ing of the rms electric and magnetic field values together with a graphic or magnetic recording is essential.

Using the above-described methods, the spatial and temporal mean of the electric field rather than the mean of the maximum values can be determined. The maximum values, obtained on the integration time of the probe, are nearly the only ones obtainable using the customary wide-band isotropic field meters.

By means of the electric field values thus obtained, the actual whole-body-averaged power absorption rate (SAR) can be estimated.

From the above-reported example, it is seen that noncontinuous monitoring of the field levels would lead to an overestimation of the electric field and of the SAR. We have shown that the SAR overestimation, equal to the ratio of the two different values of the $SAR_{\text{near-field}}$, obtained from the values of $E_{ON}$ and $E_{MAX}$, respectively, can reach a value of 3.5.

On the basis of this work, it seems that technical recommendations or guidelines should consider the measurement instrument characteristics and the methods that must be used to determine compliance with the exposure thresholds contained in the RF radiation safety regulations.

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# Decade Bandwidth Bias T's for MIC Applications up to 50 GHz

B. J. MINNIS

*Abstract* —**A new design of bias T is described capable of carrying direct currents (dc) of a number of amperes while operating over RF bandwidths of more than a decade. Realized in microstrip or stripline, a design can operate up to at least 50 GHz as either a stand-alone component or as part of a microwave integrated circuit (MIC). The bias T circuit has been treated as a combination of HP, BP, and LP filters, the BP filter forming the major part. Design of the filter is by an exact transfer function synthesis procedure involving the application of the Richards transformation [1].**

## I. INTRODUCTION

Bias T's are used for feeding dc power to active RF components in such a way that the RF behavior of the component is not adversely affected by the dc connection. They behave as diplexers which in the ideal case comprise a pair of LP and HP filters with a cutoff frequency just above zero (dc). Large current capacities and broad bandwidth capabilities have, in the past, been difficult to achieve without using elaborate, expensive methods of construction. However, the design described in this paper is capable of handling dc in excess of 3 A, and RF bandwidths of over a decade are possible with operating frequencies reaching as high as 50 GHz. It is suitable for realization in microstrip or triplate stripline and can be constructed as a low-cost individual component or, alternatively, made to form part of a more complex microwave integrated circuit (MIC).

This paper will describe the basic structure of the bias T, its equivalent circuit, and the synthesis procedure used in its design. Two practical examples will be given which cover the bands 1.8–18.2 GHz and 4.5–45.5 GHz, respectively.

## II. BASIC BIAS T AND EQUIVALENT CIRCUIT

The concept of a bias T as a diplexer comprising LP and HP filters is illustrated in Fig. 1(a). It has been common practice for designers of bias T's to concentrate on only the LP element in this combination. Often, the HP element comprises only a series capacitor, while the LP element is a more complicated arrangement of transmission lines, stubs, capacitors, and in some cases resistors. A simple example in widespread use is that illustrated in Fig. 1(b). The LP filter is a transmission line terminated in a shunt capacitor, and the maximum realizable impedance of the line limits the frequency bandwidth. Bandwidths of 3:1 are the typical limit for this bias T.

The aim with the new design was to produce a bias T with decade bandwidth performance at microwave frequencies suitable for printed-circuit realization. This has been achieved by giving attention to the design of the diplexer as a whole, instead of concentrating on the LP filter. Conceptual changes to the filter combination are indicated in the functional diagram of Fig. 2. Most significantly, a BP filter has been added, which, although shown as a pair of filters in the diagram, is a single filter with a

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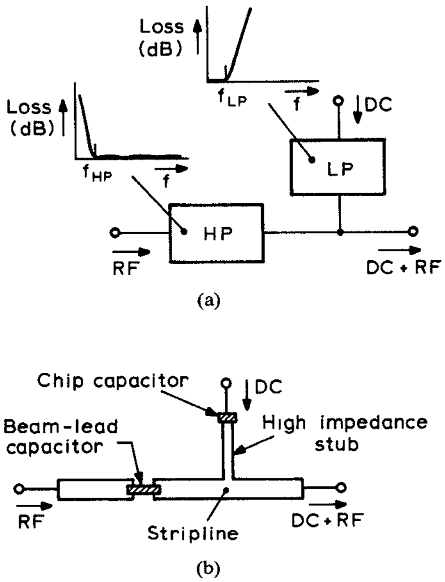



Fig. 1. Simple bias T. (a) Basic concept. (b) Stripline realization.

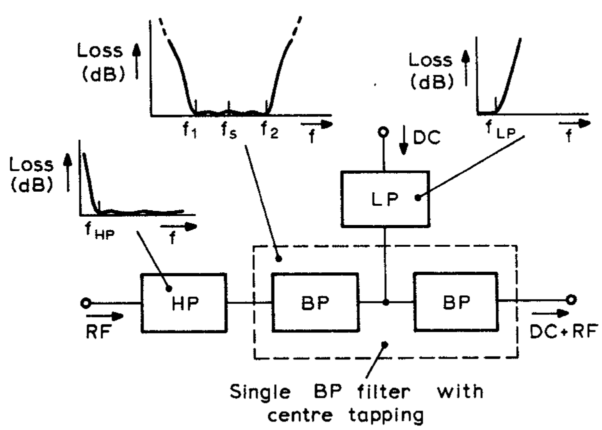



Fig. 2. Functional diagram for new bias T design.

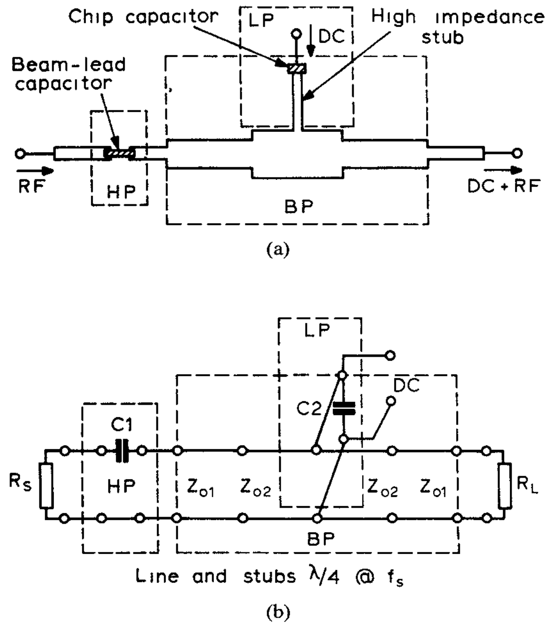



Fig. 3. Typical circuit topology. (a) Stripline realization. (b) Equivalent circuit.

center tapping for the connection of the LP filter. In general, the tapping point can be anywhere along the BP filter, including either of the two ends.

While the position of the shunt stub and the number of transmission line sections can be varied, an outline sketch of a likely stripline circuit is given in Fig. 3(a). Realizable in microstrip or stripline and using the same HP and LP filter arrangements, it bears a resemblance to the simple circuit of Fig. 1(b). However, instead of connecting the stub of the LP filter directly in shunt with a uniform transmission line, the stub forms part of a bandpass filter whose maximum realizable bandwidth is very much greater than that of the simple circuit. Fig. 3(b) gives the electrical equivalent circuit of the bias T in terms of ideal transmission lines and capacitors. Wide bandwidths can be achieved because the cascaded transmission line sections of the BP filter effectively lower the external system impedance in the vicinity of the stub and reduce its shunting effect.

The BP filter is a cascade of transmission lines and a single shunt stub which due to the capacitor at its remote end, behaves as if terminated in a short circuit to RF. All the lines and the stub have an electrical length equal to a quarter wavelength at the center frequency of the passband. Line impedances generally decrease towards the position at which the shunt stub is attached, although the shunt stub itself has an impedance considerably greater than the adjacent line. In most cases, the stub is in the center and line impedances are therefore lowest in the center. A central connection also results in equal source and load impedances. Occasionally, it will be useful to design the bias T with the stub off-center or at one end, in which case it will act as a transformer between unequal source and load impedances.

Details of the synthesis procedure for the BP filter will be given in the following section. In theory, there is no limit to the fractional bandwidth that can be covered by the filter. The consequence of specifying a large bandwidth is to produce a high stub impedance, but the stub impedance can usually be reduced to a realizable value (e.g., 100 Ω) by increasing the number of filter elements. The optimum number of elements is the smallest number which makes the stub readily realizable for the chosen frequency bandwidth. In practice, fractional bandwidths of approximately 20:1 represent the limit, with decade bandwidth (10:1) easily achieved. Two features of the filter are responsible for the limit. They are (1) the presence of the capacitor at the end of the shunt stub which fails to be a good RF short at extremely low frequencies and (2) the aspect ratio of the low-impedance lines, which can become unacceptable at extremely high frequencies. Increased bandwidth lowers the impedance of the innermost lines, and an increased center frequency reduces their length. It should be noted that the introduction of redundant unit elements and the application of admittance matrix transformations cannot improve the aspect ratio problem for very wide bandwidths.

## III. Network Synthesis

The $f$-plane equivalent circuit of the BP filter is a cascade of transmission lines and a single, shunt short-circuit stub. All these elements are a quarter wavelength at the commensurate frequency $fs$, which is the center frequency of the passband. Passband edges ($f1$ and $f2$) will be specified but the complete frequency response of the circuit is a periodic repetition of passbands and stopbands which extend to infinity. Exact synthesis of commensurate line filters of this kind has been made possible by the use of the Richards transformation [1]. This transformation is capable of mapping the periodic frequency response of a commensurate line filter in the $f$ plane into a single, nonperiodic

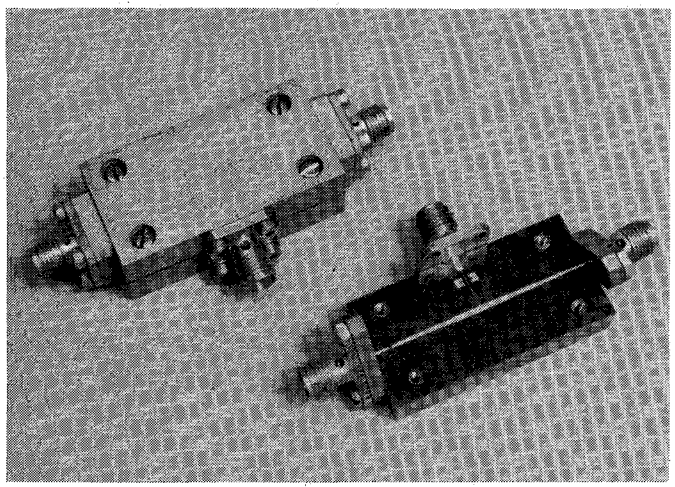


Fig. 4. 2–18-GHz bias T in stripline.

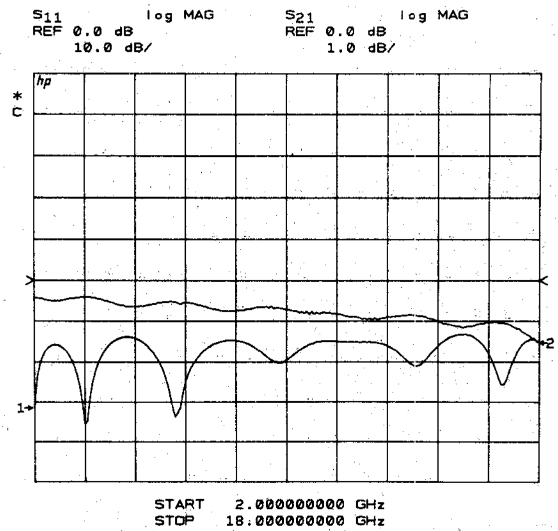



Fig. 5. Measured reflection and transmission parameters, 2–18-GHz bias T.

response on the imaginary axis of the complex frequency plane $S$. The Richards transformation can be written as $S = j\tan(\pi/2 \times f/fs)$, where $S = \sigma + j\omega$. In the case of the BP filter of the bias T, the effect of the mapping is to produce a high-pass frequency response on the $j\omega$ axis of the $S$ plane. For the equivalent circuit, the effect is to translate the shunt stub into an inductor and the transmission lines into unit elements.

Synthesis of the resultant HP filter in the $S$ plane requires the help of a suitable computer program (e.g., commercial package FILSYN). A description of the mathematical details of the synthesis is beyond the scope of this paper, but the subject is well understood and well reported in the literature [2], [3]. The essential information for the synthesis consists of the cutoff frequency $\omega 1$, which is derived from the bandwidth specification by applying the Richards transform, the number of unit elements, which together with the shunt inductor define the degree of the filter, and finally the position of the shunt inductor. Equal terminating impedances will require the inductor to be in a central position, and in this case the specified degree of the filter must be odd. The optimum number of elements will be found by iteration.

## IV. A 2–18-GHz Bias T in Stripline

As a first demonstration of the capability of the design, a bias T covering the 2–18-GHz band was constructed. To allow for manufacturing tolerances, the bandpass filter was synthesized with a passband extending over 1.8–18.2 GHz, which is a bandwidth of more than a decade. In terms of fractional bandwidth (i.e., $2(f2-f1)/(f2+f1)$), this is 164 percent. The passband has a Chebyshev equiripple response with an insertion loss ripple magnitude of 0.1 dB and corresponding ripple in return loss is 16.43 dB. Six transmission line elements, three on each side of the central stub, were required to reduce the stub impedance to around 100 Ω.

A picture showing a pair of the bias T's is given in Fig. 4. Apart for a minor difference in the type of coaxial connectors used and the fact that one of the bias T's has been temporarily split in half to reveal the circuit pattern, they are identical stripline structures. RF connections are made at the two ends of each unit and the dc feed is via the side coaxial connector. Circuit patterns were defined by conventional photolithographic techniques and etched on copper-clad dielectric material.

Three beam-lead capacitors have been mounted on the circuit. One of the capacitors is mounted across a gap in the track close to one end of the bias T, acting as the dc break (i.e., the HP filter). The other two capacitors are wired in parallel and act as the RF short circuit on the end of the shunt stub (i.e., part of the LP filter). Each of these two capacitors is connected to the stub and to an adjacent copper pad earthed with a gold foil passing through the circuit board. Clearance holes have been provided at appropriate positions in the upper dielectric board for all three capacitors.

Measurements of RF insertion loss and return loss have been made with an HP 8510 network analyzer. Results are presented in graphical form in Fig. 5. Over the band 2–18 GHz, peaks in return loss are typically 15 dB, with a single peak of 13.5 dB at 15.5 GHz. Insertion loss rises gently from 0.4 dB at 2 GHz to 1.4 dB at 18 GHz. Both responses are close to those predicted from theory when the effects of circuit losses are taken into account. The slightly high peaks in return loss and distortion in the ripple response are due to interactions between the two coaxial-to-stripline transitions. Currents in excess of 3 A have been passed through the bias T, and RF isolation at the dc connection port was found to be approximately 30 dB in the worst case.

## V. A 4.5–45.5-GHz Bias T in Microstrip

In a second application of the new design procedure, a bias T was produced covering the band 4.5–45.5 GHz. This was a scaled version of the 1.8–18.2-GHz bias T, the center frequency being a round number of 25 GHz. No second synthesis was therefore needed since the equivalent circuit is identical to that of the 1.8–18.2-GHz design except for the difference in the transmission line lengths.

A photograph of an experimental version of the 4.5–45-GHz bias T is given in Fig. 6. The figure illustrates the microstrip circuit, the use of spark-plug-type coaxial-to-microstrip transitions for the RF connections, and the provision of a large copper pad for the dc connection. It was necessary to use a thin (0.125 mm) dielectric substrate to maintain satisfactory aspect ratios for the transmission line elements of this high-frequency circuit. The circuit of Fig. 6 incorporates a pair of MIM (metal–insulator–metal) capacitors flip-chip mounted for the shunt connection at the end of the stub. A series capacitor for the dc break was added after the photograph had been taken.

RF performance of the bias T has been measured in two parts. Firstly, insertion and return losses were measured with an HP 8510 network analyzer from 1 GHz to 26 GHz, 26.5 GHz being

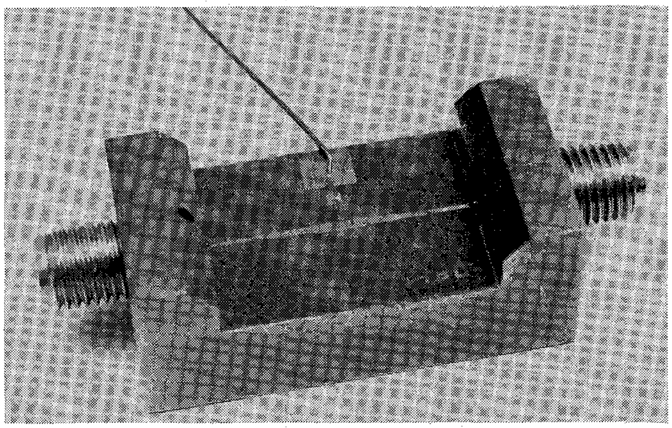
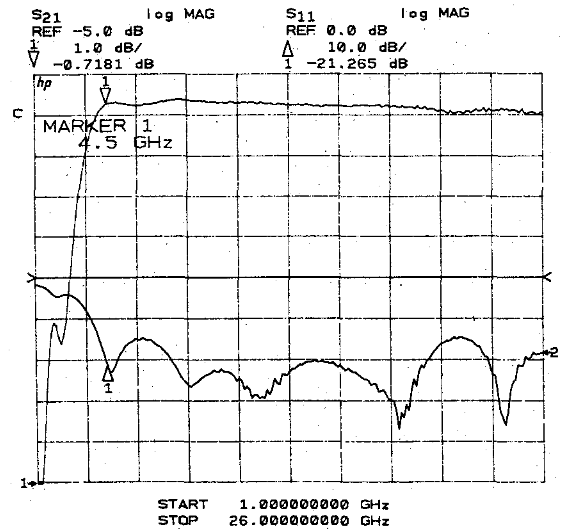


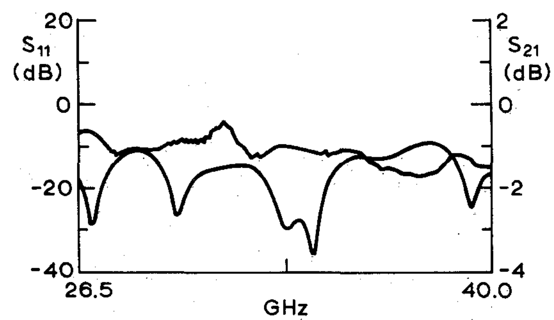
Fig. 6. 4.5–45.5-GHz bias T in microstrip.

the upper limit of the system. Secondly, the same parameters were measured over 26.5 to 40 GHz using a scalar network analyzer. Results are presented in Fig. 7(a) and (b), respectively. Up to 26 GHz, insertion loss in less than 1 dB, and peaks in return loss are better than 15 dB. From the position of the marker in Fig. 7(a), the passband lower edge can be seen to be positioned correctly at 4.5 GHz. Above 26 GHz, insertion loss is typically 1 dB, rising to a peak of 1.5 dB at 37.5 GHz. Return loss is better than 10 dB, the worst peak of 10 dB occurring at the same 37.5-GHz frequency as the peak in insertion loss. Connectors and a detector return loss of only 12 dB at 37 GHz may be responsible for the peak observed in these results. While equipment was not readily available to measure performance beyond 40 GHz, there is no indication that there will be any significant deterioration in performance up to 45 GHz.

The current-carrying capacity of the bias T is greater than 2 A.




(a)




(b)

Fig. 7. Measured reflection and transmission parameters, 4.5–45.5-GHz bias T. (a) 1–26-GHz frequency span. (b) 26.5–40-GHz frequency span.

## VI. Conclusions

The bias T design described is simple, inexpensive, and yet versatile. With the help of two practical examples in microstrip and stripline, decade bandwidths, operation into mm-wave frequencies, and large dc capacities have been demonstrated. The advance over earlier designs has been made as a result of treating the major part of the structure as a BP filter which is synthesized by exact transfer function synthesis procedures.

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