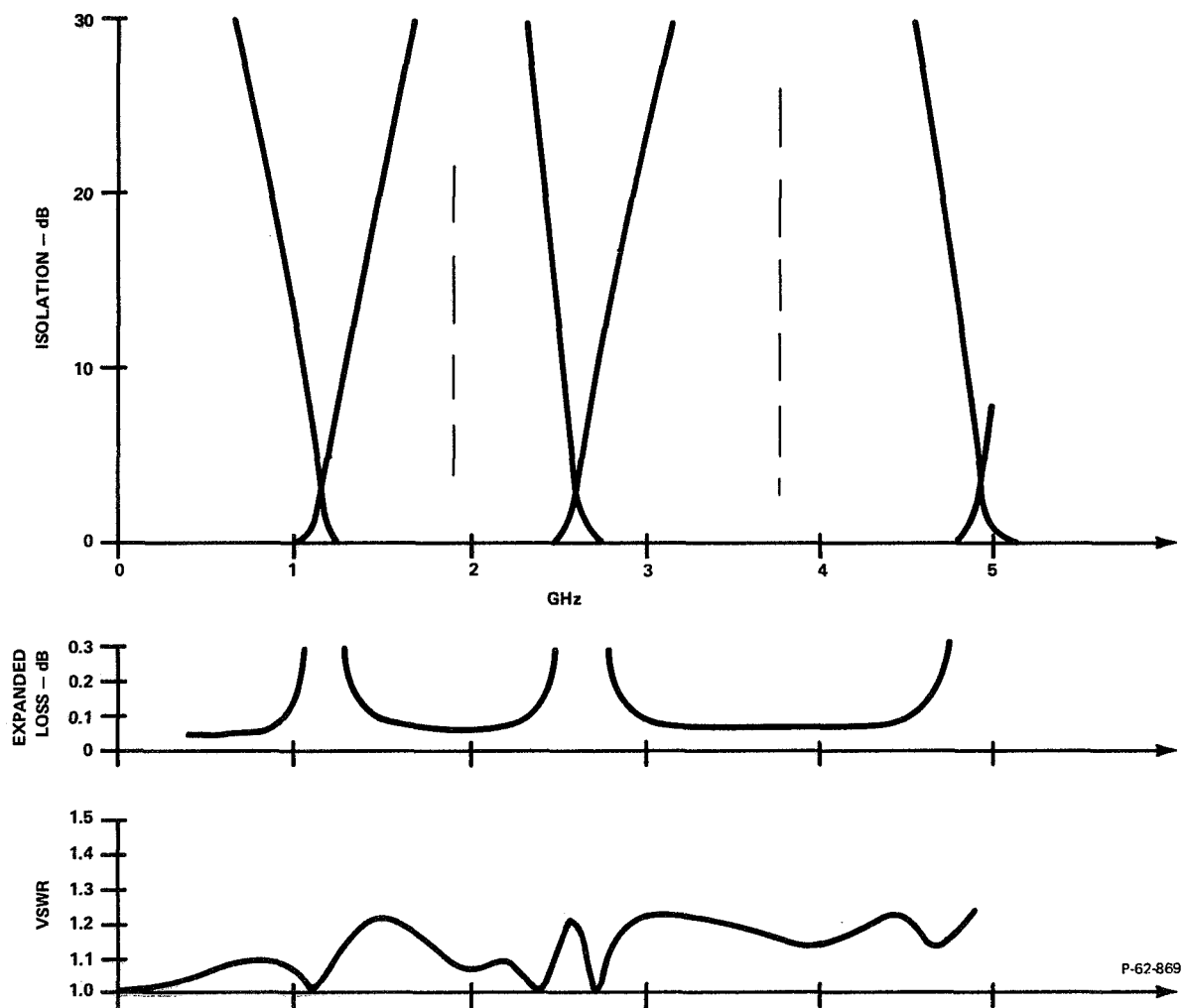


Figure 16 - Measured Performance of Primary $\omega_0 = 1.875$ GHz
Second Unit Diplexer

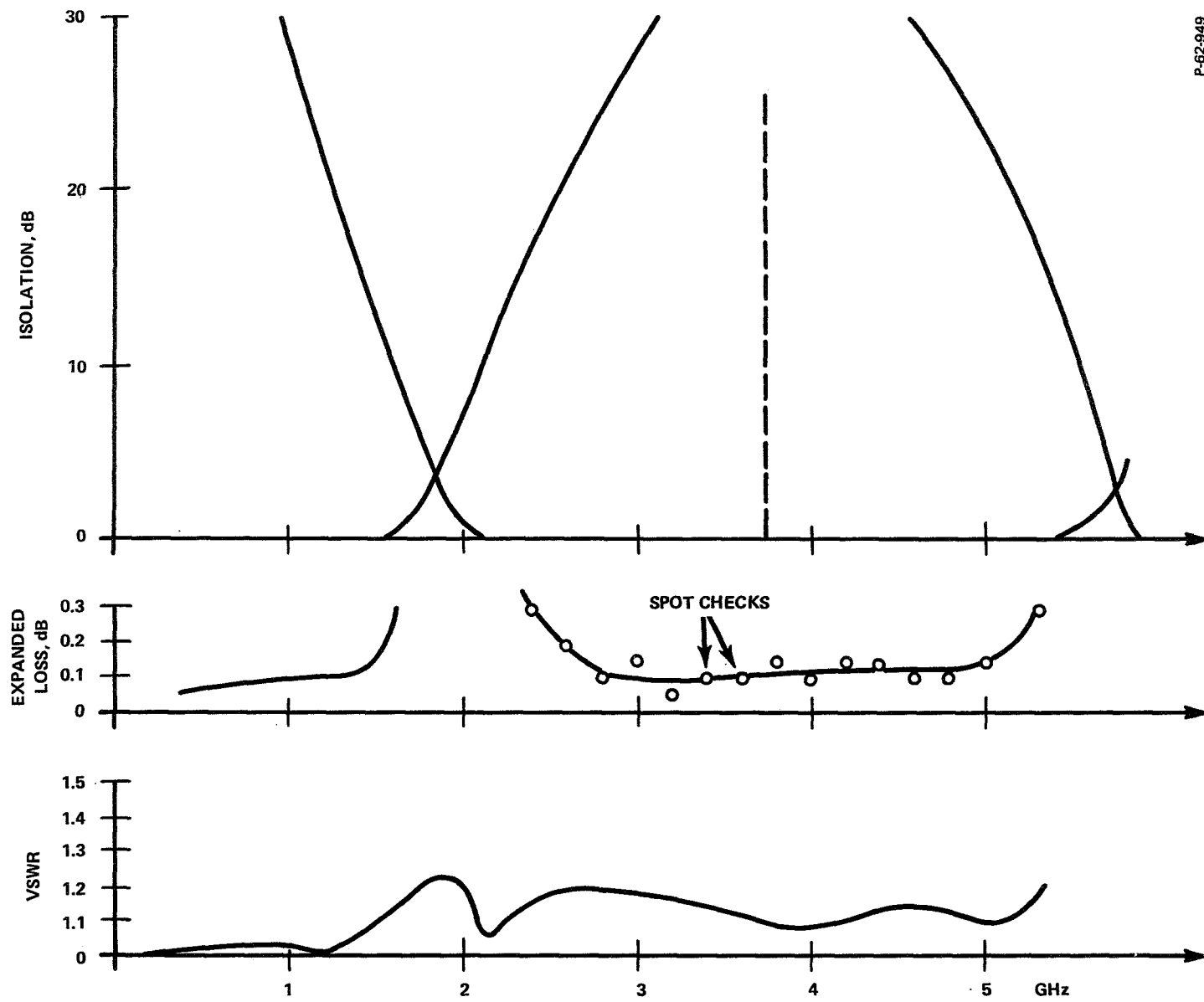
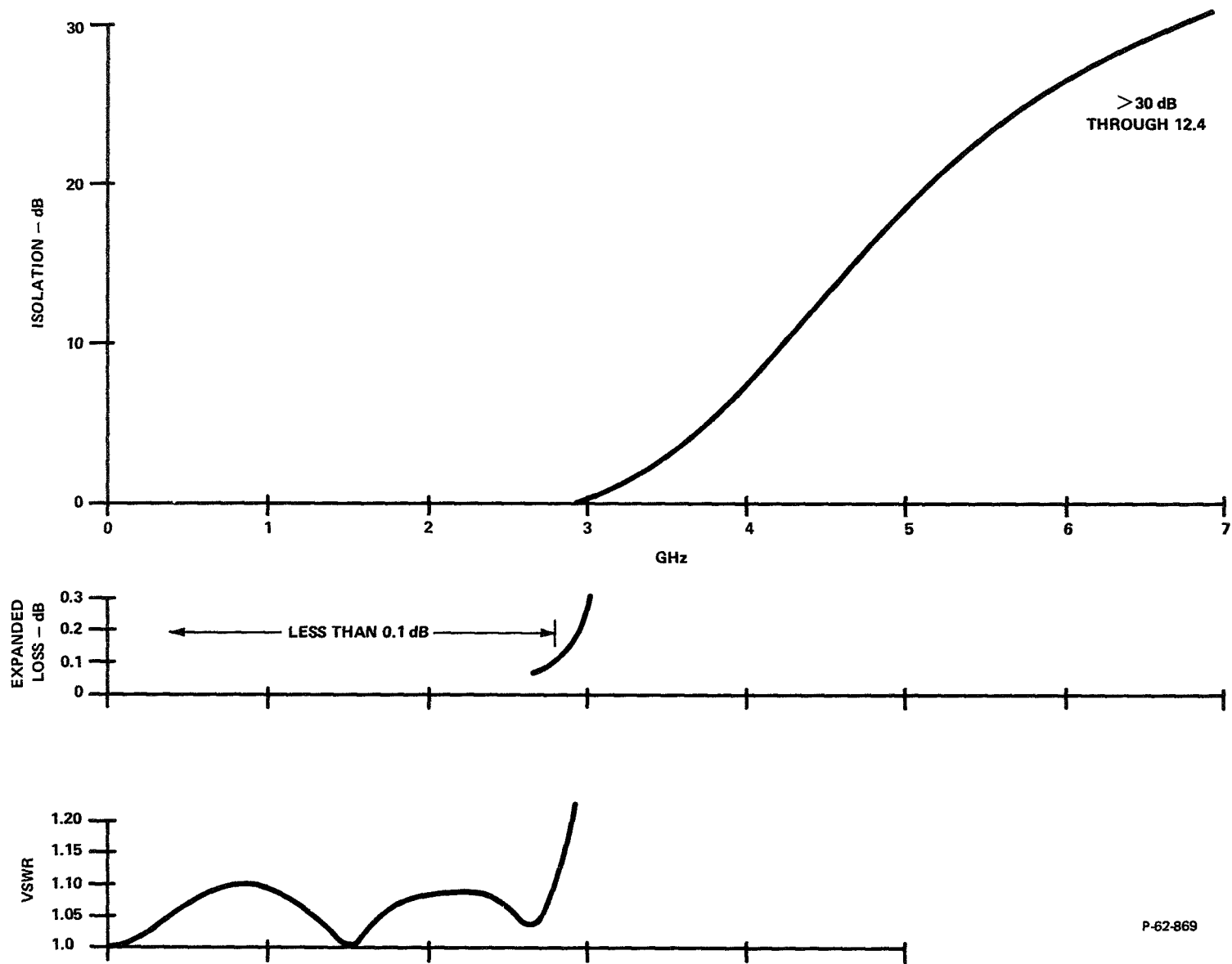


Figure 17 - Measured Response of $\omega_0 = 3.75$ GHz Second Unit Diplexer



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Figure 18 - Measured Performance of Second Unit Lowpass Filter

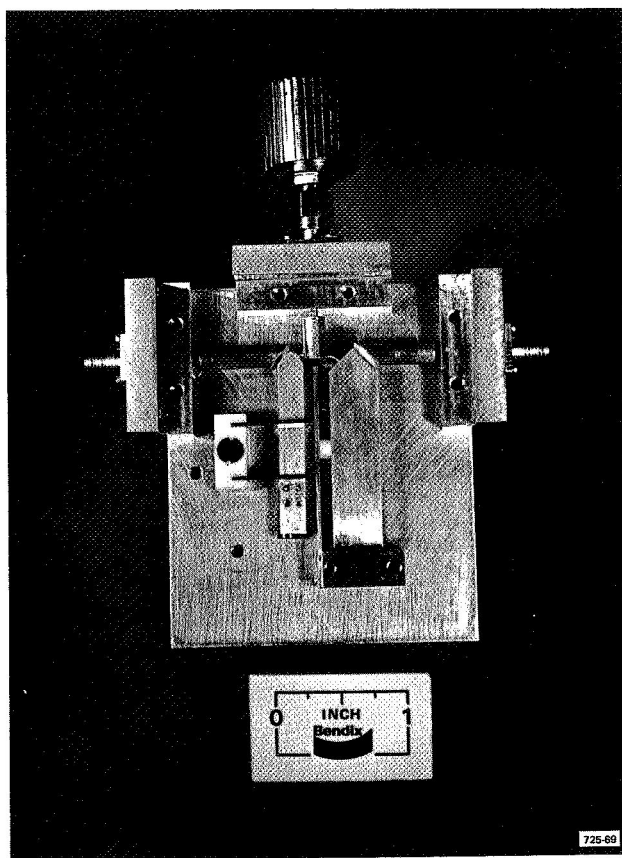


Figure 19 - Second Unit Diplexer $\omega_0 = 1.875$ GHz

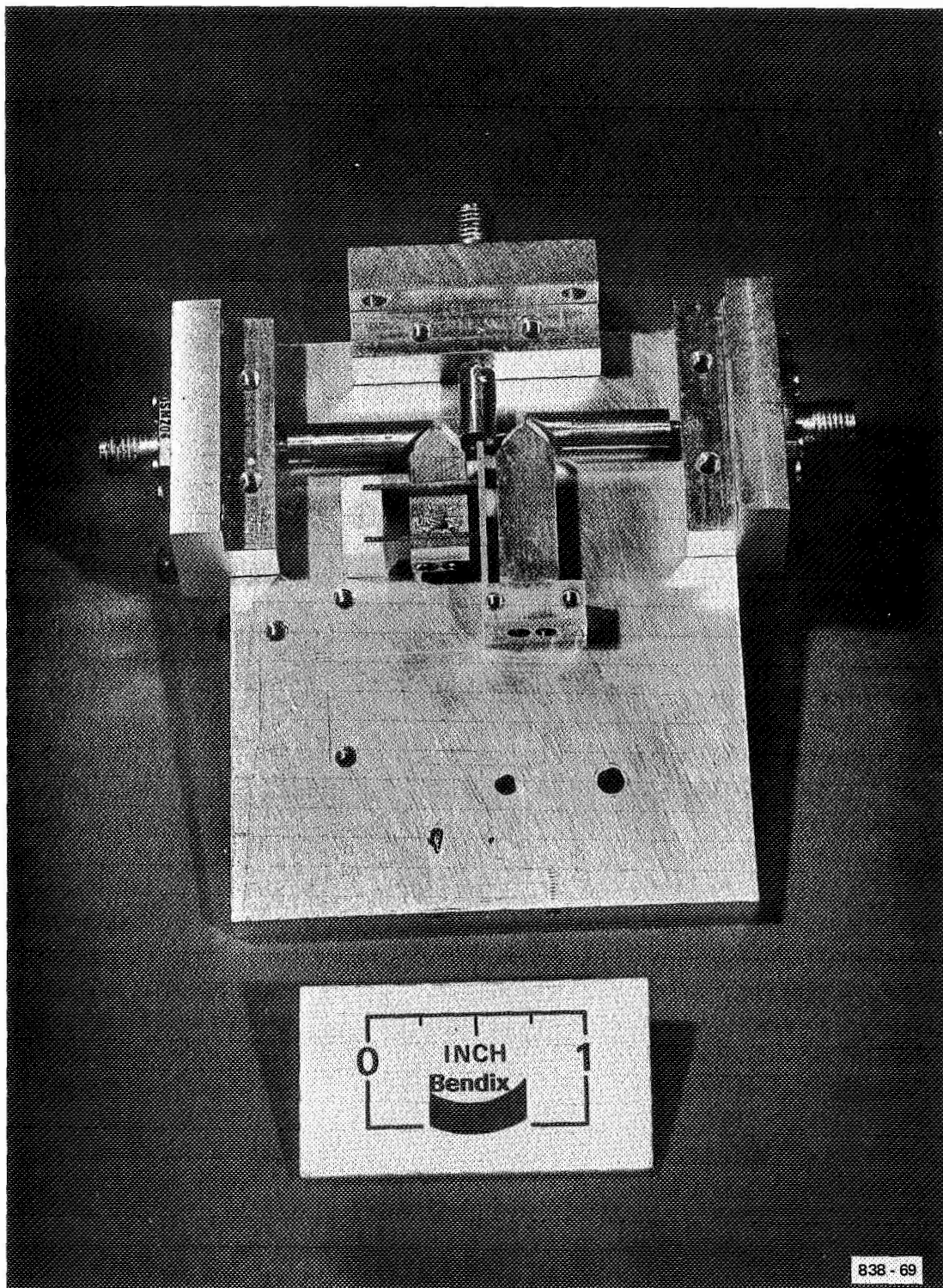


Figure 20 - $\omega_0 = 3.75$ GHz Second Unit Diplexer

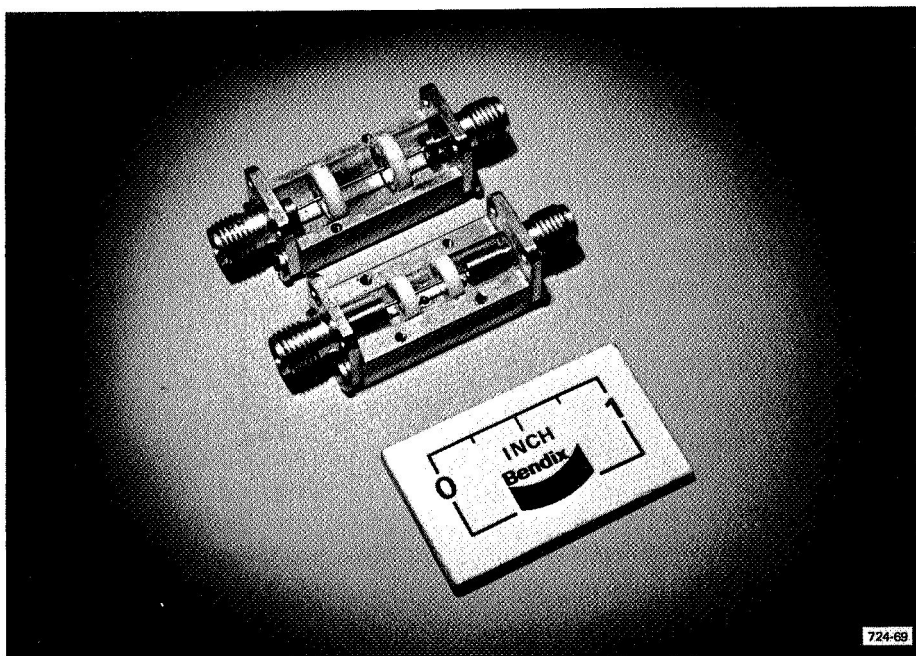


Figure 21 - Second Unit Low Pass Filter (top) and
Third Unit Low Pass Filter (bottom)

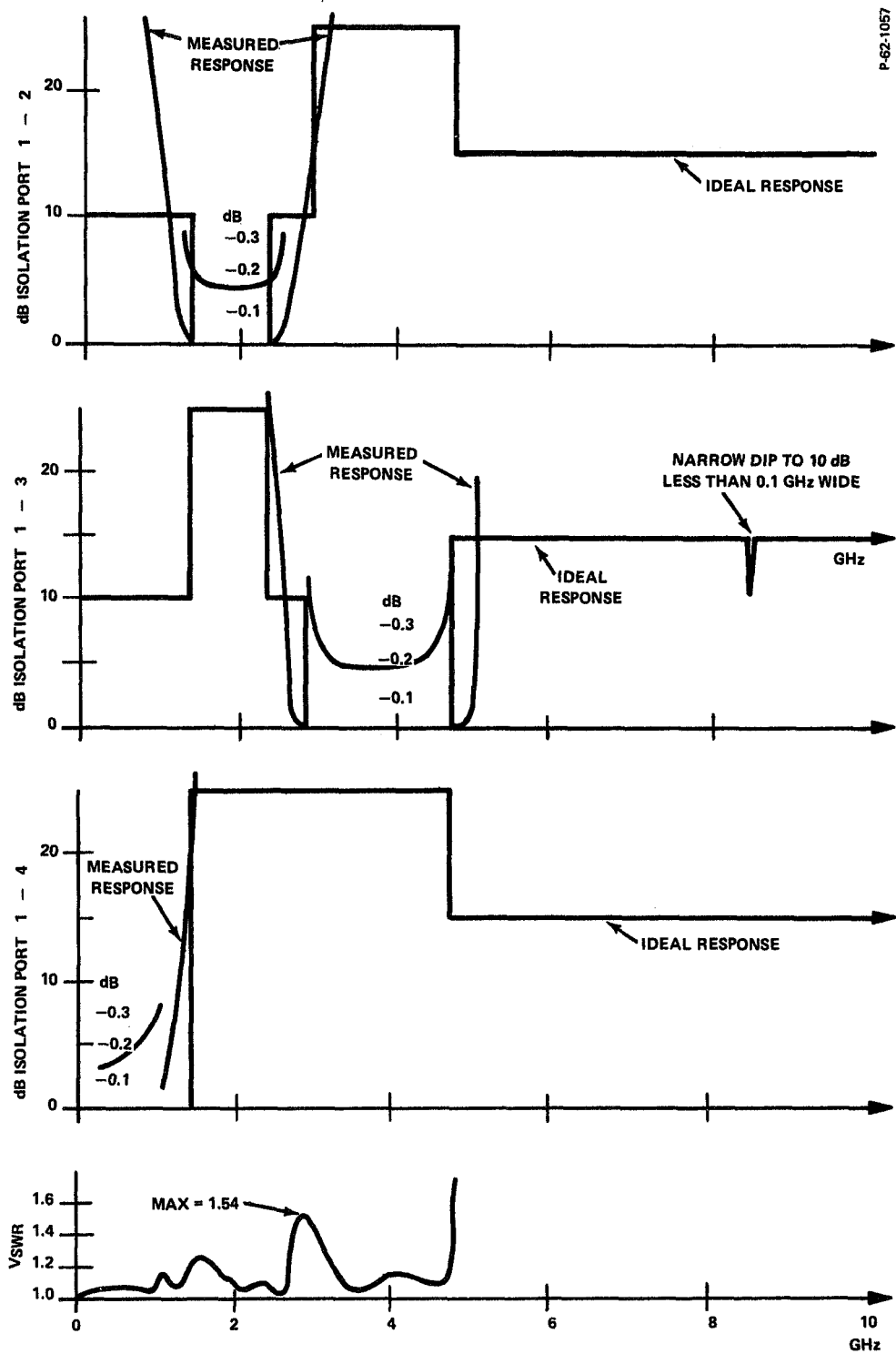


Figure 22 - Measured Response of Complete Second Unit

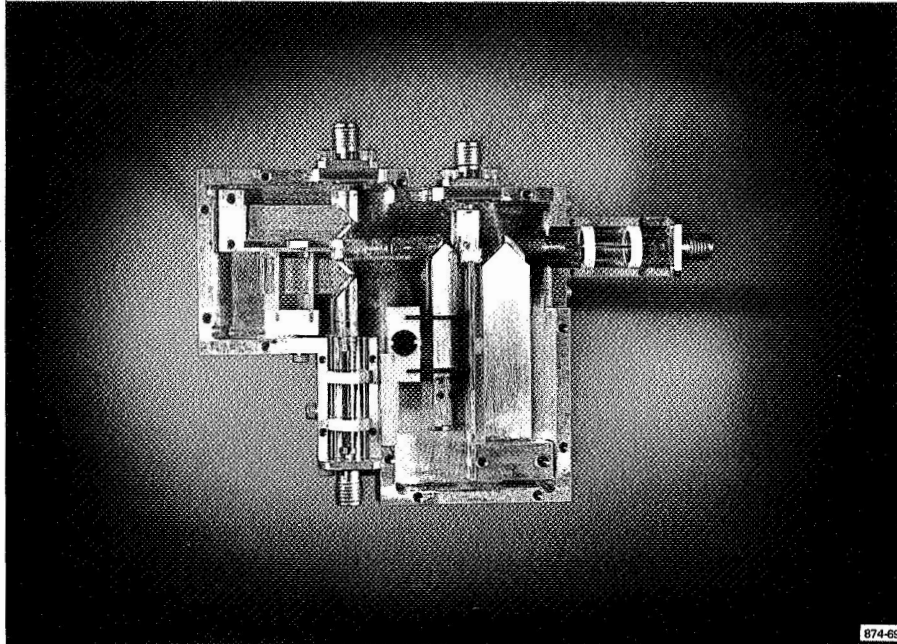


Figure 23 - Interior of Complete Second Unit

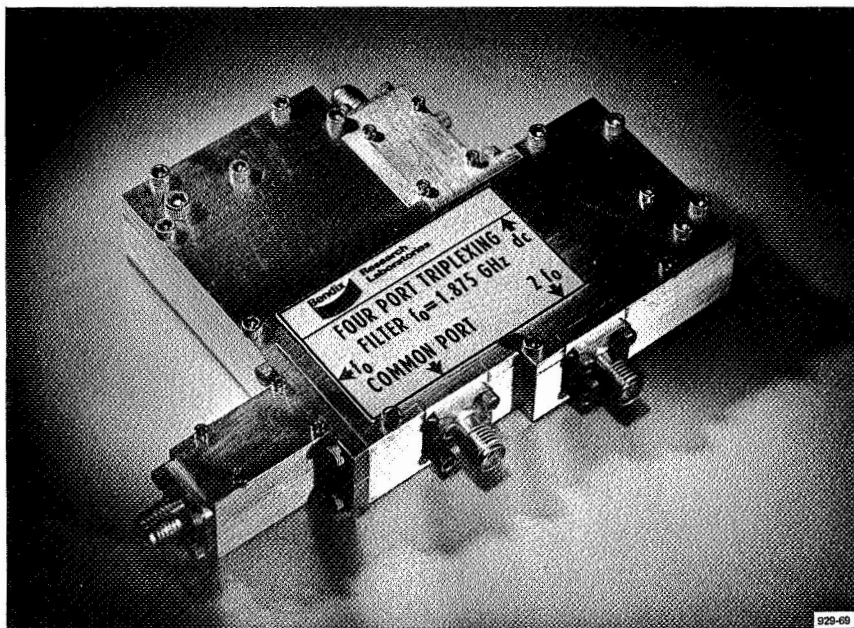


Figure 24 - Complete Second Unit

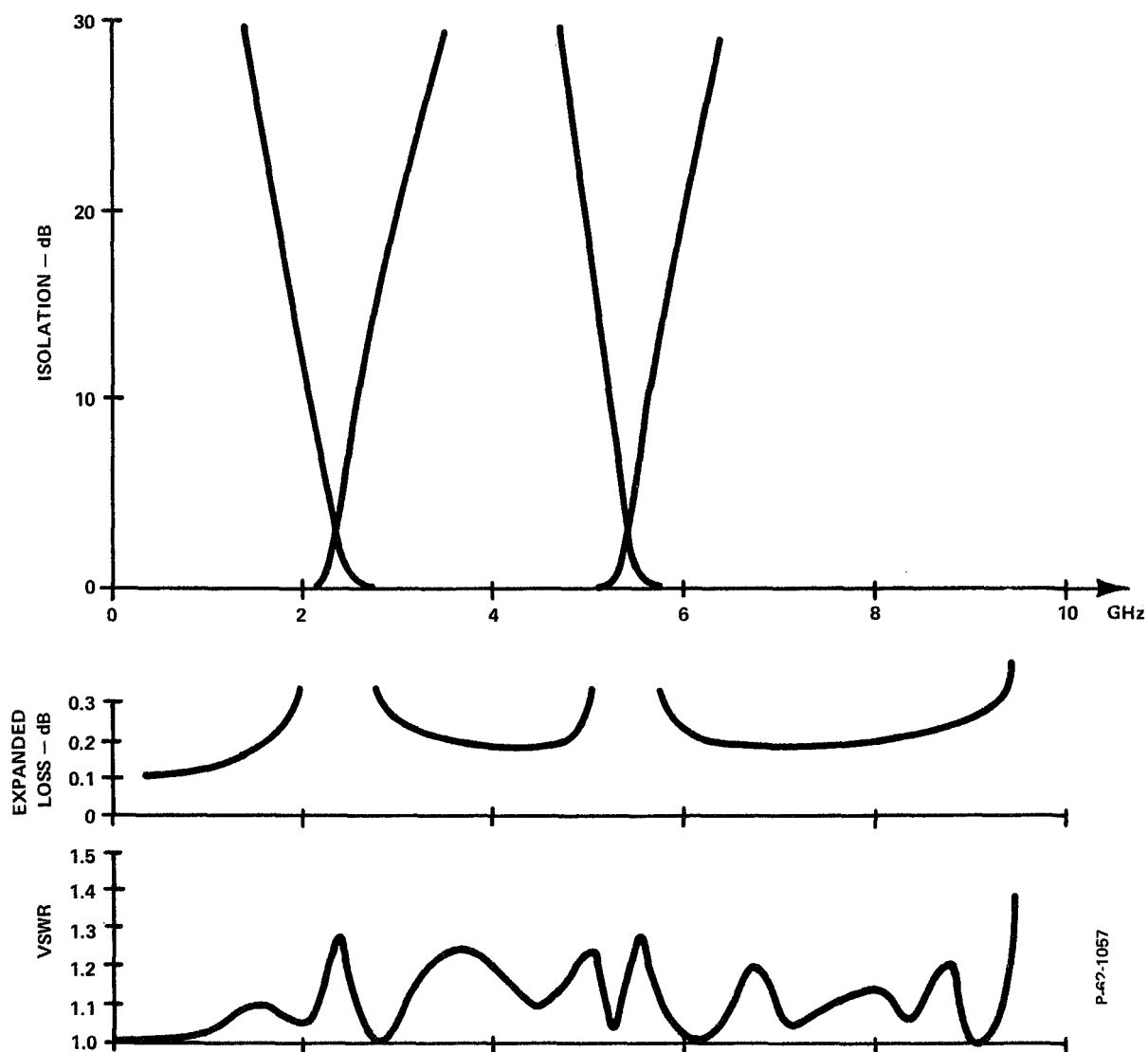


Figure 25 - Measured Response of Third Unit Primary Diplexer,
 $\omega_0 = 3.75$ GHz

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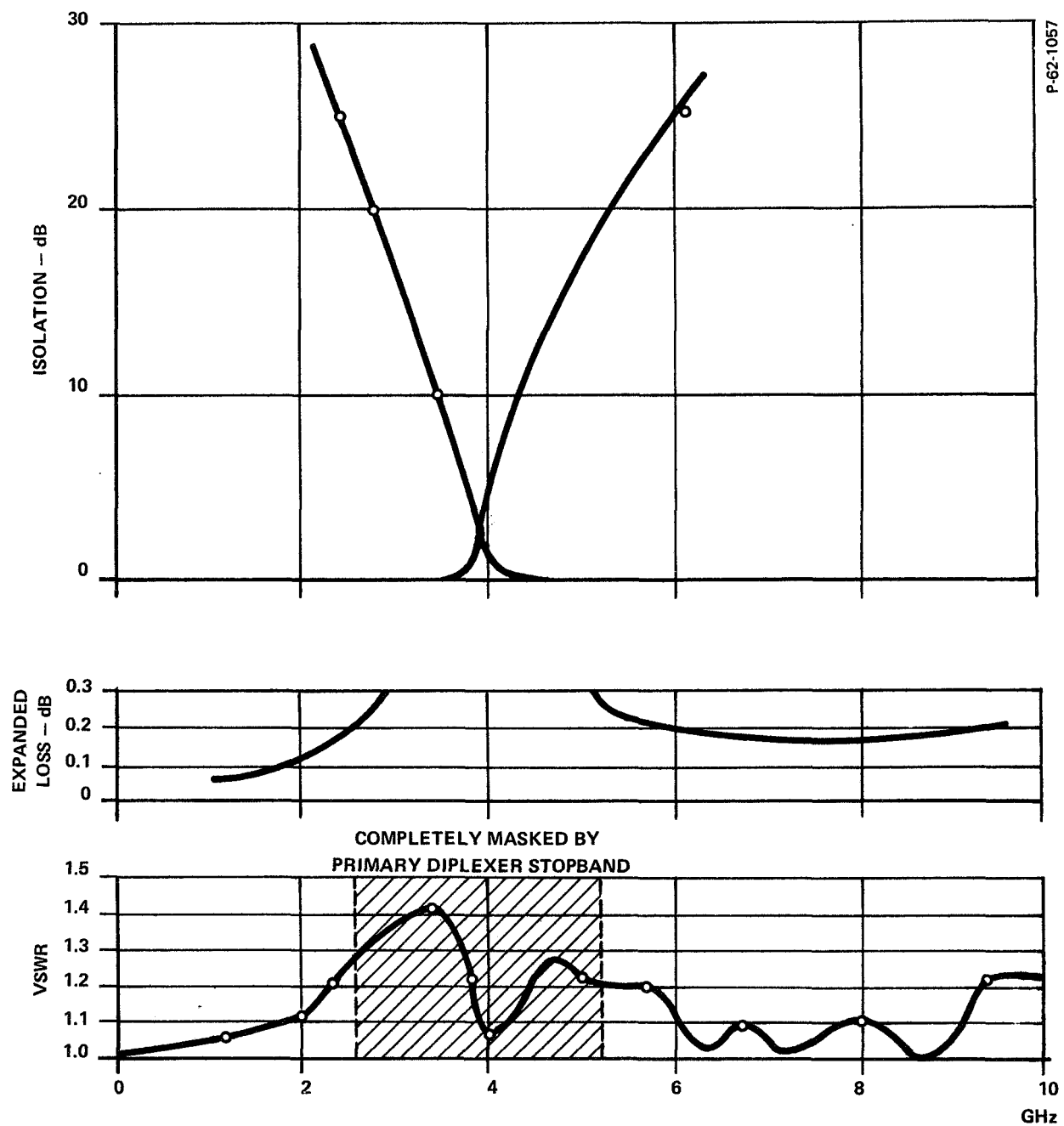
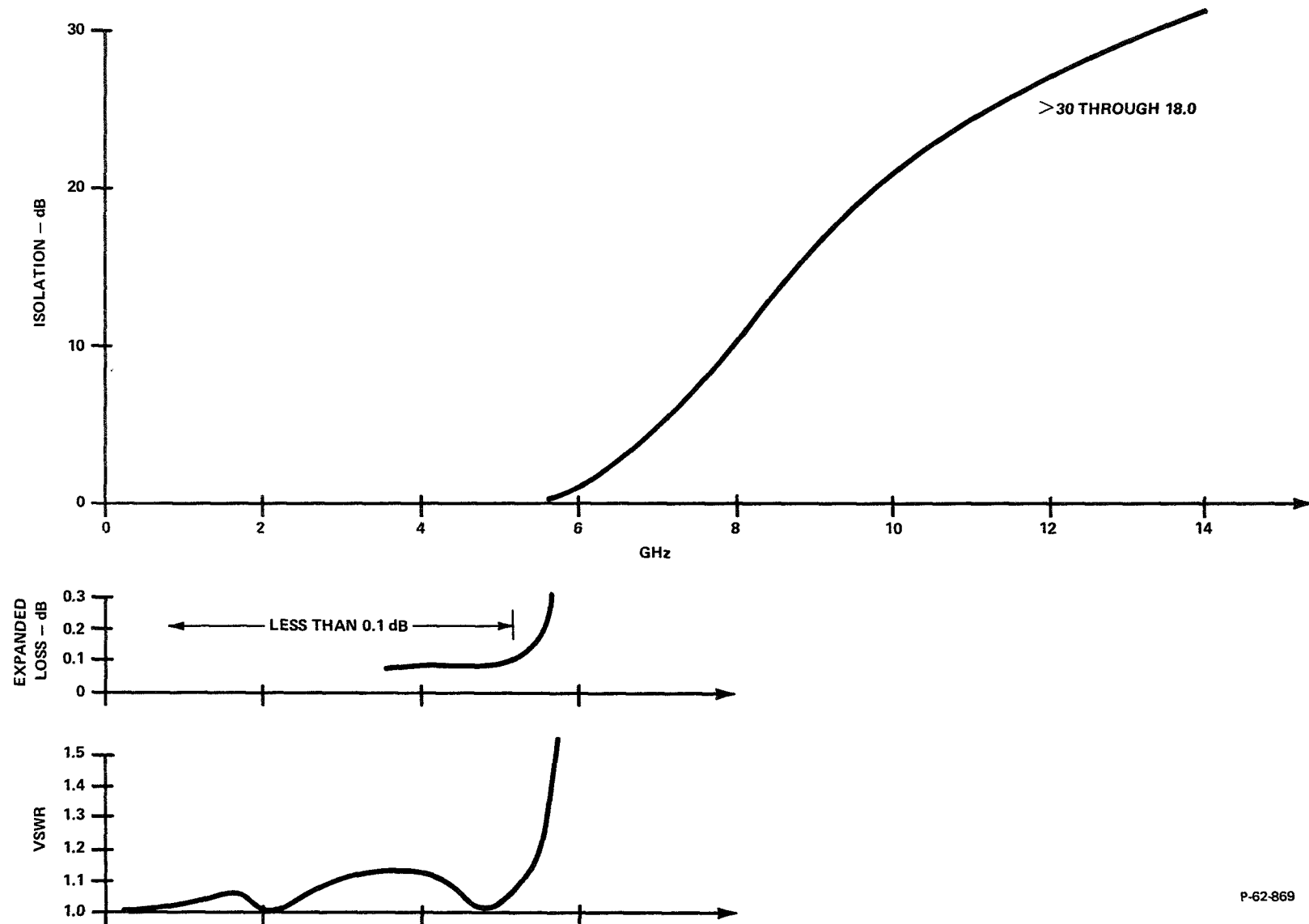


Figure 26 - Measured Response of Third Unit Diplexer,
 $\omega_0 = 7.50$ GHz



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Figure 27 - Measured Performance of Third Unit Lowpass Filter

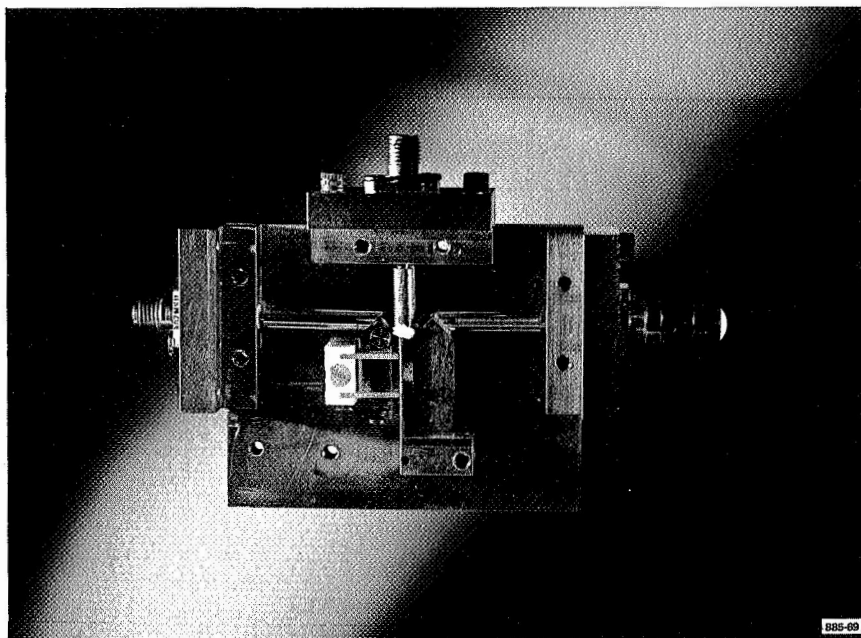


Figure 28 - Third Unit Primary Diplexer

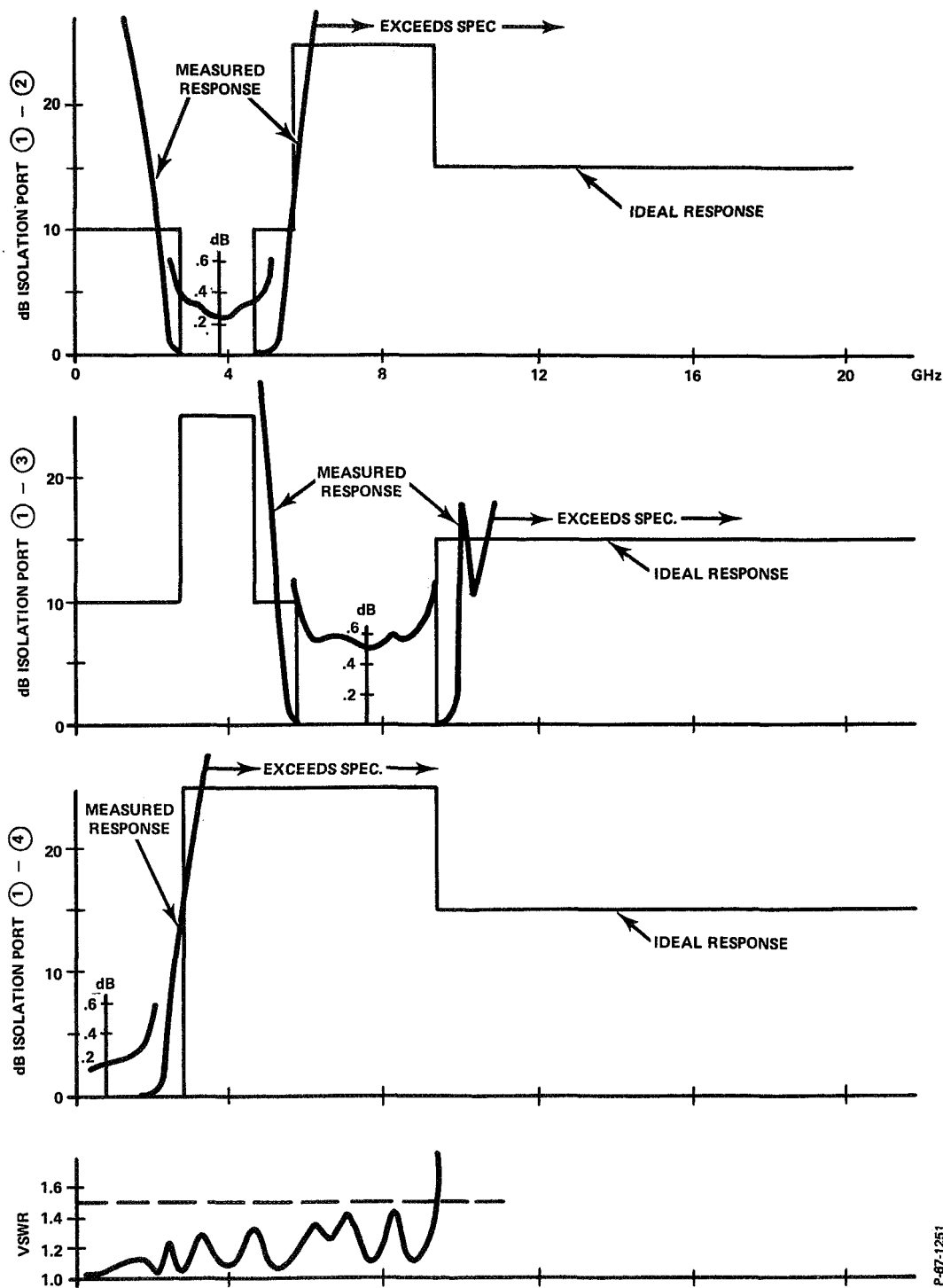


Figure 29 - Measured Response of Complete Third Unit

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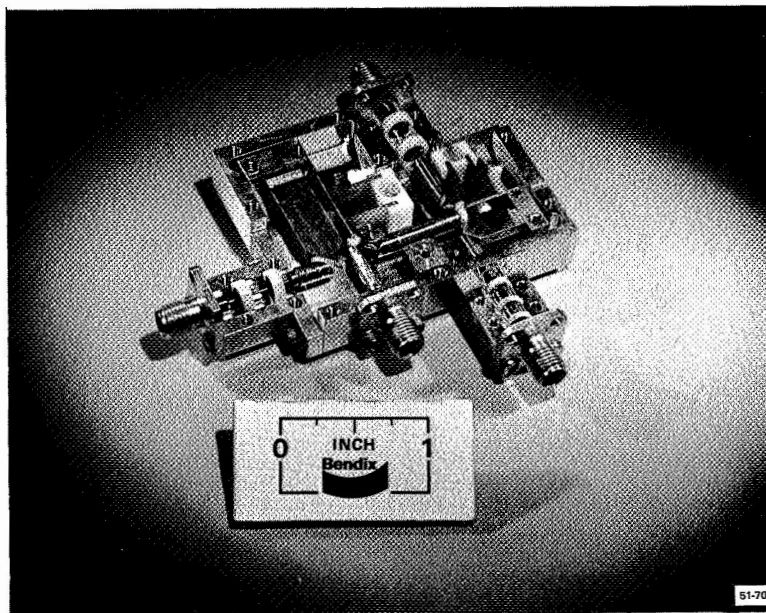


Figure 30 - Interior of Complete Third Unit

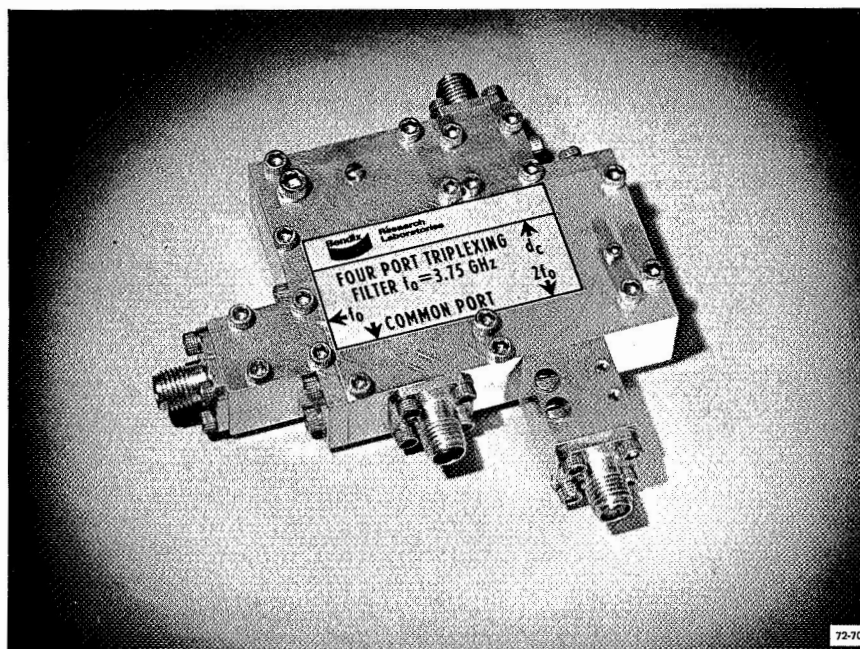


Figure 31 - Complete Third Unit

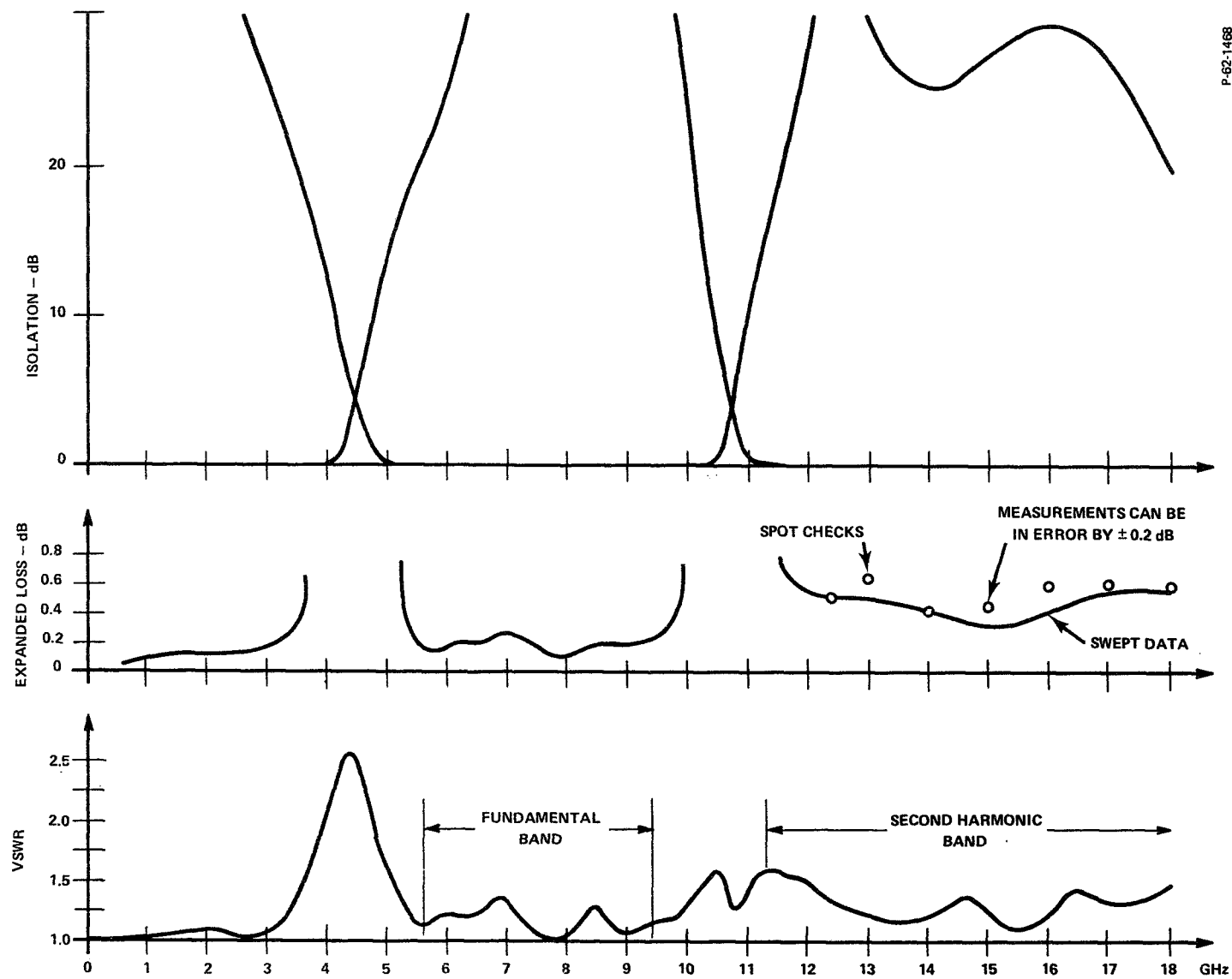


Figure 32 - Measured Performance of Fourth Unit Primary Diplexer - ω_0 - 7.50 GHz

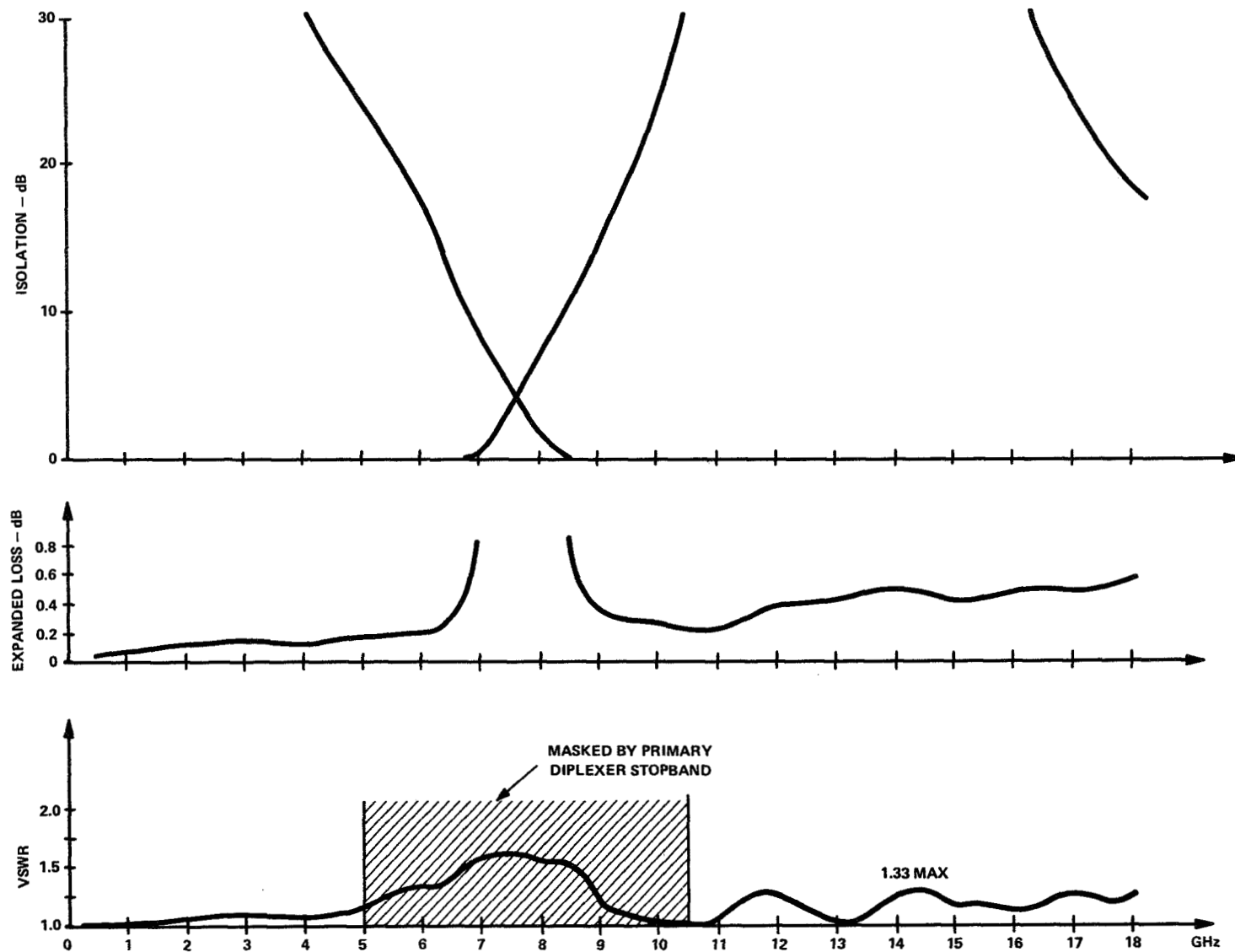


Figure 33 - Measured Performance of Fourth Unit Secondary Diplexer - $\omega_0 = 15.0$ GHz

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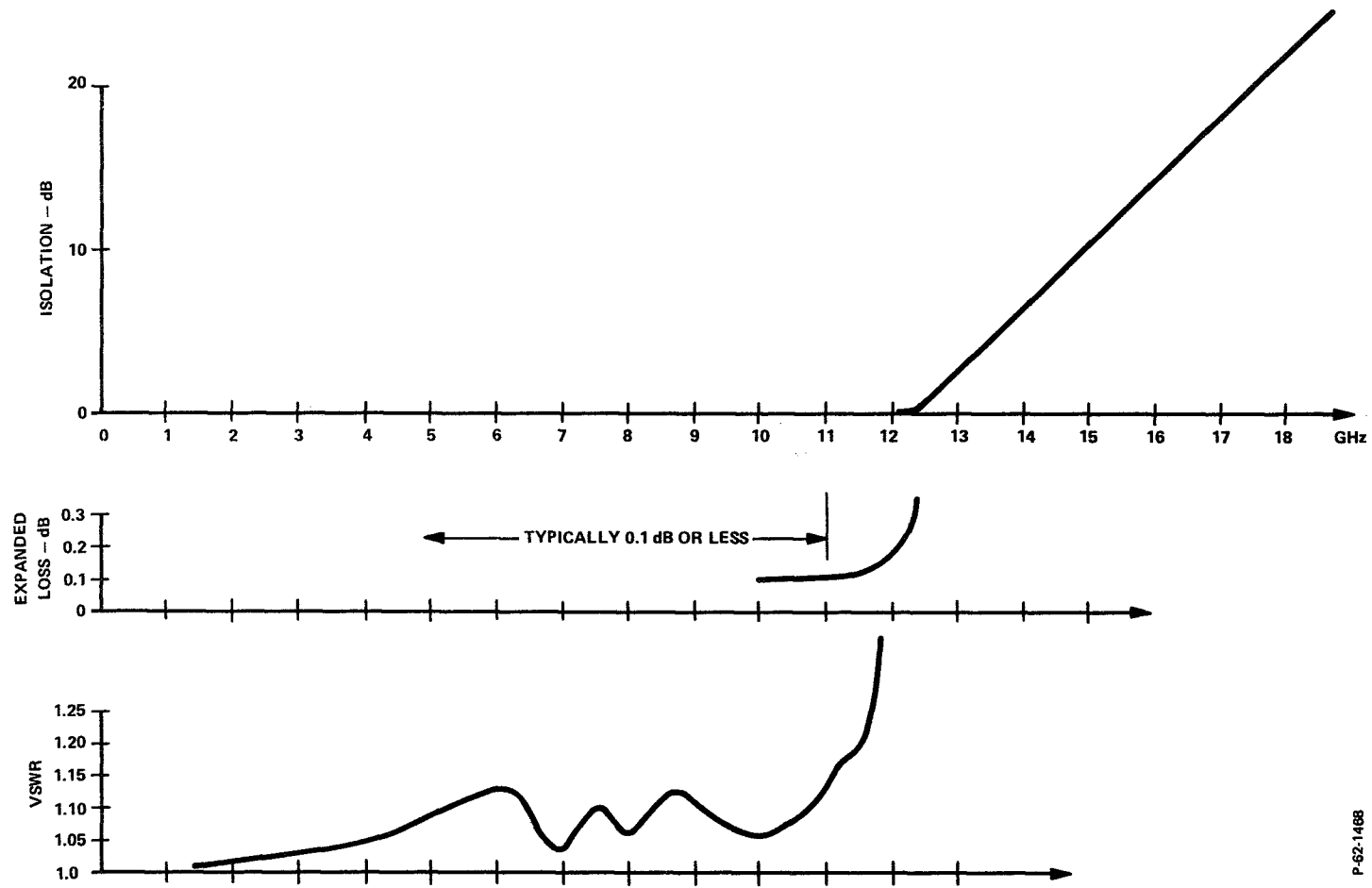


Figure 34 - Measured Performance of Fourth Unit Low Pass Filter

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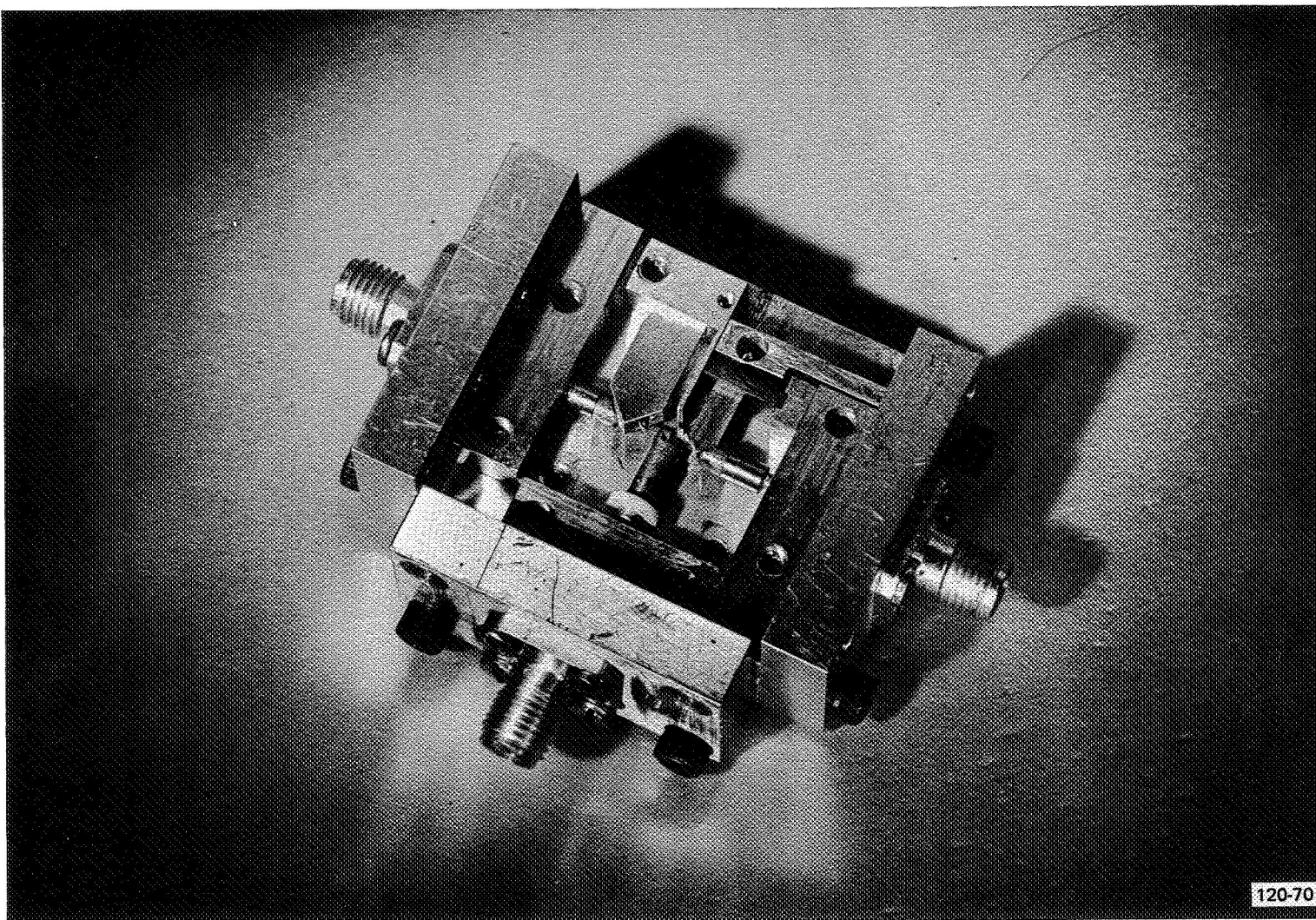


Figure 35 - Fourth Unit Primary Diplexer

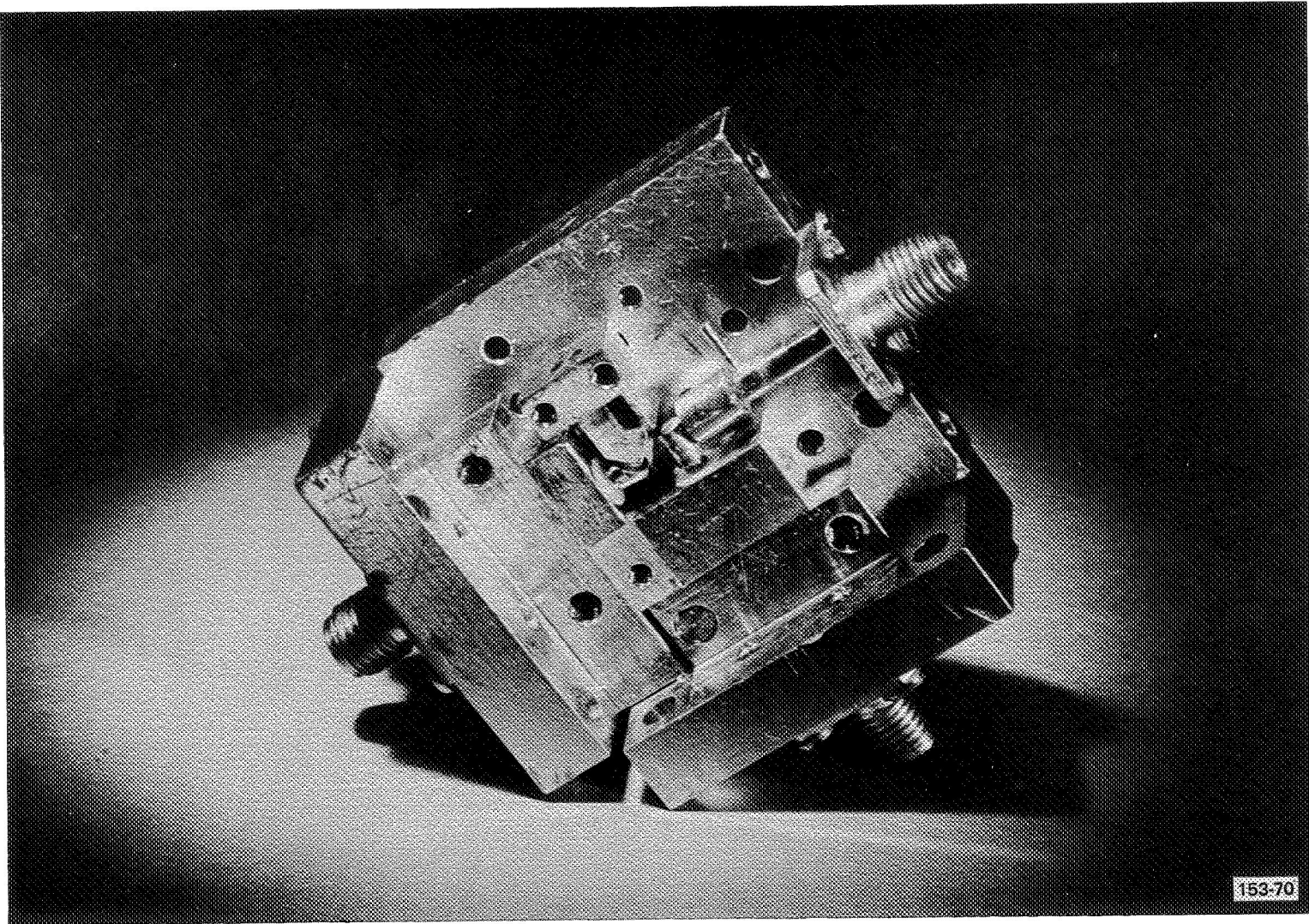


Figure 36 - Fourth Unit Secondary Diplexer

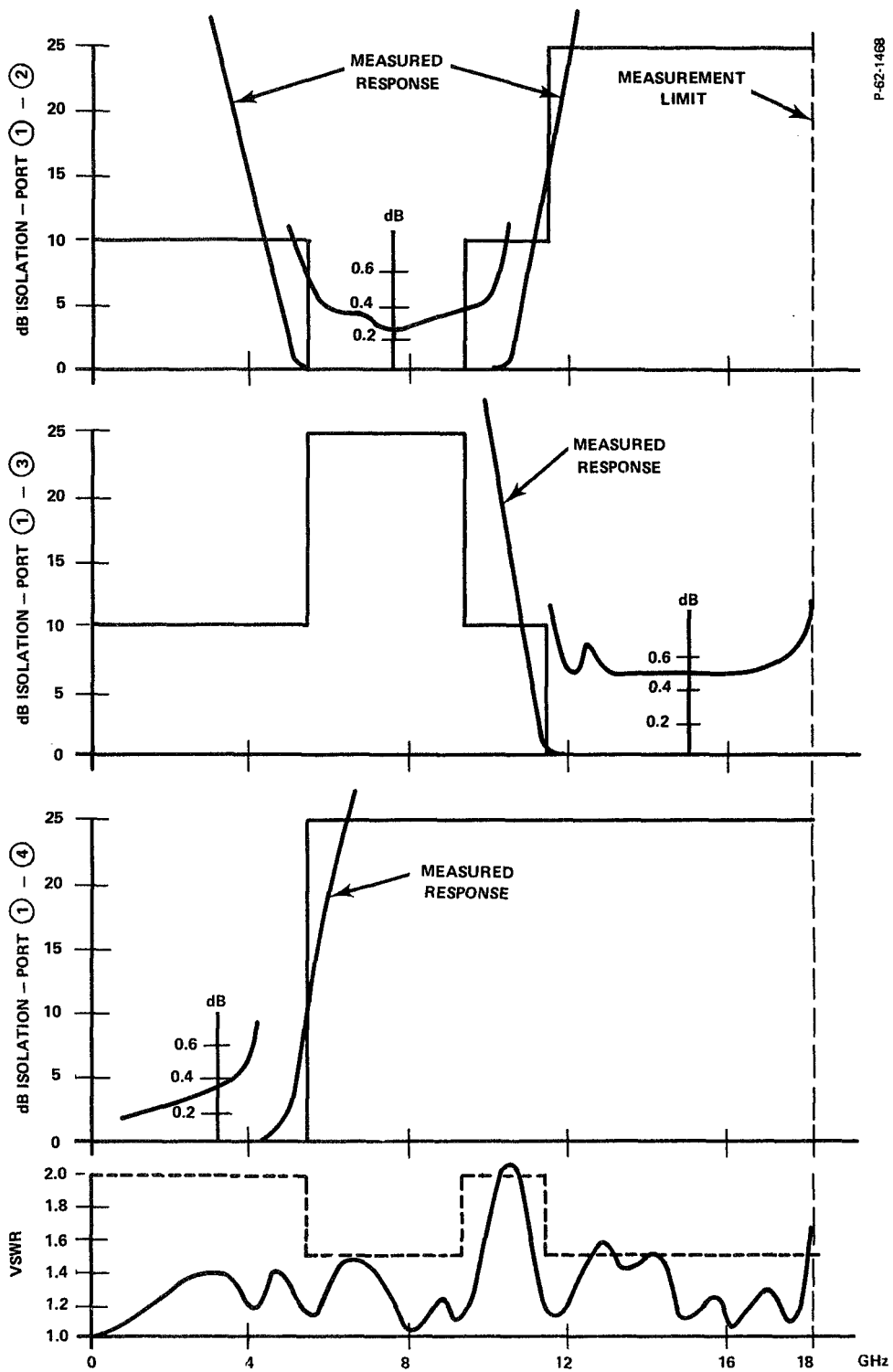


Figure 37 - Measured Performance of Complete Fourth Unit

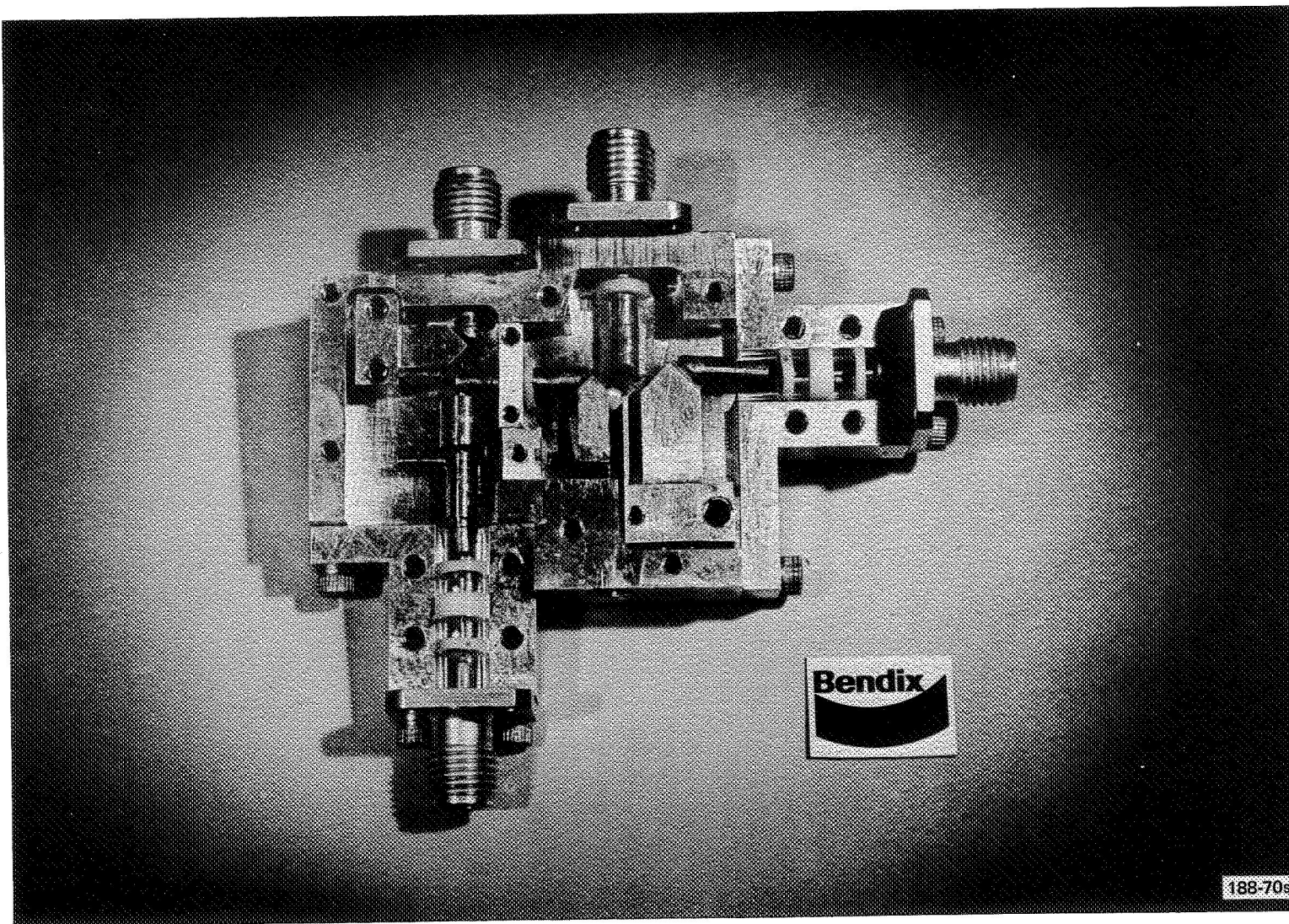
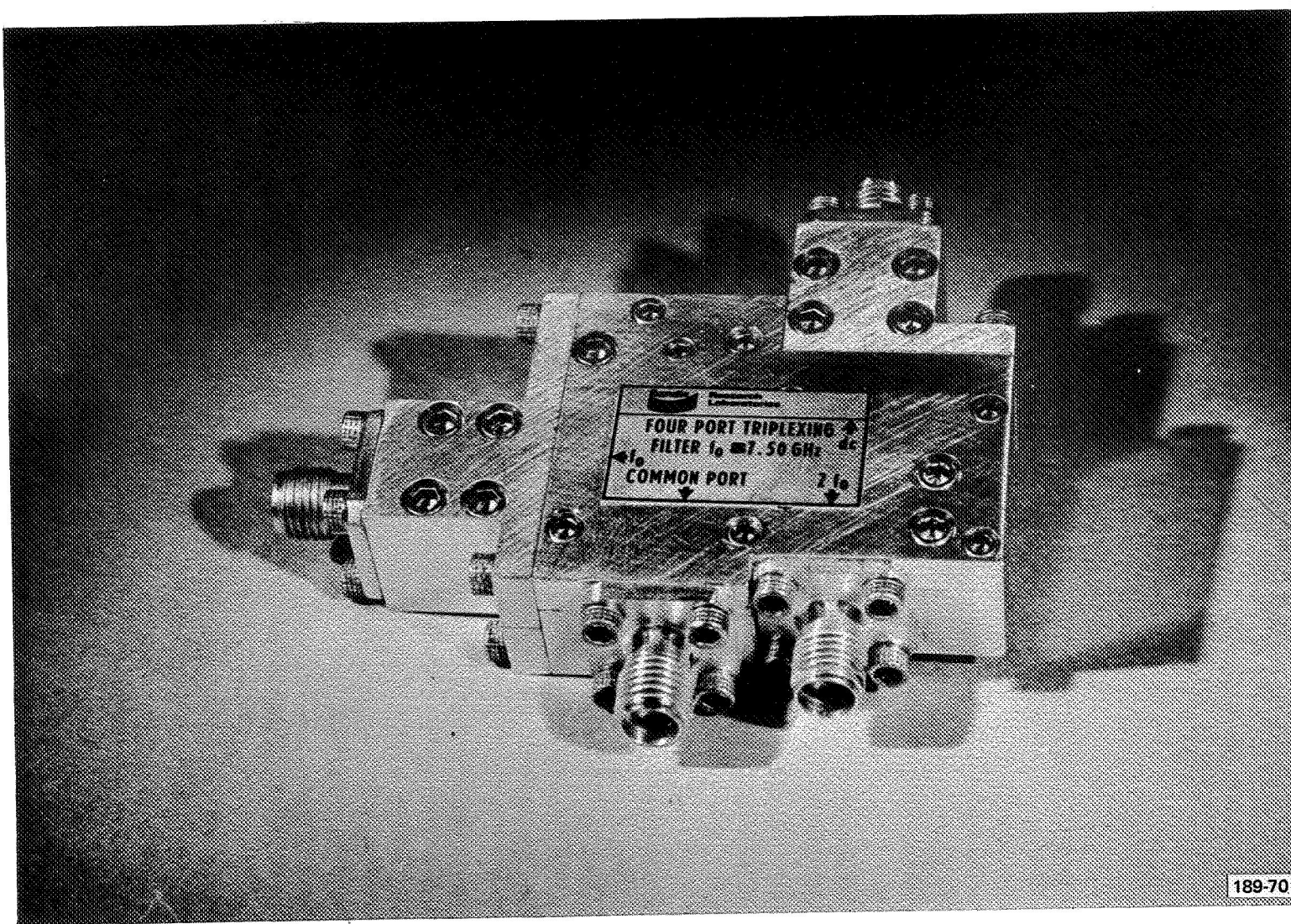


Figure 38 - Interior of Complete Fourth Unit



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Figure 39 - Exterior of Complete Fourth Unit

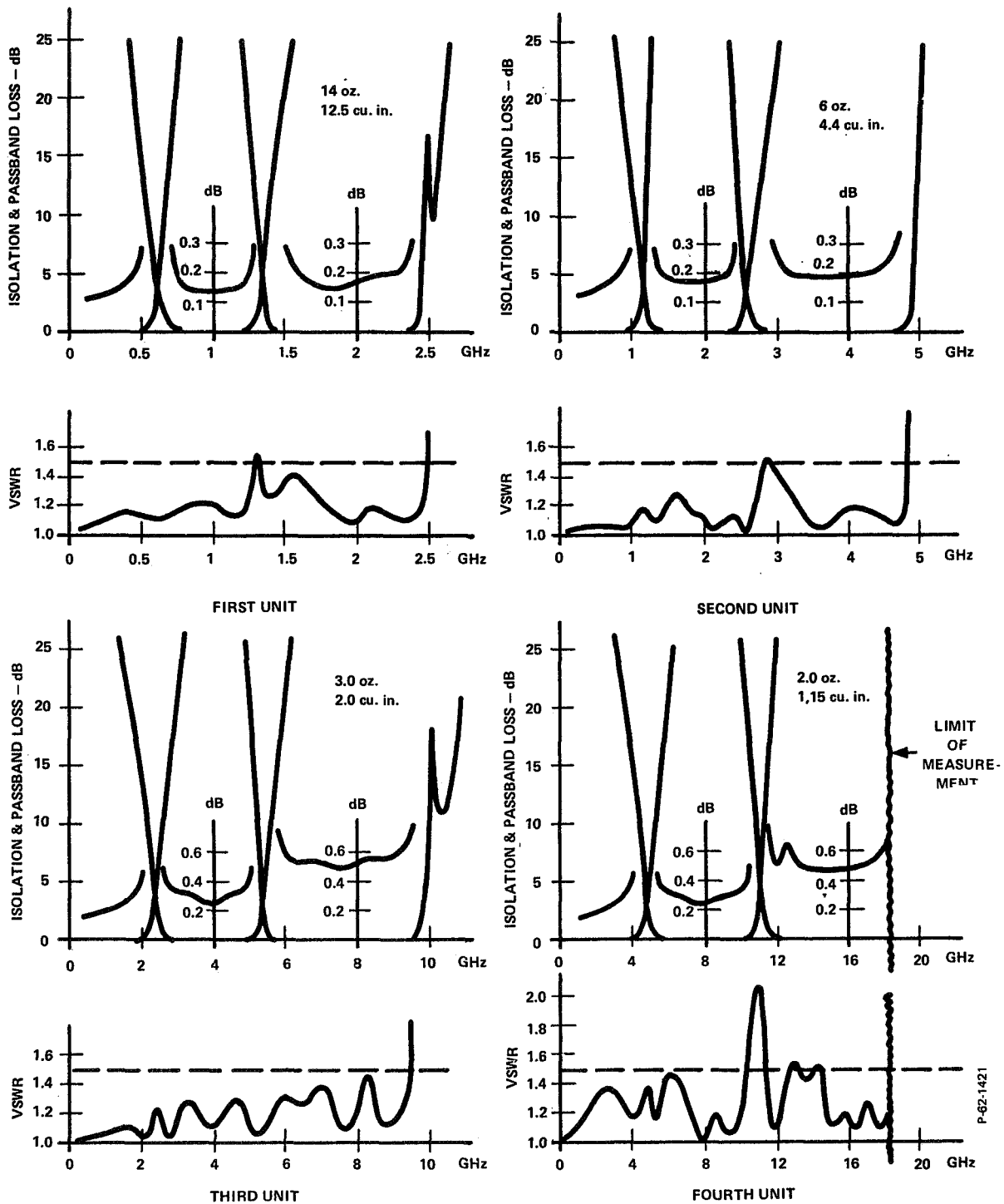


Figure 40 - Measured Performance of Four Multiplexer Units Covering L-Band Through Ku-Band